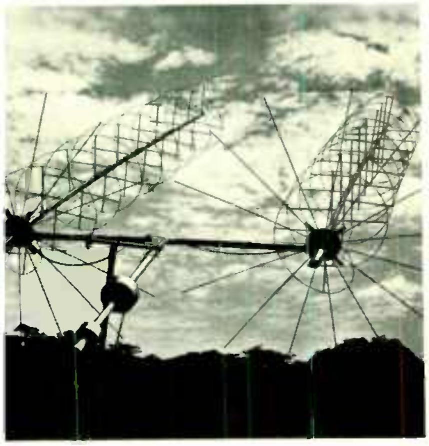
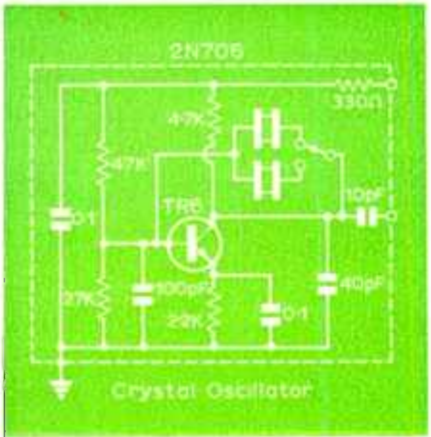
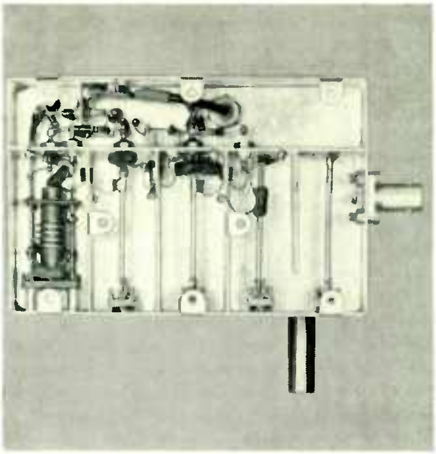
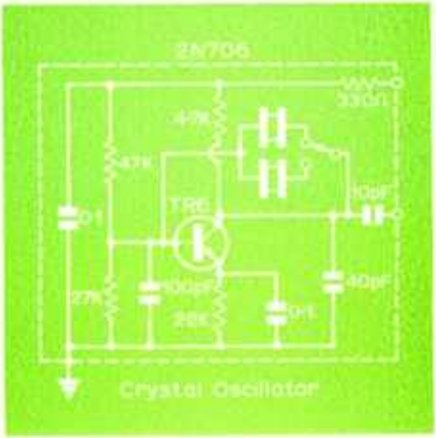
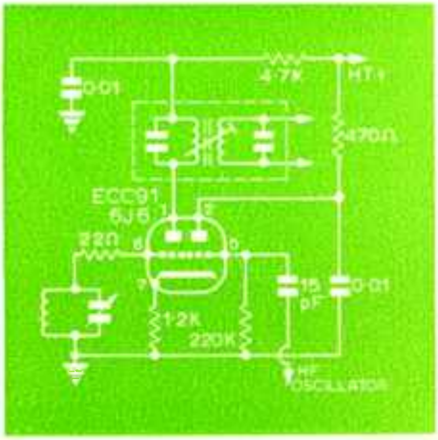
RADIO COMMUNICATION HANDBOOK

THE RADIO SOCIETY OF GREAT BRITAIN





THE RADIO COMMUNICATION HANDBOOK

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THE RADIO COMMUNICATION HANDBOOK

FOURTH EDITION

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This book is dedicated to the memory of John A. Rouse, G2AHL who was General Manager of the RSGB and Editor of the Society's publications until his death in May 1967.

He was responsible for much of the preparatory work in connection with this volume.

RADIO COMMUNICATION HANDBOOK

FOURTH EDITION

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FOREWORD

By Dr. J. A. SAXTON, D.Sc., Ph.D., A.R.C.S., F.I.E.E.

Director, Radio and Space Research Station

AMATEUR interest in radio, which many would claim to be the most interesting of all scientific hobbies, shows no sign of abating while new scientific and technological developments in the field continue apace. It is therefore not surprising that the Society has now found it desirable to bring out a fourth edition of the *Handbook* to succeed the third edition of 1961. At the time the third edition was written the seeds of a number of important advances had already been sown: thus the potentialities of space communications were becoming apparent, as was also the impact of new techniques in design, including such things as miniaturization, printed circuits and advanced solid state devices. The intervening years have seen many of these developments come to fruition, offering in the process new possibilities to amateurs as well as to professional communications engineers. Thus there have already been amateur communication satellites, though none yet launched by the United Kingdom, and there will doubtless be more, and amateurs the world over have exploited the opportunity of using them: moon-bounce experiments have also taken place and there is every incentive to press forward with the application of the latest ground-station techniques. Quite apart from all this, worthwhile refinements and improvements of previously existing radio technology have occurred over the same period and in the new edition of the handbook the Society has aimed to bring readers up to date all round.

The fourth edition therefore represents a very considerable advance on the preceding one and it is more than fifty per cent bigger. There is a new chapter on RTTY and most of the others have been extensively re-written and extended. For example, single sideband techniques are becoming of wider interest to amateurs and in the new volume the treatment of this subject has been greatly developed. There is much new material on aerials, transmitters and receivers, particularly for the h.f. band; and u.h.f. techniques, where amateur activity is also increasing, are comprehensively covered. Rather more is said than previously about wave propagation, a most important matter for a good understanding of propagation phenomena is basic to a full appreciation of the possibilities (and vagaries) of radio communication. Over the entire frequency range now available to amateurs this now demands some knowledge of upper atmospheric physics, sun-earth relationships and, in the troposphere, of radio meteorology. The reader may not find these matters covered in great depth in this volume, and would not expect to do so, but he will find enough to bring out many of the salient features and perhaps to whet his appetite for further information to be sought in more advanced treatises.

The handbook is not only invaluable to the amateur; it is also an extremely useful part of the professional library, being particularly suitable as an introduction to the many varied aspects of radio communication, and the young aspiring communications engineer will find in this book much that is essential to provide a sound basis for his subject. There is no doubt that, in the tradition of its predecessors, to which this volume is a worthy successor, the handbook is eminently suitable to find a place on every amateur's bookshelf and that it will be a great support to him in the pursuit of his hobby.

6th May, 1968 J. A. SAXTON

CHAPTER 1

PRINCIPLES

A KNOWLEDGE of the fundamental physical principles underlying radio communication is just as important to the amateur as to the professional radio engineer. It forms the starting point of all design work and is essential for proper maintenance and successful operating.

In the limited space available here it is not possible to discuss these basic principles in great detail, but a number of references are included in the bibliography at the end of the chapter as a guide to those readers who would like to make a further study of any particular part of the subject. Most public libraries will be only too glad to obtain the books listed if they are not already on their shelves.

The pattern of headings follows the syllabus of the Radio Amateurs' Examination very closely, and it is hoped that newcomers will find this chapter a useful introduction to Amateur Radio. More experienced readers may find it useful as a refresher course.

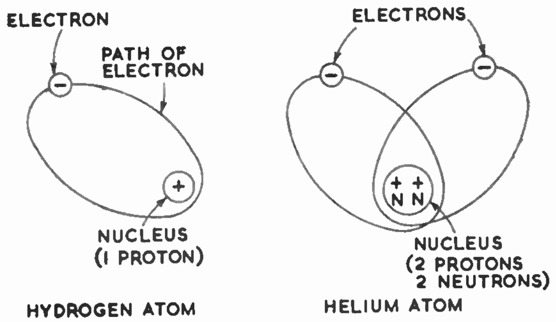


Fig. 1.1. Structure of hydrogen and helium atoms.

ATOMS AND ELECTRONS

All matter is composed of molecules, which is the name given to the smallest quantity of a substance which can exist and still display the physical and chemical properties of that substance. There is a very great number of different sorts of molecules. Further study of a molecule discloses that it is made up of smaller particles called atoms and it has been found that there are about 102 different types of atoms. All molecules are made up of combinations of atoms selected from this range. Examples of different atoms are atoms of hydrogen, oxygen, iron, copper and sulphur, and examples of how atoms are combined to form molecules are (*a*) two atoms of hydrogen and one of oxygen to form one molecule of water, and (*b*) two atoms of hydrogen, one of sulphur and four of oxygen to form one molecule of sulphuric acid.

Atoms are so small that they cannot be seen even under the most powerful microscopes. Their behaviour, however, can be studied and from this it has been discovered that atoms are made up of a positively charged relatively heavy core or *nucleus* around which are moving a number of much lighter particles each negatively charged, called *electrons*. Atoms are normally electrically neutral; that is to say, the amount of positive electricity associated with the nucleus is exactly balanced by the total amount of negative electricity associated with the electrons. One type of atom differs from another in the number of positive and neutral particles (called *protons* and *neutrons*) which make up the nucleus and the number and arrangement of the orbital electrons which are continually moving around the nucleus.

Large atoms, such as those of uranium, are complex, but small atoms, such as those of hydrogen and helium, are relatively simple, as can be seen in **Fig. 1.1.**

The important fact which emerges from the preceding paragraphs is that all matter is made up of positively and negatively charged particles, which means that electricity is latent in everything around us.

CONDUCTORS AND INSULATORS

The ease with which the electrons in a substance can be detached from their parent atoms varies from substance to substance. In some substances there is a continual movement of electrons in a random manner from one atom to another, and the application of an electrical pressure or voltage (for example from a battery) to the two ends of a piece of wire made of such a substance will cause a drift of electrons along the wire called an *electric current*; electrical conduction is then said to take place. It should be noted that if an electron enters the wire from the battery at one end it will be a different electron which immediately leaves the other end of the wire. By arbitrary convention, the direction of current flow is said to be from positive to negative.

Materials which exhibit this property of electrical conduction are called *conductors*. All metals belong to this class. Materials which do not conduct electricity are called *insulators*, and the table shown above gives a few examples of commonly used conductors and insulators.

Conductors	Insulators
Silver	*Mica*
Copper	*Quartz*
Aluminium	*Glass*
Brass	*Ceramics*
Steel	*Ebonite*
Mercury	*Plastics*
Carbon	*Air*
Certain liquids	*Oil*

SOURCES OF ELECTRICITY

When two dissimilar metals are immersed in certain chemical solutions, or *electrolytes*, an electromotive force (e.m.f.) is created by chemical action within the cell so that if the pieces of metal are joined externally to the cell there will be a continuous flow of electric current. Such a device is called a *simple cell* and such a cell comprising copper and zinc rods immersed in diluted sulphuric acid is shown in Fig. 1.2 (A). The flow of current is from the copper to the zinc plate in the external circuit; i.e. the copper forms the positive (+) terminal of the cell and the zinc forms the negative (−) terminal.

In a simple cell of this type hydrogen forms on the copper electrode, and this gas film has the effect of increasing the internal resistance of the cell and also setting up within the cell a counter or polarizing e.m.f. which rapidly reduces the effective e.m.f. of the cell as a whole. This polarization effect is overcome in practical cells by the introduction of chemical agents surrounding the anode for the purpose of removing the hydrogen by oxidation as soon as it is formed. Such agents are called *depolarizers*.

Primary Cells. Practical cells in which electricity is produced in this way by direct chemical action are called *primary cells*; a common example is the Leclanché cell, the construction of which is shown diagrammatically in Fig. 1.2 (B). The zinc case is the negative electrode and a carbon rod is the positive electrode. The black paste surrounding the carbon rod may contain powdered carbon, manganese oxide, zinc chloride, sal-ammoniac and water, the manganese oxide acting as depolarizer by combining with hydrogen formed at the anode to produce another form of manganese oxide and water. The remainder of the cell is filled with a white paste which may contain plaster of Paris, flour, zinc chloride, sal-ammoniac and water. The cell is sealed with pitch except for a small vent which allows accumulated gas to escape. The e.m.f. developed by a single dry cell is about 1·5 volts and cells may be connected in series (positive terminal to negative terminal and so on) until a battery of cells, usually referred to simply as a *battery*, of the desired voltage is obtained. The symbol used to denote a cell in a circuit diagram is shown in Fig. 1.2 (C). The long thin stroke represents the positive terminal and the short thick stroke the negative terminal. Several cells joined in series to form a battery are shown; for higher voltages it becomes impracticable to draw all the individual cells involved and it is sufficient to indicate merely the first and last cells with a dotted line between them with perhaps a note added to state the actual voltage. The amount of current which can be derived from a dry cell depends on its size and the life required and may range from a few milliamperes to an ampere or two.

Secondary Cells. In primary cells some of the various chemicals are used up in producing the electrical energy—a relatively expensive and wasteful process. The maximum current available also is limited. Another type of cell, called a *secondary cell*, *storage cell* or *accumulator*, offers the advantage of being able to provide a higher current and is capable of being charged by feeding electrical energy into the cell to be stored chemically, and be drawn out or discharged later as electrical energy again. This process of charging and discharging the cell is capable of repetition almost indefinitely.

The most common type of secondary cell is the lead-acid cell such as that used in motor-car batteries. It consists of two sets of specially prepared lead plates immersed in a mixture of sulphuric acid and water. Briefly the action of the cell is as follows: in the discharged state the active material on each plate is lead sulphate. During the charging process the lead sulphate in the positive plate is changed to lead peroxide and sulphuric acid, and the lead sulphate in the negative plate to a form of lead called spongy lead and sulphuric acid. During discharge the reverse action takes place and lead sulphate forms again on both plates. The state of the charge of the cell may be checked by measuring the specific gravity of the electrolyte with a hydrometer since the concentration of sulphuric acid increases during charging and decreases again as the cell is discharged. Typical values of specific gravity are 1·250 for a fully charged cell and 1·175 for a discharged cell. The terminal voltage of a single lead-acid cell when fully charged and left standing is about 2·05 volts. In the discharged condition the voltage falls to about 1·85 volts.

To obtain a long life cells should not be overcharged since this causes flaking and buckling of the plates. Cells should not be left in a discharged state because the lead sulphate may undergo a physical change which it is difficult to break down in the re-charging process and the capacity of the cell will be impaired; cells in this condition are said to be "sulphated," a condition which may sometimes be partially eradicated by prolonged charging at a low current.

Mechanical Generators. Mechanical energy may be converted into electrical energy by moving a coil of wire in a magnetic field. Direct-current or alternating-current generators are available in all sizes but the commonest types

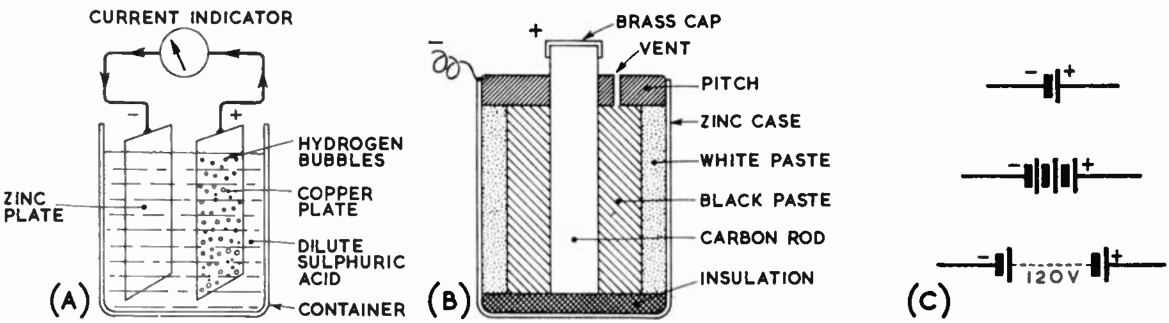


Fig. 1.2. The electric cell. A number of cells connected in series is called a battery. (a) A simple electric cell consisting of copper and zinc electrodes immersed in dilute sulphuric acid. (b) Sectional drawing showing construction of a dry cell. (c) Symbols used to represent single cells and batteries in circuit diagrams.

likely to be met in amateur radio work are (i) a.c. petrol-driven generators of up to 1 or 2 kW output such as are used for supplying portable equipment and (ii) small motor generators, sometimes called *dynamotors*, which furnish up to about 100 watts of power and comprise a combined low-voltage d.c. electric motor and a d.c. generator so that a high-tension supply may be derived from a 6- or 12-volt car battery.

ELECTRICAL UNITS

(a) Unit of Quantity. Since all electrons, whatever kind of atom they belong to, carry the same charge, the amount of electricity associated with the electron could be used as the unit of quantity of electricity. It is, however, too small for use as a practical unit, and a more convenient unit is called the *coulomb*; this is equivalent to the charge on approximately 6×10^{18} (six million million million) electrons. The analogy here between the molecule of water as a unit of quantity of water and the practical unit, the gallon, is obvious.

A quantity of electricity or number of coulombs is usually denoted by the symbol q.

(b) Unit of Current Flow. Continuing with the water analogy, whereas a flow of water is spoken of as "x gallons per second" the flow of electricity can be expressed as "x coulombs per second." A current of one coulomb per second is called an *ampere* (abbreviated to *amp.*) and the strength of a current is said to be "x amperes." A current flow is usually denoted by the symbol I. The currents used in radio are often very small fractions of an ampere and for convenience the two smaller units *milliampere* (meaning a thousandth of an ampere) and *microampere* (meaning a millionth of an ampere) are used. Thus a current of 0·003 ampere is written as 3 milliamperes. See **Table 1.1** for abbreviations.

The relation between the total quantity of electricity (q) which has passed a point in a wire, the time of flow (t) and the rate of flow (I) is therefore—

Quantity (coulombs) = *Current Flow* (amperes) × *Time* (seconds)

or in symbols—

$$q = I \times t.$$

(c) Unit of Electric Pressure. In order to make electricity flow continuously through a circuit it is necessary to have some device which can produce a continuous supply of electrons. This may be a battery in which the supply of electrons is produced by chemical action or a dynamo or generator in which mechanical energy is turned into electrical energy. The battery or generator produces an electrical pressure sometimes called an electromotive force or e.m.f. which may be used to force a current through a circuit. The unit of electrical pressure is the *volt*, and voltages are usually denoted in formulae by the symbol E or V.

(d) Unit of Electrical Resistance. The ease with which an electric current will flow, or can be conducted, through a wire will depend on the dimensions of the wire and the material from which it is made. The opposition of a circuit to the flow of current is called the *resistance* of the circuit and is usually denoted by the symbol R in formulae. The resistance of a circuit is measured in *ohms*, and a circuit is said to have a resistance of 1 ohm if the voltage between its ends is 1 volt when the current flowing is 1 amp. For convenience, because the resistances used in radio equipment may be up to 10 million ohms two larger units called the *kilohm* (meaning 1000 ohms) and the *Megohm* (meaning 1 million ohms) are used. Thus 47,000 ohms may be spoken of as 47 kilohms. See **Table 1.1.** for abbreviations.

TABLE 1.1
Units and Symbols

Quantity	Symbol used in formulæ	Unit	Abbreviation
Quantity	q	Coulomb	–
Current	I	Ampere	A
Voltage	E or V	Volt	V
Time	t	Second	s
Resistance	R	Ohm	Ω
Capacitance	C	Farad	F
Inductance	L	Henry	H
Mutual Inductance	M	Henry	H
Power	W	Watt	W
Frequency	f	Cycle/second	c/s
Frequency	f	Hertz	Hz
(One Hertz equals one cycle per second)			
Wavelength	λ	Metre	m

The following multiples and sub-multiples of the above units are commonly used:

Microampere	(μA)	= 1 millionth of an ampere
Milliampere	(mA)	= 1 thousandth of an ampere
Microvolt	(μV)	= 1 millionth of a volt
Millivolt	(mV)	= 1 thousandth of a volt
Kilovolt	(kV)	= 1,000 volts
Megohm	(MΩ)	= 1,000,000 ohms
Microfarad	(μF)	= 1 millionth of a farad
Micro-microfarad	(μμF)	= 1 million-millionth of a farad
Picofarad	(pF)	= 1 million-millionth of a farad
Microsecond	(μS)	= 1 millionth of a second
Millisecond	(mS)	= 1 thousandth of a second
Milliwatt	(mW)	= 1 thousandth of a watt
Megacycle/second	(Mc/s)	= 1,000,000 cycles per second
Megahertz	(MHz)	= 1,000,000 Hertz
Kilocycle/second	(kc/s)	= 1000 cycles per second
Kilohertz	(kHz)	= 1000 Hertz
Microwatt	(μW)	= 1 millionth of a watt
Kilowatt	(kW)	= 1000 watts
Metre	(m)	= 100 centimetres
Kilometre	(km)	= 1000 metres

RESISTANCE AND CONDUCTANCE

Different materials may be compared as conductors by measuring the resistance of samples of the materials of the same size and shape. The resistance (in ohms) of conducting material of one centimetre in length and one square centimetre in cross-sectional area is called the *specific resistance* or *resistivity* of the material. Specific resistance is quoted as x ohms per centimetre cube or more simply as x ohm-cm. If the specific resistance of a material is known, the actual resistance of, say, a piece of wire made from that material can be calculated since the resistance of a wire increases proportionally with its length and inversely with its cross-sectional area.

If the length of a wire is l cm, its cross-sectional area a sq.

cm, and its specific resistance S ohm-cm, its actual resistance will be given by the formula—

$$R = \frac{S \times l}{a} \text{ ohms}$$

Approximate specific resistances of typical materials used in radio equipment are given in **Table 1.2**. At radio frequencies the resistance of a wire may be very much greater than its direct-current value because radio-frequency currents only travel along the surface shell of the wire and not through the whole body as in the case of direct current. This is called the *skin effect* and is important in making high-quality radio coils.

Sometimes instead of speaking of the resistance of a circuit the *conductance* of the circuit is referred to: this is simply the reciprocal of the resistance. The unit of conductance is the *mho*, and conductance is usually denoted in formulae by the symbol G.

The relation between resistance and conductance is therefore—

$$G = \frac{1}{R} \quad \text{or conversely—} \quad R = \frac{1}{G}$$

Thus a resistance of 10 ohms has a conductance of 0·1 mho.

TABLE 1.2

Specific Resistance of Commonly Used Conductors

Silver	$1{\cdot}6 \times 10^{-6}$ ohm-cm
Copper	$1{\cdot}7 \times 10^{-6}$
Aluminium	3×10^{-6}
Brass	7×10^{-6}
Iron	12×10^{-6}
Special high-resistance alloys:	
Manganin	48×10^{-6}
Eureka	49×10^{-6}
Nichrome	108×10^{-6}

Ohm's Law

Ohm's Law states that the ratio of the voltage applied across a resistance to the current flowing through that resistance is constant. This ratio is equal to the value of the resistance.

Writing this as a formula—

$$\frac{\textit{Voltage}}{\textit{Current}} = \textit{a constant} = \textit{Resistance}$$

or in symbols—

$$\frac{E}{I} = R$$

Note that if E is in volts and I in amperes, then R is in ohms.

It will be seen that this formula relates current, voltage and resistance, so that if two of these quantities are known the third may be calculated by suitably rearranging the formula in the following ways:

$$E = I \times R \text{ (giving } E\text{, knowing } I \text{ and } R) \quad \ldots \text{(i)}$$

$$I = \frac{E}{R} \text{ (giving } I\text{, knowing } E \text{ and } R) \quad \ldots \text{(ii)}$$

$$R = \frac{E}{I} \text{ (giving } R\text{, knowing } E \text{ and } I) \quad \ldots \text{(iii)}$$

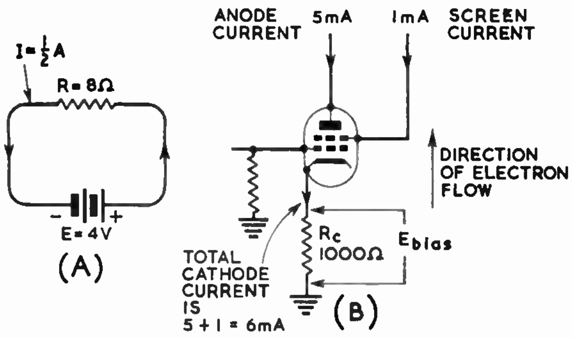


Fig. 1.3. Applications of Ohm's Law.

The application of Ohm's Law is illustrated by the following examples.

Example I. *Consider the circuit shown in* **Fig. 1.3 (A)** *which consists of a battery E of voltage 4 volts and a resistance R of 8 ohms. What is the magnitude of the current in the circuit?*

Here $E = 4$ volts and $R = 8$ ohms. Let I be the current flowing in amperes. Then from Ohm's Law (Formula ii)—

$$I = \frac{E}{R} \text{ amperes} = \frac{4}{8} \text{ amperes} = \frac{1}{2} \text{ ampere}$$

It should be noted that in all calculations based on Ohm's Law care must be taken to ensure that I, E and R are in amperes, volts and ohms respectively if errors in the result are to be avoided.

Example II (from a *Radio Amateurs' Examination* paper). *What is the standing bias voltage produced by a cathode bias resistor of 1000 ohms when the characteristics of the valve are such that the anode current is 5 mA and the screen-grid current is 1 mA?*

The circuit is shown in **Fig. 1.3 (B)** in which the total cathode current is the sum of the anode and screen-grid currents. The cathode current flowing through the cathode resistor will cause a voltage to appear across its terminals which is the required standing bias voltage.

Let E_{bias} = standing bias voltage in volts,
I_c = total cathode current in amperes,
R_c = resistance of the cathode resistor in ohms.

Then from Ohm's Law—

$$E_{bias} = (I_c \times R_c) \text{ volts}$$

Here $I_c = (5 + 1) = 6$ milliamperes $= 0{\cdot}006$ ampere.
$R_c = 1000$ ohms.

Substituting in the formula—

$$E_{bias} = 0{\cdot}006 \times 1000 = 6 \text{ volts}$$

Electrical Power

When a current of electricity flows through a resistance e.g. in an electric fire, the resistance gets hot and electrical energy is turned into heat. The actual rise in temperature depends on the amount of power dissipated in the resistance and the shape and size of the resistance. Sometimes the power dissipated is so small that the temperature rise is not very noticeable, but nevertheless whenever an electric current flows through a resistance power is dissipated therein. The unit of electrical power is the *watt*, usually denoted in formulae by the symbol W. The amount of electrical power dissipated in a resistance is equal to the product of the voltage across the resistance and the current flowing through the resistance. Thus—

Power (watts) = *Voltage* (volts) × *Current* (amperes)

or in symbols—

$$W = E \times I \qquad \ldots \text{(i)}$$

Since from Ohm's law—

$$E = I \times R \text{ and } I = E/R$$

the formula for the power dissipated in a resistance may also be written in two further forms:

$$W = E^2/R \qquad \ldots \text{(ii)}$$

$$W = I^2 \times R \qquad \ldots \text{(iii)}$$

These formulae are useful for finding, for example, the power input to a transmitter or the power dissipated in various resistances in an amplifier so that suitably rated resistances can be selected. To take a practical case, consider again the circuit of Fig. 1.3 (A). The power dissipated in the resistance may be calculated as follows:

Here $E = 4$ volts and $R = 8$ ohms. Let W be the power in watts. Then—

$$W = \frac{E^2}{R} = \frac{4 \times 4}{8} = 2 \text{ watts.}$$

It must be stressed again that the beginner should always see that all values are expressed in terms of volts, amperes and ohms in this type of calculation. Careless use of Megohms or milliamperes, for example, may lead to an answer several orders too large or too small.

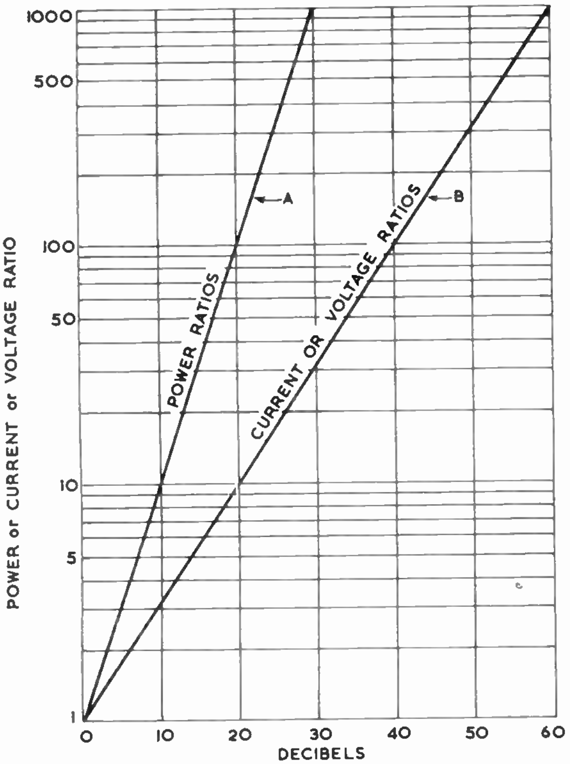


Fig. 1.4. Graph relating decibels to power or current or voltage ratio.

Decibels and Nepers

In radio and line transmission engineering power ratios are often expressed in terms of units proportional to the logarithm of their ratio. In this way gains and losses in amplifiers, networks and transmission paths can be added and subtracted instead of having to be multiplied together.

Two such units are in common use. The first is the decibel, abbreviated *db*, based on common logarithms and widely used in the British Commonwealth and in the USA. The second is the neper, based on natural logarithms and widely used in continental Europe.

TABLE 1.3

Decibels	1	2	3	6	10	14	20	40
Power Ratios	1·26	1·58	2	4	10	25	100	10^4
E or I Ratios	1·12	1·26	$\sqrt{2}$	2	3·16	5	10	100

If there are two power levels, P_1 and P_2, for example at the input and output of an amplifier, the power ratio in decibels, N, is given by the expression

$$N = 10 \log_{10} \frac{P_1}{P_2} \text{ db}$$

Expressed in nepers

$$N = \tfrac{1}{2} \log_e \frac{P_1}{P_2} \text{ nepers}$$

Nepers can be converted easily into decibels, and vice versa, since from the two equations above it follows that:

1 neper = 8·69 decibels

Curve A in Fig. 1.4 can be used for determining the number of decibels corresponding to a given power ratio and vice-versa.

Since power can be expressed in terms of current in, or voltage across a resistance, it follows that, for a given value of resistance, voltage and current ratio can also be expressed in decibels (or nepers).

If $\quad P_1 = I_1^2 R = \dfrac{E_1^2}{R} \quad$ and $\quad P_2 = I_1^2 R = \dfrac{E_2^2}{R}$

then for a current ratio of I_1/I_2

$$N = 20 \log_{10} \frac{I_1}{I_2} \text{ db}$$

and for a voltage ratio of E_1/E_2

$$N = 20 \log_{10} \frac{E_1}{E_2} \text{ db}$$

Fig. 1.4, curve B, shows the relationship between decibels and voltage and current ratios.

Table 1.3 gives some useful approximate power, voltage and current ratios in terms of decibels.

It should be noted that *strictly speaking* current and voltage ratios should only be expressed in decibels when the resistance at which they are measured is the same. In practice, however, the voltage gains of amplifiers having very different input and output resistances are often quoted in this way; it is an arbitrary use of the decibel which although sometimes convenient in practice often leads to confusion.

TABLE 1.4

The Resistor Colour Code

0 black	5 green	Tolerance colours:
1 brown	6 blue	± 5 per cent .. gold
2 red	7 violet	±10 per cent .. silver
3 orange	8 grey	±20 per cent .. no colour
4 yellow	9 white	

Resistors Used in Radio Equipment

As already mentioned, a resistor through which a current is flowing may get hot. It follows therefore that in a piece of radio equipment the resistors of various types and sizes that are needed must be capable of dissipating the power as required without overheating.

Generally speaking, radio resistances (or *resistors* as they are usually called) can be divided roughly into two classes, (*a*) low power up to 3 watts and (*b*) above 3 watts. The low-power resistors are usually made of carbon and may be obtained in a wide range of resistance values from about 10 ohms to 10 Megohms and in power ratings of 1/10 watt to 3 watts. Typical carbon resistors are shown below in **Fig. 1.5**. For higher powers, resistors are usually wire-wound on ceramic formers and the very fine wire is protected by a vitreous enamel coating. Typical wire-wound resistors are shown on the right.

Resistors, particularly the small carbon types, are usually colour-coded to indicate the value of the resistance in ohms and sometimes also the tolerance or accuracy of the resistance. The standard colour code is shown in **Table 1.4**.

The colours are applied either as *bands* at one end of the resistor as shown in **Fig. 1.6 (A)**, or as *body*, *tip* and *dot* colours as shown in **Fig. 1.6 (B)**. As an example, what would be the value of a resistor with the following markings?

(A) Colour Bands—*yellow*	(B) Body—*blue*
violet	Tip—*grey*
orange	Dot—*red*
silver	(*No tolerance colour on other tip.*)

Type A. The yellow first band signifies that the first figure is 4, the violet second band signifies that the second figure is 7, while the orange third band signifies that there are three zeros to follow: the silver fourth band indicates a tolerance of ±10 per cent. The value of the resistor is therefore 47,000 ohms ±10 per cent (47 K ohms ±10 per cent).

Type B. The blue body gives the first figure as 6, the grey tip gives the second figure as 8, while the red dot signifies that there are two zeros to follow: in the absence of any colour on the other tip the tolerance must be taken as ±20 per cent. The value of the resistor is therefore 6,800 ohms ±20 per cent (6·8 K ohms ±20 per cent).

So far only fixed resistors have been mentioned. Variable resistors, sometimes called *potentiometers* or *volume controls*, are also used. These are usually panel-mounted by means of a threaded bush through which a ¼-in. diameter spindle protrudes and to which the control knob is fitted. Low-power high-value variable resistances use a carbon resistance element and high-power lower resistance types (up to 100,000 ohms) use a wire-wound element.

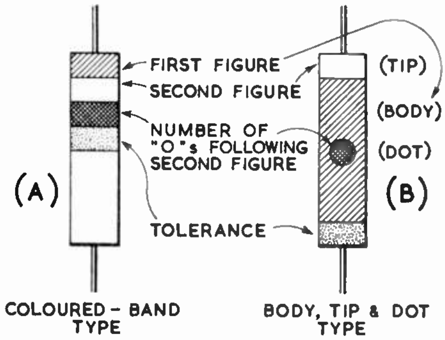


Fig. 1.6. Standard resistance value markings.

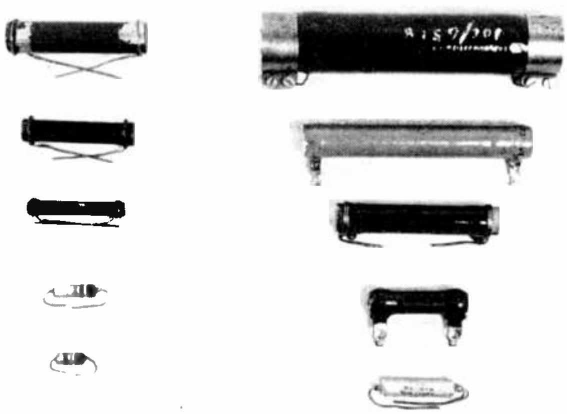


Fig. 1.5. Typical resistors used in radio equipment. The specimens shown on the left are carbon resistors and those on the right are wire-wound.

Resistances in Series and Parallel

Resistors may be joined in series or parallel to obtain some special desired value of resistance. When in series, resistors are connected as shown in **Fig. 1.7 (A)** and the total resistance of the resistors is equal to the sum of the separate resistances. The parallel connection is shown in **Fig. 1.7 (B)**, and with this arrangement the reciprocal of the total resistance is equal to the sum of the reciprocals of the separate resistances.

If R is the total resistance, these formulae can be written as follows:

Series Connection	Parallel Connection
$R = R_1 + R_2 + R_3 + \text{etc.}$	$\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \text{etc.}$

Considering the case of only *two* resistors in parallel—

$$\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} = \frac{R_1 + R_2}{R_1 R_2}$$

which by inversion gives—

$$R = \frac{R_1 R_2}{R_1 + R_2}$$

This is a useful formula, since with it the value of two

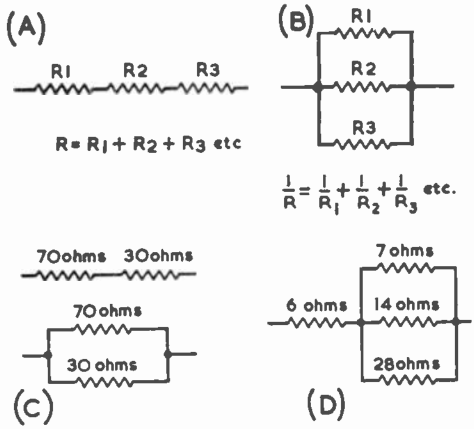


Fig. 1.7. Resistors in various combinations: (A) series, (B) parallel, (C) series and parallel (D) series-parallel. The calculation of the resultant resistances in (C) and (D) is explained in the text.

resistors in parallel can easily be calculated without the use of reciprocals.

To illustrate these rules two examples are given below.

Example I. *Calculate the resistance of a 30 ohm and a 70 ohm resistor connected first in series and then in parallel.*

In series connection—

$$R = 30 + 70 = 100 \text{ ohms}$$

In parallel connection, using the special formula for two resistors in parallel—

$$R = \frac{R_1 R_2}{R_1 + R_2}$$

$$= \frac{30 \times 70}{30 + 70} = 21 \text{ ohms}$$

These two calculations are illustrated in **Fig. 1.7** (C).

Example II. *Three resistors of 7, 14 and 28 ohms are connected in parallel. If another resistor of 6 ohms is connected in series with this combination what is the total resistance of the circuit?*

The circuit is shown in **Fig. 1.7** (D). Taking the three resistors in parallel first, these are equivalent to a single resistance of R ohms given by—

$$\frac{1}{R} = \frac{1}{7} + \frac{1}{14} + \frac{1}{28} = \frac{4 + 2 + 1}{28} = \frac{1}{4}$$

Therefore—

$$R = 4 \text{ ohms}$$

This in series with the 6 ohm resistor gives a total resistance of 10 ohms for the whole circuit.

CAPACITORS AND CAPACITANCE

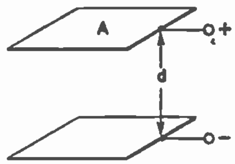
Capacitors (or condensers) have the property of being able to store a charge of electricity. They consist essentially of two conducting plates or strips separated by an insulating medium called a *dielectric*. When a capacitor is charged there is a voltage difference between its plates, and the larger the plates and/or the smaller their separation the greater will be the charge that the capacitor holds for any given voltage across its plates.

The unit of *capacitance* (or capacity) is called the *farad* and is the capacitance of a capacitor which holds a charge of one coulomb when the voltage across its plates is one volt. A farad is too large a unit for practical purposes and capacitors are usually measured in *microfarads* (μF), one millionth of a farad and in micro-microfarads ($\mu\mu$F) or *picofarads* (pF), one million-millionth of a farad.

The capacitance of a capacitor depends on the area of its plates and on the distance by which they are separated and the material between them. When the space between the plates of a capacitor is occupied by some other insulating medium than air the capacity of the capacitor is increased, the area and spacing being assumed the same in both cases. The factor by which the dielectric increases the capacity compared with air is called the *dielectric constant* of the material. This factor is sometimes called the *permittivity* of the material and is denoted by the symbol K. Typical values of K are—air 1, paper 2·5, glass 5, mica 7. Certain ceramics have much higher values of K. If the dielectric is a vacuum, as in the case of the inter-electrode capacity in a valve the same value of K as for air may be assumed. (Strictly, $K = 1$ for a vacuum and is slightly higher for air.) The voltage at which a capacitor breaks down depends on the spacing between the plates and the type of dielectric used. Capacitors are often labelled with the maximum working voltage which they are designed to withstand and this figure should not be exceeded.

Fig. 1.8 shows an elementary capacitor made of two parallel metal plates of area A sq. cm separated by a distance d cm. If the plates are of unequal area, only the smaller one

Fig. 1.8. Parallel-plate capacitor. The capacitance is proportional to the area A and inversely proportional to the spacing *d*.



should be taken into account since it is actually the cross-sectional area of the active dielectric which determines the capacity. If the dielectric constant of the material between the plates is K, the capacitance of the capacitor will be—

$$C = \frac{0{\cdot}0885\ KA}{d} \text{ pF}$$

where A is in square centimetres and d is in centimetres. If A is in square inches and d is in inches the formula becomes—

$$C = \frac{0{\cdot}224\ KA}{d} \text{ pF}$$

As an example, suppose two metal plates each 1 square inch in area are spaced 0·004 inch apart in air, the capacity would be—

$$C = \frac{0{\cdot}224 \times 1 \times 1}{0{\cdot}004} \text{ pF} = 56 \text{ pF}$$

If mica (for which the value of K is 7) had been used as a dielectric the capacitance would be—

$$C = \frac{0{\cdot}224 \times 7 \times 1}{0{\cdot}004} = 392 \text{ pF.}$$

Capacitors used in Radio Equipment

The values of capacitors used in radio equipment extend from about 1 pF to 1000 μF. They are designed for various

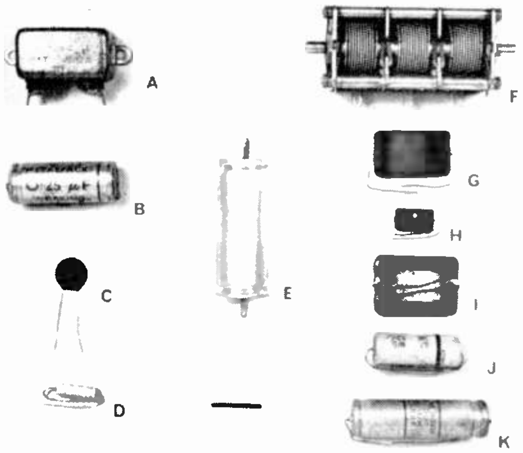


Fig. 1.9. Typical capacitors used in radio equipment.

A —Metal-cased block paper, chassis mounting.
B —Metal-cased tubular paper with insulating sleeve.
C —Disc ceramic.
D—Tubular ceramic.
E —Electrolytic.
F —Precision three-gang tuning capacitor, silver plated brass.
G—Mica, encapsulated.
H—Mica, miniature encapsulated.
I —Mica, wax coated.
J —Tubular paper.
K —Metal-cased tubular paper

working voltages and are of many different types of construction. The important varieties are described below.

(a) Air-dielectric Capacitors (Capacity range 5-1000 pF max.).

Air-dielectric capacitors are usually of the variable type such as are used for tuning transmitters and receivers. High-voltage types for transmitters have relatively large spacings between the plates and the construction of a typical ganged air-spaced variable capacitor is shown at *F* in Fig. 1.9.

(b) Mica Capacitors (Capacity range 5-10,000 pF).

Stacked-foil type. These are made from alternate thin metal plates and sheets of mica, stacked one above the other, clamped, and usually moulded in Bakelite as shown at *G* in Fig. 1.9. They are commonly used as bypass capacitors in radio-frequency circuits.

Silvered-mica type. These are made by spraying a coat of silver on each side of a sheet of mica to form the electrodes or plates. Silvered-mica capacitors are highly stable in value and are used in r.f. equipment where high frequency stability is required, e.g. in oscillators.

(c) Paper Capacitors (Capacity range 0·001-10 μF).

Paper capacitors, shown at *A*, *B*, *J* and *K* in Fig. 1.9, are made from long strips of thin aluminium foil and paper which are wound into a tight roll and then either moulded in Bakelite or sealed in cardboard or metal tubes, or for the larger sizes in rectangular metal cans. Another type of paper capacitor uses metallized paper electrodes in which the foils are replaced by metal sprayed on each side of a strip of paper. This type is smaller than the foil type for a given capacity and working voltage.

(d) Ceramic Capacitors (Capacity range 1 pF–0.01 μF).

Ceramic capacitors are made by coating two sides of a ceramic disc, tube, or cup, with silver. Ceramic materials having a high dielectric constant are available so that relatively high capacities can be obtained in a small physical size. They are used, for example, as bypass capacitors in very high frequency equipment, where small size and low inherent inductance are necessary. They should not be used in oscillators, however, as they usually vary considerably in capacity with temperature changes, the change being positive or negative depending on the type of ceramic material used. Ceramic capacitors are shown at *C* and *D* in Fig. 1.9. Ceramic capacitors used for temperature compensation of stable oscillators.

(e) Electrolytic Capacitors (Capacity range 2–2000 μF).

Electrolytic capacitors are made of aluminium-foil plates and have a semi-liquid conducting compound, often in the form of a special impregnated paper, between them. The dielectric is a thin insulating layer which is formed by electro-chemical action on one of the foils when a d.c. polarizing voltage is applied to the capacitor. This film is so thin that it is possible to make large-capacity electrolytic capacitors of small physical size. There is an upper limit to the working voltage of electrolytic capacitors of about 700 volts and there is also some leakage of current between the plates inside the capacitor. An electrolytic capacitor is

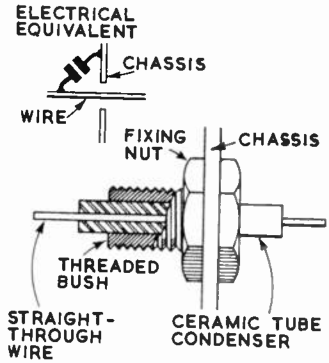


Fig. 1.10 (a). Feed-through capacitor.

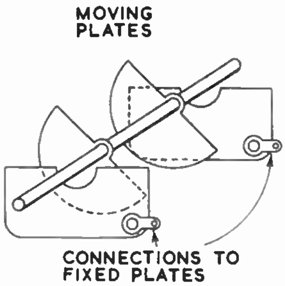


Fig. 1.10 (b). Basic construction of a split-stator capacitor.

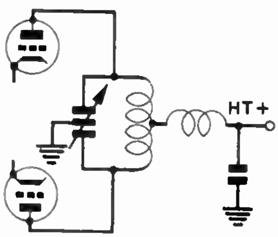


Fig. 1.10 (c). Push-pull circuit using a split-stator capacitor.

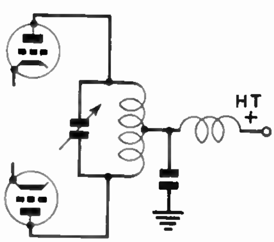


Fig. 1.10 (d). Push-pull circuit using an ordinary single-section capacitor.

shown at E in Fig. 1.9. Electrolytic capacitors are generally used for reducing the ripple voltage in power supply units, and also for decoupling purposes. In d.c. circuits the correct polarity of electrolytic capacitors must be observed, or permanent damage will result: the capacitor may even explode. The maximum working voltage must not be exceeded or similar results may ensue.

(f) Feed-through Capacitors

There is always a small amount of inductance associated with a capacitor, partly in the connecting leads and partly in the capacitor. At very high frequencies (above about 30 Mc/s) it is often difficult to obtain an effective low-reactance path to earth through a bypass capacitor because of this stray inductance. Feed-through capacitors are specially designed to overcome this difficulty. They consist of a ceramic tube, silvered inside and outside to give the required capacity. The tube is soldered to a bush which may be bolted to a chassis and thereby give a very low inductance path from the outside silvering to earth. The inner silvering is connected to a wire which passes through the tube and protrudes at each end as shown in **Fig. 1.10 (a)**, the whole device forming a very convenient method of decoupling or filtering supply leads where they enter or leave a screened compartment.

(g) Split-stator Capacitor

A split-stator capacitor is a special type of variable capacitor which is used mainly in push-pull radio-frequency amplifier circuits in transmitters. It is made up of two separate sets of fixed and moving plates as shown diagrammatically in **Fig. 1.10 (b)**. The advantages of the push-pull circuit of **Fig. 1.10 (c)** using a split-stator capacitor over that of **Fig. 1.10 (d)** using an ordinary variable capacitor are that the circuit is symmetrically balanced to earth and the shaft is at earth potential. In the case of the circuit with the split-stator capacitor the balance of the circuit is not so dependent on the position of the tapping point on the coil.

Capacitors in Series and Parallel

Capacitors can be connected in series or parallel, as shown in **Fig. 1.11**, either to obtain some special capacity value using a standard range of capacitors, or perhaps in the case of series connection to obtain a capacitor capable of withstanding a greater voltage without breakdown than is provided by a single capacitor. When capacitors are connected in parallel, as in **Fig. 1.11 (A)**, the total capacity of the combination is equal to the sum of the separate capacities. When capacitors are connected in series, as in **Fig. 1.11 (B)**, the reciprocal of the equivalent capacity is equal to the sum of the reciprocals of the separate capacities.

If C is the total capacity these formulae can be written as follows:

PARALLEL CONNECTION	SERIES CONNECTION
$C = C_1 + C_2 + C_3$ etc.	$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$ etc.

Similar to the formula for *resistors in parallel,* a useful equivalent formula for two *capacitors in series* is—

$$C = \frac{C_1C_2}{C_1 + C_2}$$

The use of these formulae is illustrated by an example taken from a Radio Amateurs' Examination paper.

Example. *Two capacitors of 4 and 12 pF are connected in series; two others of 8 and 24 pF are also connected in series. What is the equivalent capacitance if these series combinations are joined in parallel?*

The circuit is shown in **Fig. 1.11 (C)**. Using the formula for two capacitors in series the two series arms X and Y can be reduced to single equivalent capacities C_X and C_Y as shown in **Fig. 1.11 (D)**. Thus—

$$C_X = \frac{4 \times 12}{4 + 12} = \frac{4 \times 12}{16} = 3 \text{ pF}$$

$$C_Y = \frac{8 \times 24}{8 + 24} = \frac{8 \times 24}{32} = 6 \text{ pF}$$

These two capacitances may now be compounded in parallel to give the total effective capacity represented by the single capacitor C_T in **Fig. 1.11 (E)**.

$$C_T = C_X + C_Y = 3 + 6 = 9 \text{ pF}$$

The total equivalent capacity of the four capacitors connected as described is therefore 9 pF.

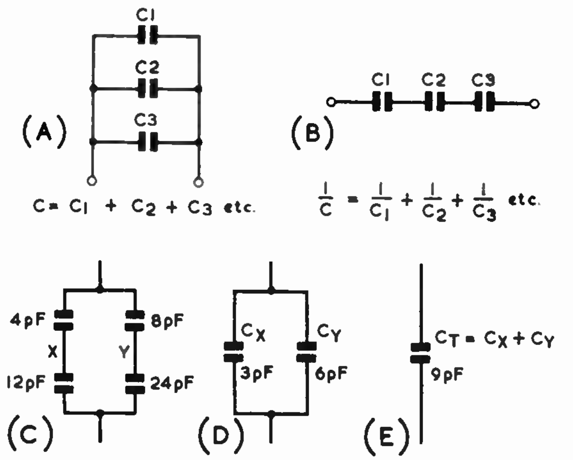


Fig. 1.11. Capacitors in various combinations: (A) parallel, (B) series, (C) series-parallel. The calculation of the resultant capacitance of the combination shown in (C) requires first the evaluation of each series arm *X* and *Y* as shown in (D): the single equivalent capacitance of the combination is shown in (E).

MAGNETISM

Permanent Magnets. A magnet will attract pieces of iron towards it by exerting a magnetic force upon them. The field of this magnetic force can be demonstrated by sprinkling iron filings on a piece of thin cardboard under which is placed a bar magnet. The iron filings will map out the magnetic field as sketched in **Fig. 1.12 (A)**. It will be seen that the field is most intense near the ends of the magnet, the centres of intensity being called the *poles*, and *lines of force* spread out on either side and continue through the material of the magnet from one end to the other.

If such a magnet is suspended so that it can swing freely in a horizontal plane it will always come to rest pointing in one particular direction, namely towards the earth's magnetic poles, the earth itself acting as a magnet. A compass needle is simply a bar of magnetized steel. One end of the magnet (N) is called a *north pole*, which is an abbreviation of " north-seeking pole " and the other end (S) a *south pole* or " south-seeking pole." It is an accepted

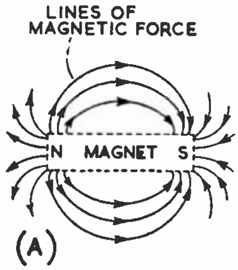


Fig. 1.12 (A). Magnetic field produced by a bar magnet.

convention that magnetic force acts in the direction from *N* to *S* as indicated by the arrows on the lines of force in Fig. 1.12 (A).

If two magnets are arranged so that the north pole of one is near the south pole of another, there will be a force of attraction between them, whereas if similar poles are opposite one another, the magnets will repel one another: see **Fig. 1.12 (B)**.

Magnets made from certain kinds of iron and nickel alloys retain their magnetism more or less permanently, and find many uses in radio equipment, such as a loudspeakers, polarized relays, headphones, cathode-ray tube focusing arrangements and magnetron oscillators.

Other types of iron and nickel alloys, e.g. soft iron, are

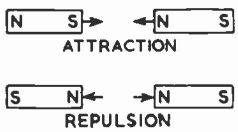


Fig. 1.12 (B). Attraction and repulsion between bar magnets.

not capable of retaining magnetism, and therefore cannot be used for making " permanent " magnets. They are effective in transmitting magnetic force, however, and are used as cores in electromagnets and transformers.

Electromagnets. A current of electricity flowing through a straight wire exhibits a magnetic field, the lines of force of which are in a plane perpendicular to the wire and concentric with the wire. If a piece of cardboard is sprinkled with iron filings, as shown in **Fig. 1.12 (C)**, they will arrange themselves in rings round the wire, thus illustrating the magnetic field associated with the flow of current in the wire. Observation of a small compass needle placed near the wire

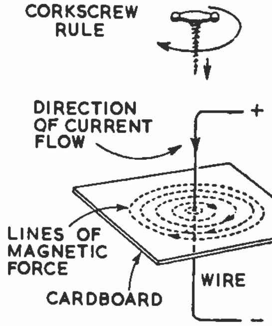


Fig. 1.12 (C). Magnetic field produced by current flowing in a straight wire.

would indicate that for a current flow in the direction illustrated the magnetic force acts clockwise round the wire. A reversal of current would reverse the direction of the magnetic field.

The *corkscrew rule* enables the direction of the magnetic field round a wire to be found. Imagine a " right-handed " corkscrew being driven into the wire so that it progresses in the direction of current flow; the direction of the magnetic field around the wire will then be in the direction of rotation of the corkscrew.

The magnetic field surrounding such a straight wire is relatively weak, but a strong magnetic field can be produced by a current if instead of a straight wire a coil of wire or *solenoid* is used: moreover the field can be greatly strengthened if a piece of soft iron, called a *core*, is placed inside the

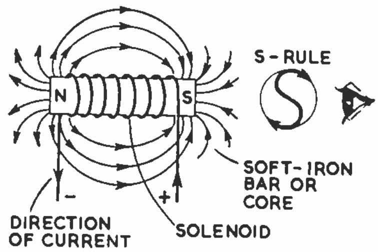


Fig. 1.12 (D). The " S " rule for determining the polarity of an electromagnet.

coil. The extent by which the strength of the solenoid magnet is increased by the introduction of the core is called the *permeability* of the core (*cf.* dielectric constant). **Fig. 1.12 (D)** shows the magnetic field produced by a solenoid, which it will be seen is very similar to that of a bar magnet as shown in Fig. 1.12 (A). A north pole is produced at one end of the coil and a south pole at the other. Reversal of the current will reverse the polarity of the electromagnet. The polarity of a solenoid can be deduced from the " *S* " rule, which states that the pole which faces an observer looking at the end of a solenoid is a south pole if to him the current is flowing (i.e. from positive to negative) in a clockwise direction; see Fig. 1.12 (D).

The strength of a magnetic field produced by a current is directly proportional to the current, a fact made use of in moving coil meters. It also depends on the number of turns of wire, the area of the coil, and the permeability of the core.

Interaction of Magnetic Fields. Similar to the attraction and repulsion of permanent magnets, there can be interaction between the fields of electromagnets or between a permanent magnet and an electromagnet, and in just the same way the interaction is manifest as a force causing relative motion between the two. This interaction forms the basis of several types of electromechanical devices which are used in radio such as loudspeakers, earphones, and moving-coil meters.

The principle of the loudspeaker is shown in **Fig. 1.13 (A)**. A strong permanent magnet *A* forms part of a soft-iron magnetic circuit *BCD* which has a narrow annular gap between the circular pole-pieces *C* and *D*: a strong radial magnetic field is thereby produced in this gap. Speech-frequency currents are passed through the coil *E* which is free to move in the annular gap and is fixed to a cone *F*. Interaction between the magnetic field in the gap and that due to the current flowing in the coil causes the coil to move

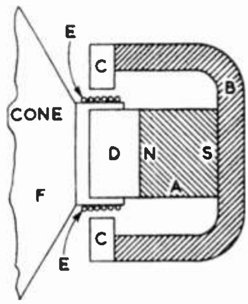


Fig. 1.13 (A). The principle of the moving coil loudspeaker.

backwards or forwards according to the strength and polarity of the speech signal: the cone therefore vibrates in sympathy with the speech currents and causes sound waves to be produced.

The principle of the earphone is shown in **Fig. 1.13 (B).** A permanent magnet M is situated close to a circular diaphragm Z made of magnetic material. The permanent magnet's field holds the diaphragm in a state of stress. Around the arms of the U-shaped magnet are two coils wound with many turns of fine wire and through which speech currents are passed. The magnetic field due to the speech current flowing in the coils adds to or subtracts from the field holding the diaphragm in a state of stress causing it to vibrate in sympathy with the speech signal and emit corresponding sound waves.

The principle of the moving-coil meter is discussed in Chapter 19 (*Measurements*).

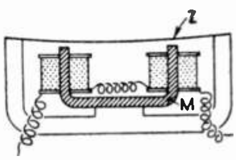


Fig. 1.13 (B). The principle of the earphone.

ELECTROMAGNETIC INDUCTION

If a bar magnet is plunged into a solenoid coil as indicated in **Fig. 1.14 (A)** the moving-coil microammeter connected across the coil will show a deflection. The explanation of this phenomenon, known as *electromagnetic induction*, is that the movement of the magnet's lines of force past the turns of the coil cause a voltage to be induced in the coil which in turn causes a current to flow through the meter. The magnitude of the effect depends on the strength and rate of movement of the magnet and the size of the coil. Withdrawal of the magnet causes a reversal of the current. No current flows unless the lines of force are moving relatively to the coil. The same effect is obtained if a coil of wire is arranged to move relatively to a fixed magnetic field. Dynamos and generators depend for their operation on the principle of electromagnetic induction.

If a pair of coils of wire are arranged as shown in **Fig. 1.14 (B)**, when the switch K is open there is no magnetic field from the coil P linking the turns of the coil S, and the current through S is zero. Closing K will cause a magnetic field to appear due to the current in the coil P. This field, as it builds up from zero, will induce a voltage in S and cause a current to flow through the meter for a short time until the field due to P has reached a steady value when the current through S falls to zero again. The effect is only momentary and is completed in a small fraction of a second. The change in current in the circuit P is said to have induced a voltage in the circuit S. The fact that a changing current in one circuit can induce a voltage in another circuit is the principle underlying the operation of transformers.

Self-Inductance

If a steady current is flowing through a coil there will be a steady magnetic field due to that current. A current change will tend to alter the strength of the field which in turn will induce in the coil a voltage tending to oppose the change being made. This process is called *self-induction.* A coil is said to have *self-inductance*, usually abbreviated to *inductance*. It will have a value of one *henry* (H) if, when the current through the coil changes at a rate of one ampere per second, the voltage appearing across its terminals is one volt. Inductance is usually denoted by the symbol L in formulae. As the inductance values used in radio equipment may be only a very small fraction of a henry the units *millihenry* (mH) and *microhenry* (μH) meaning one thousandth and one millionth of a henry respectively are commonly used.

The inductance of a coil depends on the number of turns and on the area of the coil and the permeability of the core material on which the coil is wound. The inductance of a coil of a certain physical size and number of turns can be calculated to a fair degree of accuracy from formulae or they can be derived from coil charts.

Mutual Inductance

A changing current in one circuit can induce a voltage in a second circuit: see Fig. 1.14 (B). The strength of the voltage induced in the second circuit S depends on the closeness or " tightness " of the magnetic coupling between the circuits; for example, if both coils are wound together on an iron core practically all the lines of force or magnetic flux from the first circuit will link with the turns of the second circuit. Such coils would be said to be *tightly coupled* whereas if the coils were both air-cored and spaced some distance apart they would be *loosely coupled.*

The mutual inductance between two coils is measured in henrys, and two coils are said to have a *mutual inductance* of one henry if when the current in the primary coil changes at

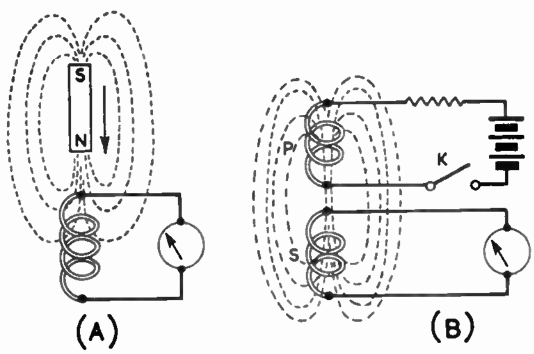


Fig. 1.14. Electromagnetic induction: (A) Relative movement of a magnet and a coil causes a voltage to be induced in the coil; (B) when the current in one of a pair of coupled coils changes in value, a current is induced in the second coil.

a rate of one ampere per second the voltage across the secondary is one volt. Mutual inductance is denoted in formulae by the symbol M.

The mutual inductance between two coils may be measured by joining the coils in series (*a*) so that the sense of their windings is the same and (*b*) so that they are reversed. The total inductance is then measured in each case.

If L_a and L_b are the total measured inductances, and L_1 and L_2 are the separate inductances of the two coils and M is the mutual inductance—

$$L_a = L_1 + L_2 + 2M$$
$$L_b = L_1 + L_2 - 2M$$
$$\therefore L_a - L_b = 4M$$
$$\text{i.e.} \quad M = \frac{L_a - L_b}{4}$$

The mutual inductance is therefore equal to one-quarter of the difference between the series-aiding and series-opposing readings.

Inductors used in Radio Equipment

Inductors, like capacitors, are used in radio equipment where a circuit is required to be frequency-dependent. For instance they are used in conjunction with capacitors to form tuned circuits which select wanted from unwanted frequencies in receivers and transmitters, and they are also used in the form of *chokes* in filters which remove hum in high-tension power supplies.

The value of inductance used in any circuit depends largely on the frequency at which the circuit is operating: for example, coils might be of many henrys inductance for use as power-supply filter chokes where the frequency is 50 or 100 cycles per second or as small as 0·1 microhenry when used as tuning coils at a frequency of, say, 144 Mc/s.

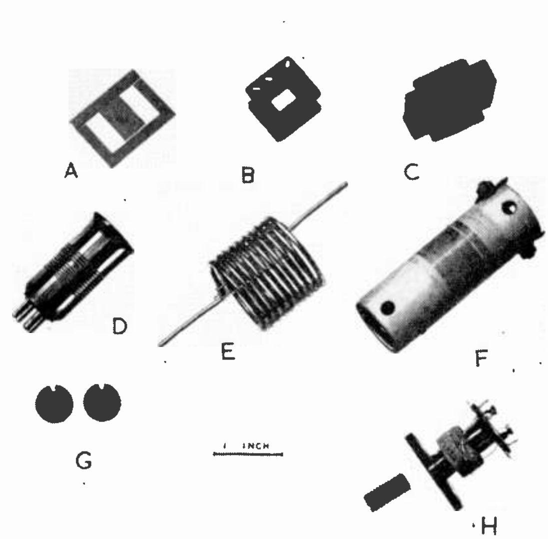


Fig. 1.15. Typical inductances and magnetic cores. A — Iron laminations (T- and U-type); B — Bobbin; C — Assembled iron-cored inductance; D — Plug-in air-cored coil; E — Low-loss transmitter tank coil; F — High-stability oscillator coil; G — Small ferrite "pot" core; H — Wave-wound coil with adjustable iron core.

Broadly speaking, coils can be divided into two classes—(*a*) low-frequency coils or chokes and (*b*) radio-frequency coils and chokes. Low-frequency chokes range in inductance from 0·5 to 500 henrys and are wound with many turns of wire on iron cores. In **Fig. 1.15** the general method of construction of such chokes is shown in *A*, *B* and *C*. The core is made up of thin soft iron or special-alloy stampings or laminations *A*, insulated from each other so as to reduce the power lost due to eddy currents induced in the core. The coil is wound on a bobbin *B*, which is then packed with laminations as shown at *C*. Laminations may be obtained in a wide range of sizes and permeabilities.

Radio-frequency coils are air-cored, dust iron-cored or ferrite cored. Air-cored coils are usually wound on low-loss cylindrical or ribbed formers. In Fig. 1.15 a plug-in coil such as is often used in amateur short-wave transmitters and receivers is shown at *D*. For maximum efficiency, coils should be of such dimensions that the length of the winding is not less than half and not greater than twice the diameter.

A perfect radio-frequency coil would have inductance only and no resistance, and the current through it and the voltage across it would be *in quadrature* (i.e. 90° out-of-phase) and no power would be lost in the coil. In practice, however, coils have an appreciable effective series resistance, which may be many times the d.c. resistance of the coil, and power is wasted in the coil. This power loss shows up in the case of a transmitting coil, where currents are high in a rise in temperature of the coil and its former. Coils in which this loss resistance is low are said to be "low-loss" or "high Q." The losses in a coil at radio frequencies are due to (*a*) eddy currents in surrounding metal objects, including the core if there is one, (*b*) skin effect (see *Resistance and Conductance*), and (*c*) dielectric loss in the coil former. For a low-loss coil such as would be used in the output tank circuit of a short-wave transmitter, heavy-gauge wire (e.g. No. 14 s.w.g.) is used, wound as shown at *E* in Fig. 1.15 with the minimum of dielectric material inside, the turns of wire being supported by strips of good-quality insulating material which can just be seen cemented to the wires. The coil should be mounted so that there is no part of the metal chassis or screening within a distance at least equal to the diameter of the coil in any direction. For high frequency-stability, such as is needed in a transmitter master-oscillator, coils should be wound under tension on grooved ceramic formers: an example is shown at *F* in Fig. 1.15.

For larger inductances, such as would be used in intermediate-frequency amplifiers, the coils are usually fitted with a special kind of iron core (" dust iron ") the permeability of which allows fewer turns of wire to be used for a given inductance; the resistance is thereby decreased and the efficiency of the coil is increased. Solid iron cores cannot be used since the eddy-current losses would more than offset the gain in coil efficiency just mentioned, and a fine iron-alloy powder moulded in an insulating medium is normally used. Rather higher efficiency can be obtained by the use of *ferrites* which are non-metallic magnetic materials of very high resistivity and therefore low eddy-current loss. In Fig. 1.15, *G* shows two halves of a ferrite " pot " core. The coil is wave-wound, or simply wound on a bobbin and placed inside the two halves so that it is completely surrounded by the ferrite material

A typical i.f. coil of wave-wound construction with an adjustable screw-type dust-iron core is shown at H in Fig. 1.15. The variation of the position of the dust-iron core alters the inductance value. Sometimes an adjustable brass core is used instead of the iron core, but whereas the effect of the iron is to *increase* the inductance because of its high magnetic permeability the effect of the brass is to *reduce* the inductance; the reason for this is that the brass, besides being non-magnetic, has a very low resistance and therefore acts as an efficient short circuited single turn secondary. Although copper would theoretically give a superior performance, silver-plated brass has a sufficiently high conductivity and is the material normally employed.

The use of magnetic or conductive cores to control the inductance value of a coil is known as *slug tuning*.

For convenience in aerial tuning in a transmitter a tapped coil may be used. If a continuously variable inductance is required the coil can be made to rotate while a variable tapping is made by a sliding pulley following the wire spiral. This type of variable inductance has the disadvantage of relying on three rubbing contacts in series, viz. between the pulley and the wire, between the pulley and the rod on which it moves, and the pivot at the end of the coil. Nevertheless it is capable of giving satisfactory service if properly maintained.

A variable inductance with no rubbing contacts may be made by using two series-connected coils which can be turned relatively to one another through 180° from the series-aiding to series-opposing condition. Such a device depends on the variation of the mutual inductance between the two coils and is called a *variometer*.

Inductors in Series and Parallel

Provided that there is no mutual coupling between inductors when they are connected in series, the total inductance obtained is equal to the sum of the separate inductances. When they are in parallel the reciprocal of the total inductance is equal to the sum of the reciprocals of the separate inductances.

If L is the total inductance (no mutual coupling) the relationships are as follows:

SERIES CONNECTION	PARALLEL CONNECTION
$L = L_1 + L_2 + L_3$ etc.	$\frac{1}{L} = \frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3}$ etc.

For the special case of *two* inductances in parallel—

$$L = \frac{L_1 L_2}{L_1 + L_2}$$

It may be noted that this formula is of the same type as that relating to two *resistors in parallel* or *capacitors in series*. The use of these series and parallel inductance formulae is illustrated by the following example taken from a Radio Amateurs' Examination paper:

Example. *Two inductors of 10 and 20 microhenrys are connected in series; two others of 30 and 40 microhenrys are also connected in series. What is the equivalent inductance if these series combinations are connected in parallel? Assume that there is no mutual induction.*

The 10 and 20 μH coils in series are equivalent to (10 + 20) = 30 μH.

The 30 and 40 μH coils in series are equivalent to (30 + 40) = 70 μH.

These two equivalent inductances of 30 μH and 70 μH respectively are in parallel and will therefore be equivalent to one single inductance of—

$$\frac{30 \times 70}{30 + 70} = 21 \mu\text{H}$$

This is the value of the equivalent inductance of the four coils in this series-parallel arrangement.

Time Constant

Fig. 1.16 (A) shows a circuit in which a capacitor C can either be charged from a battery of voltage E, or discharged, through a resistor R according to whether the switch S is in position a or b.

If at some instant, t_a, the switch is thrown from b to a current will start to flow into the capacitor with an initial value E/R. As the capacitor charges the voltage across the plates increases and the current through the circuit therefore falls away, as shown in the charging portion of Fig. 1.16 (b); when fully charged to the voltage E the current will have dropped to zero.

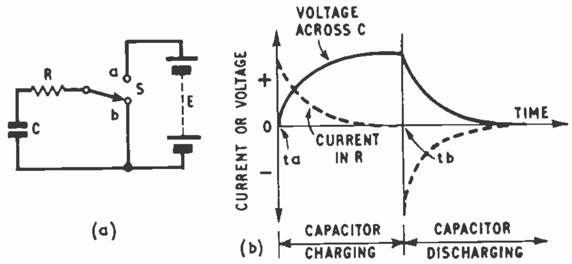


Fig. 1.16. In (a) a capacitor C can be charged or discharged through the resistance R by operating the switch S. The curves of (b) show how the voltage across the capacitor and the current into and out of the capacitor vary with time as the capacitor is charged and discharged. The curve for the rise and fall of current in an LR circuit is similar to the voltage curve for the CR circuit.

When, at some time t_b, the switch is thrown back to b the capacitor will discharge through the resistor R, the current being in the opposite direction to the charging current, starting at a value $-E/R$ and dying away to zero. As the capacitor discharges the voltage across its plates falls to zero as shown in the discharge portion of Fig. 1.16 (b). Graphs of this shape are called *exponential* curves.

It is evident that a finite time is required to charge and discharge a capacitor through a resistor; the larger R or C the longer the time taken. This time is specified by the *Time Constant* of the CR circuit, being the time taken for a capacitor to discharge to 37 per cent, i.e. $\frac{1}{\epsilon}$th, of its initial voltage, or to charge to 63 per cent $(1-\frac{1}{\epsilon})$th, of its final voltage. The time constant, in seconds, is given by the product of the capacity in Farads and the resistance in ohms:

Thus $T = CR$ seconds.

As an example, the time constant of a capacity of 0·1 microfarads (10^{-7} Farads) and a resistance of 47 Kilohms ($4{\cdot}7 \times 10^4$ ohms) is:

$$\begin{aligned} T &= 10^{-7} \times 4{\cdot}7 \times 10^4 \\ &= 4{\cdot}7 \times 10^{-3} \text{ seconds} \\ &= 4{\cdot}7 \text{ milliseconds} \end{aligned}$$

Examples, in which the time constants of CR circuits are important, are to be found in radio receiver circuits associated with the second detector and a.g.c. rectifier. The time constant of the signal detector diode load and reservoir capacitor is chosen to be long compared with the period of the i.f. signal to give good smoothing, or filtering, but not so long that the voltage across the load resistor cannot follow the highest speed audio amplitude variations likely to be encountered. On the other hand the time constant of the a.g.c. circuit is chosen to be long compared with the period of the lowest audio frequency amplitude variations to avoid suppression, due to negative feedback, of the modulation in the a.g.c. controlled stages, but at the same time short enough to follow the fastest fading likely to be encountered.

Time constants are very important in pulse circuits which are expected to handle fast transient waveforms.

The rise and fall of current in an inductance in circuit with a resistance also takes a definite time depending on the values of L and R in question. The larger the value of L and the smaller the value of R the longer it takes for the current to rise or decay. The shape of the current curves for an LR circuit are similar in shape to the voltage curves for the CR circuit shown in Fig. 1.16 (b).

The time constant of an LR circuit is defined as the time taken for the current to reach 63 per cent, $(1-\frac{1}{e})$th, of its final value, or to decay to 37 per cent $\frac{1}{e}$th of its initial value. The time constant in seconds is given by the inductance in Henrys divided by the resistance in ohms:

Thus: $$T = \frac{L}{R} \text{ seconds.}$$

ALTERNATING CURRENT

Previous sections have been concerned mainly with current flowing in one direction through a circuit, such a current being produced, for example, by a battery, i.e. *direct current* (usually abbreviated to *d.c.*).

There is another important type of current flow in which current flows backwards and forwards alternately through a circuit. Such currents are called *alternating currents* (*a.c.*).

An alternating current could be obtained by connecting a resistor to a battery through a reversing switch as shown in **Fig. 1.17 (A)**. On operating the switch S backwards and forwards from a to b the current from the battery would flow in alternate directions through the resistor R; the graph in **Fig. 1.17 (B)** shows how the current would vary with the passage of time and the reversals of the switch. Such a current waveform, in which the current flows at a steady value for equal times in both directions is called a *square-wave.*

The most common alternating current is that used for the electricity supply mains in which, in Great Britain and most of Europe, the direction of current flow is continuously changing from positive to negative and back again at a rate of 50 cycles per second. In the USA and Canada the rate, or frequency, is 60 cycles per second. In each cycle the direction changes twice, but for the purpose of most calculations it is the complete cycle which is more significant than the two half-cycles in opposite directions.

The alternating voltage of the electricity supply does not have a square waveform but fluctuates with a gradual and smooth variation like the swinging of a pendulum. Such a current would be produced by the continuous and steady rotation of a coil in a magnetic field as indicated in **Fig. 1.17 (C)**, when the circuit is completed by some device such as a resistor. This is the basic principle used in the *alternator*, a machine for generating alternating current. The waveform of this current has a shape which is defined mathematically as a *sine-wave*, shown in **Fig. 1.17 (D)**, and the current is said to vary *sinusoidally.*

Alternating currents of much higher frequency are used in radio communication, the frequencies commonly used in short-wave amateur work ranging between 1,800,000 and 30,000,000 cycles per second (1·8-30 Mc/s) and in more advanced work up to 430,000,000 cycles per second (430 Mc/s) or even higher.

Characteristics of Alternating Currents

From the preceding paragraph it can be seen that one feature which distinguishes one alternating current from another is the rate at which the complete cycles of current-reversal take place; this is called the *frequency* of the alternating current and is measured in *cycles per second.*

A second distinguishing feature of an alternating current is its magnitude or *amplitude*, by which is meant the maximum value reached during one cycle or alternation: see **Fig. 1.17 (D)**.

It is possible to have two alternating currents whose frequency and amplitude are exactly equal but with a time

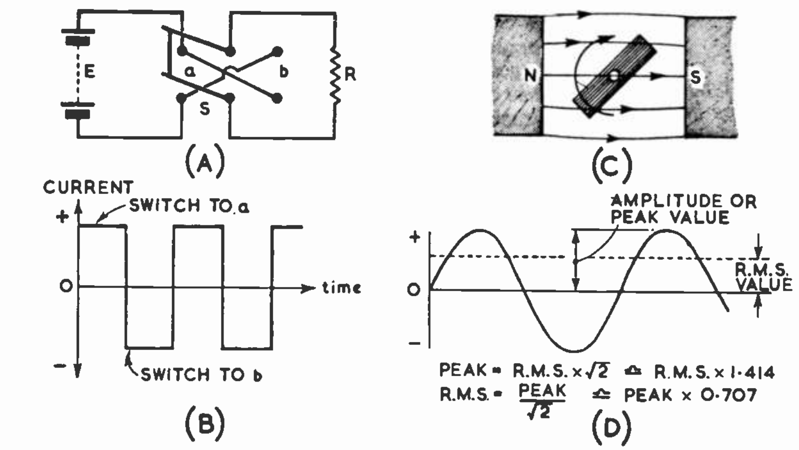


Fig. 1.17. Alternating current. A simple circuit with a current-reversing switch shown at (A) produces a square-wave current through the resistor R as shown in (B). When a coil is rotated in a magnetic field as in (C) the voltage induced in the coil has a sinusoidal waveform (D).

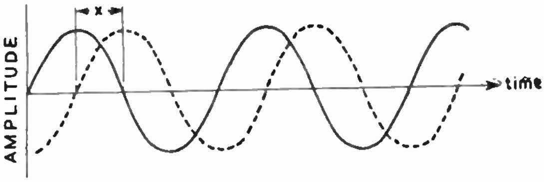


Fig. 1.18. Phase difference between two alternating currents. Provided that they both have the same frequency the phase difference will remain constant.

lag, or *phase difference*, between the two so that they are not performing the same part of their cycles at the same instant. This is illustrated in **Fig. 1.18** in which the solid and dotted sine waves are of the same amplitude and frequency but differ in phase by one-quarter of the time taken for one cycle as indicated by the time x in the diagram.

Thus an alternating current, or indeed any alternating quantity, can be defined by the three properties—frequency, amplitude and phase.

Another term used with reference to alternating current is the *period* of a wave, which is the time taken to perform one cycle. The period of a wave is the reciprocal of the frequency; thus—

$$Frequency = \frac{1}{Period} \qquad Period = \frac{1}{Frequency}$$

Frequency is usually denoted by the symbol f and period by the symbol T. As an example, the period of the 50 cycles per second mains is given by—

$$T = \frac{1}{f} = \frac{1}{50} \text{ second}$$

The value of an a.c. voltage or current can be specified by its amplitude. This is often called the *peak* voltage or *peak* current, being the highest value reached during a cycle. Thus the peak value of any a.c. voltage or current is equal to its amplitude (in volts or amperes) as shown in Fig. 1.17 (D).

The peak value, however, is not the most common way of specifying an alternating voltage or current. The usual value adopted is the *root-mean-square* or *r.m.s.* value which is equal to the d.c voltage or current which would produce the same amount of power in a resistive load as the alternating voltage or current would produce The r.m.s. value is less than the peak value, the two being related if the waveform is sinusoidal as follows:

$$\text{r.m.s. value} = \frac{\text{peak value}}{\sqrt{2}} = 0{\cdot}707 \times \text{peak value}$$

or conversely—

$$\text{peak value} = \text{r.m.s. value} \times \sqrt{2} = 1{\cdot}414 \times \text{r.m.s. value}$$

An a.c. mains supply of 240 volts r.m.s. thus has a peak voltage of 240 × 1·414 or approximately 340 volts.

A.C. Circuit Containing Resistance Only

Ohm's Law can be used to find the current flowing through a resistor when a certain alternating voltage is applied across its terminals. The current will flow backwards and forwards through the resistor under the influence of the applied voltage and will be in phase with it. The voltage and current waveforms for the resistive circuit of **Fig. 1.19 (A)** are shown in **Fig. 1.19 (B)**.

The power dissipated in the resistor can be calculated direct from the usual power formulae, provided that r.m.s. values for voltage and current are used, viz.:

$$W = EI \qquad W = I^2R \qquad W = \frac{E^2}{R}$$

If peak values $\hat{E}$ for voltage and $\hat{I}$ for current are used these formulae become—

$$W = \frac{\hat{E}\hat{I}}{2} \qquad W = \frac{\hat{I}^2R}{2} \qquad W = \frac{\hat{E}^2}{2R}$$

A.C. Circuit Containing Capacitance Only: Reactance of Capacitor

When an alternating voltage is applied to a capacitor as shown in **Fig. 1.19 (C)** the capacitor will be alternately charged in one direction and then charged in the opposite direction as the supply alternates. At each reversal it becomes momentarily discharged. An observer watching the flow of current in the wires connecting the capacitor to the a.c. supply would see an alternating current flowing backwards and forwards through the wires. This current flows into and out of the capacitor and not actually through the capacitor. A careful study would show that the current is one-quarter period (or 90°) out-of-phase with the voltage, and in advance of it, as shown in **Fig. 1.19 (D)**. This must be so, for when the capacitor is fully charged the voltage is a maximum and the current is zero. No power is dissipated in such a circuit, a current of this type being called a reactive current or *wattless current*.

The magnitude of the current flowing in the circuit depends on the capacity of the capacitor and the frequency of the supply, and the capacitor therefore exhibits a property of current restriction or opposition somewhat similar

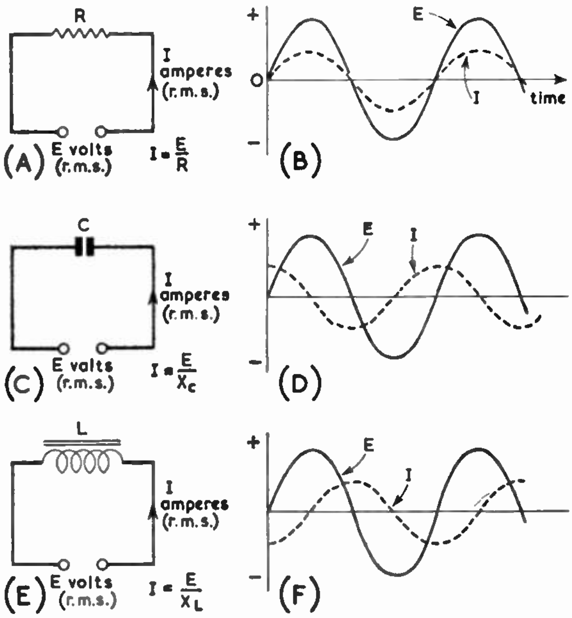


Fig. 1.19. Voltage and current relationships in a.c. circuits comprising (A) resistance only, (C) capacitance only, (E) inductance only.

to that exhibited by a resistor except that the effect is frequency-dependent and there is a 90° phase difference between the voltage and the current. This property, which is called the *capacitive reactance* of the capacitor is measured in ohms and is usually denoted by the symbol X_c; it is given by the formula—

$$X_C = \frac{1}{2\pi fC}$$

where f is the frequency in cycles per second and C is the capacity in farads: the value of π is 3·14.

Sometimes for convenience, the symbol ω is used to represent $2\pi f$. It is known as the *angular frequency* and is expressed in radians per second.

The following example, taken from a Radio Amateurs' Examination paper, shows how this formula is applied.

Example. *Calculate the reactance offered, at frequencies of 50 c/s and 50 kc/s respectively, of a capacitance of 2 microfarads.*

In the formula for capacitive reactance—

$$X_C = \frac{1}{2\pi fC} \text{ ohms}$$

Before any calculation is made it must be remembered to express f and C in the correct units.

Considering first the frequency of 50 c/s—

$$X_C = \frac{1}{2\pi \times 50 \times 2 \times 10^{-6}} \text{ ohms}$$

$$= \frac{10{,}000}{2\pi} = 1592 \text{ ohms}$$

Considering next the frequency of 50 kc/s (i.e. 50,000 c/s)—

$$X_C = \frac{1}{2\pi \times 50{,}000 \times 2 \times 10^{-6}} \text{ ohms}$$

$$= \frac{10}{2\pi} = 1{\cdot}592 \text{ ohms}$$

It is worth noting that the reactance is inversely proportional to the frequency, so that it would be possible to derive the reactance at 50 kc/s directly from the reactance at 50 c/s simply by dividing by 1,000.

The current flowing "through" a capacitor can be calculated by using Ohm's Law, regarding the reactance of the capacitor X_C as replacing the more usual resistance R, viz.;

$$I = \frac{E}{X_C} \qquad \left(cf.\ I = \frac{E}{R}\right)$$

A.C. Circuit Containing Inductance Only: Reactance of Coil

The opposition of an inductance to alternating current flow is called the *inductive reactance* of the coil: see **Fig. 1.19 (E)**. It is proportional to the inductance of the coil and also to the frequency and is denoted by the symbol X_L.

It may be calculated from the formula—

$$X_L = 2\pi fL$$

As in the case of a capacitor, the current in an inductance and the voltage across it are one-quarter period (or 90°) out-of-phase and no power is dissipated in a purely inductive circuit. In an inductive circuit, however, the phase of the current is 90° behind the voltage, i.e. exactly opposite to the case of the capacitive circuit: see **Fig. 1.19 (F)**.

Example. *Calculate the inductive reactance of a coil of 2 henrys inductance at frequencies of 50 c/s and 50 kc/s respectively.*

The formula for inductive reactance is—

$$X_L = 2\pi fL$$

Since f is 50 c/s and L is 2 henrys—

$$X_L = 2\pi \times 50 \times 2 \text{ ohms}$$
$$= 200 \times 3{\cdot}14 \text{ ohms}$$
$$= 628 \text{ ohms}$$

Because inductive reactance is directly proportional to frequency, the reactance of the 2H coil at 50 kc/s will be one thousand times its reactance at 50 c/s, namely 628,000 ohms.

The current through an inductance and the voltage across it are connected by the modified Ohm's Law formula—

$$I = \frac{E}{X_L} \qquad \left(cf.\ I = \frac{E}{R}\right)$$

Susceptance

A term sometimes used in calculations is the *susceptance* of a coil or a capacitor. Susceptance is simply the reciprocal of reactance in the same way that conductance is the reciprocal of resistance. Susceptance is in general denoted by the symbol B and is measured in *mhos*.

Thus—

$$B_L = \frac{1}{X_L} \qquad B_C = \frac{1}{X_C}$$

A.C. Circuit Containing L, C and R: Impedance

An a.c. circuit may contain resistance, inductance and capacitance in series, as shown in **Fig. 1.20 (A)**, each of which opposes the flow of current in a different way as mentioned in the previous sections. The opposition of the composite

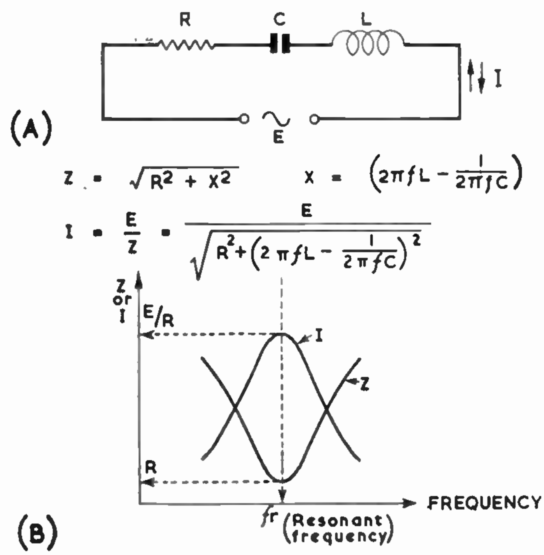


Fig. 1.20. The series-resonant circuit. The curves shown at (B) indicate how the impedance and the current vary with frequency in the type of circuit shown at (A).

circuit to alternating current flow is called the *impedance* of the circuit and is denoted by the symbol Z.

The impedance of the whole circuit cannot be found by simply adding the reactances and resistances together, since the relative current and voltage phases associated with each component are not the same. The inductive and capacitive reactance have exactly opposite phase effects and consequently they must first be subtracted from one another to find the total reactance in the circuit, thus—

$$X = (X_L - X_C)$$

The impedance is then found by compounding the resistance with the total reactance of the circuit. This has to be done by taking the square root of the sum of the squares of these two quantities; thus—

$$Z = \sqrt{R^2 + X^2}$$

The current flowing through the circuit may then be found using Ohm's Law but with Z replacing R as follows:

$$I = \frac{E}{Z}$$

If the current in the circuit is known, the voltage across each individual component can be calculated using Ohm's Law with the appropriate value of the resistance or reactance.

Series-Resonant Circuit: Acceptor Circuit

Considering the circuit of **Fig. 1.20 (A)**, current flowing may be calculated using the formula $I = E/Z$, where Z is the impedance of the circuit and E is the applied alternating voltage. When Z is written in terms of R, L, C and f, the expression of I becomes—

$$I = \frac{E}{\sqrt{R^2 + \left(2\pi fL - \frac{1}{2\pi fC}\right)^2}}$$

At one particular frequency depending on the exact values of C and L, the capacitive and inductive reactances will be equal and opposite. Under this condition, called *resonance*, the impedance of the circuit will be a minimum, equal to R and the current will be a maximum. Resonance is a very important phenomenon as it is used to select a desired frequency from other unwanted frequencies. **Fig. 1.20 (B)** shows how the current and impedance for the circuit of Fig. 1.20 (A) vary with frequency, the current at resonance being E/R and the impedance at resonance being R.

Series-resonant circuits are sometimes referred to as *acceptor circuits* because at the resonance frequency the impedance is a minimum and they therefore accept maximum current at this frequency.

The frequency at which a certain coil and capacitor resonate when connected together can be found by equating the inductive and capacitive reactances: thus—

$$2\pi fL = \frac{1}{2\pi fC}$$

Solving for f—

$$f = \frac{1}{2\pi\sqrt{CL}}$$

If L is in henrys and C is in farads, f will be in cycles per second.

This formula, next to Ohm's Law, is probably the most used of all in radio work as it permits the calculation of the inductance of a coil which will tune to a desired frequency with a given capacitance. Its use, together with other formulae relating to the resonant circuit, is illustrated by the following worked examples taken from various Radio Amateurs' Examination papers.

Example I. *What value of inductance is required in series with a capacitor of 500 pF for the circuit to resonate at a frequency of 400 kc/s? (Assume no resistance.)*

From the resonance formula—

$$f = \frac{1}{2\pi\sqrt{LC}}$$

the inductance is— $L = \dfrac{1}{4\pi^2 f^2 C}$

Expressing the frequency and the capacitance in the proper units ($f = 400 \times 10^3$ c/s and $C = 500 \times 10^{-12}$F)—

$$L = \frac{1}{4\pi^2 \times (400 \times 10^3)^2 \times (500 \times 10^{-12})}\ \text{H}$$

Taking $\pi^2 = 10$, this becomes—

$$L = \frac{1}{3200}\ \text{H}$$

$$= 312{\cdot}5 \text{ microhenrys.}$$

Example II. *If an inductance of 50 microhenrys is in series with a capacitance of 500 picofarads what is the resonant frequency? (π^2 may be taken as 10.)*

At resonance—

$$f = \frac{1}{2\pi\sqrt{LC}}$$

$$\text{or } f^2 = \frac{1}{4\pi^2 LC}$$

Expressing the inductance and the capacitance in the proper units ($L = 50 \times 10^{-6}$H and $C = 500 \times 10^{-12}$F)—

$$f^2 = \frac{1}{4 \times 10 \times (50 \times 10^{-6}) \times (500 \times 10^{-12})}$$

$$= 10^{12}$$

Therefore—

$$f = 10^6 \text{ c/s} = 1 \text{ Mc/s.}$$

Example III. *If the effective series inductance and capacitance of a vertical aerial are 20 μH and 100 pF respectively and the aerial is connected to a coil of 80 μH inductance what is the approximate resonant frequency?*

In this example the aerial and coil together will resonate at a frequency determined by the capacitance and the sum of the aerial effective inductance and the loading coil inductance.

At resonance—

$$f = \frac{1}{2\pi\sqrt{LC}}$$

Here the relevant values of inductance and capacitance expressed in the proper units are—

$$L = (20 + 80) \times 10^{-6}\text{H}$$

$$C = 100 \times 10^{-12}\text{F}$$

Therefore—

$$f = \frac{1}{2\pi\sqrt{(100 \times 10^{-6}) \times (100 \times 10^{-12})}} \text{ c/s}$$

$$= \frac{1}{2\pi \times 10^{-7}} \text{ c/s}$$

$$= 1{\cdot}6 \text{ Mc/s approximately.}$$

Example IV. *An alternating voltage of 10 volts at a frequency of 5/π Mc/s is applied to a circuit of the following elements in series: (i) a capacitance of 100 pF, (ii) a non-inductive resistor of 10 ohms.*

(*a*) *What value of inductance in series is required to tune the circuit to resonance?*

(*b*) *At resonance, what is the current in the circuit?*

(*a*) For the calculation of the inductance, the resistance can be ignored since it has no effect on the resonant frequency, which is given by—

$$f = \frac{1}{2\pi\sqrt{LC}}$$

Rearranged, this becomes—

$$L = \frac{1}{4\pi^2 f^2 C}$$

Expressing the frequency and the capacitance in the proper units ($f = 5 \times 10^6/\pi$ cycles per second; $C = 100 \times 10^{-12}$F)—

$$L = \frac{1}{4\pi^2 \times \left(\frac{25 \times 10^{12}}{\pi^2}\right) \times (100 \times 10^{-12})} \text{ henrys}$$

$$= \frac{1}{10{,}000} \text{ H}$$

$$= 100 \text{ microhenrys.}$$

(*b*) At resonance, the inductive and capacitive reactances cancel out and the circuit will have a purely resistive impedance of 10 ohms. The current I through the circuit at resonance can then be calculated directly from Ohm's Law: $I = E/R$. Since $E = 10$ volts and $R = 10$ ohms

$$\text{Current at resonance } I = \frac{10 \text{ volts}}{10 \text{ ohms}}$$

$$= 1 \text{ ampere.}$$

Magnification Factor: Q

At resonance the voltage across the coil (or the capacitor) in the circuit of Fig. 1.20 (A) can be considerably greater than that supplied to the circuit by the generator. The current at resonance is determined by the value of the resistor R whereas the voltage across the coil (or the capacitor) is given by the product of the current and the appropriate reactance which may be many times the value of R. The ratio of the voltage across the coil (or the capacitor) to that across the resistor is called the *magnification factor* or Q of the circuit. If I is the current at resonance—

$$Q = \frac{IX_L}{IR} = \frac{2\pi fL}{R} = \frac{\omega L}{R}$$

$$\text{or } Q = \frac{IX_C}{IR} = \frac{1}{2\pi fCR} = \frac{1}{\omega CR}$$

where $\omega = 2\pi f$

The Q of a tuned circuit is determined mainly by the coil since good-quality capacitors have negligible losses. The Q of a tuned circuit determines its *selectivity*, i.e. its ability to pick out a wanted signal from a number of unwanted signals on adjacent frequencies. A high Q-value corresponds to good selectivity.

Parallel-Resonant Circuit: Rejector Circuit

If a coil and capacitor are connected in parallel as in **Fig. 1.21 (A)** resonance effects appear when the current is of such a frequency that the reactance of the coil and capacitor are equal in magnitude and opposite in sign. The resonant frequency of a parallel-tuned circuit is, for practical purposes, given by the same formula as for the series circuit, viz.—

$$f = \frac{1}{2\pi\sqrt{LC}}$$

At resonance the impedance of a parallel-tuned circuit is resistive and is a maximum as shown in **Fig. 1.21 (B)**. If the coil and capacitor were perfect reactive components, the impedance at resonance would be infinite. In practice the capacitor can be looked upon as a nearly perfect reactance, but there is always appreciable resistance associated with the coil which limits the resonant impedance of the circuit. The value of the impedance of a parallel-tuned circuit at resonance is called the *dynamic resistance* of the circuit. This is a fictitious resistance and appears to exist only for alternating currents of the resonant frequency; the d.c. resistance of the circuit is, of course, relatively very low.

Parallel-tuned circuits are sometimes called *rejector circuits* because at resonance they have a high impedance and therefore reject current at their resonant frequency.

Dynamic Resistance

There is a series-to-parallel transformation which can be made between the two circuits of **Fig. 1.21 (C)** in which the series resistance r can be replaced by an equivalent shunt resistance R. Providing the value of r is small compared with the reactance of the coil, as is usually the case in radio circuits, the value of R is given by the formula—

$$R = \frac{X_L^2}{r} = \frac{\omega^2 L^2}{r}$$

where $\omega = 2\pi f$

It follows that the parallel-tuned circuit of Fig. 1.21 (A) can be replaced by that shown in **Fig. 1.21 (D)** in which a perfect coil and capacitor are shunted by a resistor R. At resonance the capacitive and inductive reactances cancel out and the impedance of the circuit reduces to the fictitious resistance R_D shown in **Fig. 1.21 (E)**: this is the dynamic resistance of the circuit.

Since at resonance $\omega L = 1/\omega C$, the value of the dynamic resistance of a parallel-tuned circuit is given by—

$$R_D = \frac{L}{Cr}$$

Further, since $Q = \omega L/r$, the dynamic resistance can also be expressed as—

$$R_D = \frac{Q}{\omega C}$$

From this it can be seen that for a high dynamic resistance (which is necessary for high gain in an r.f. amplifier when a

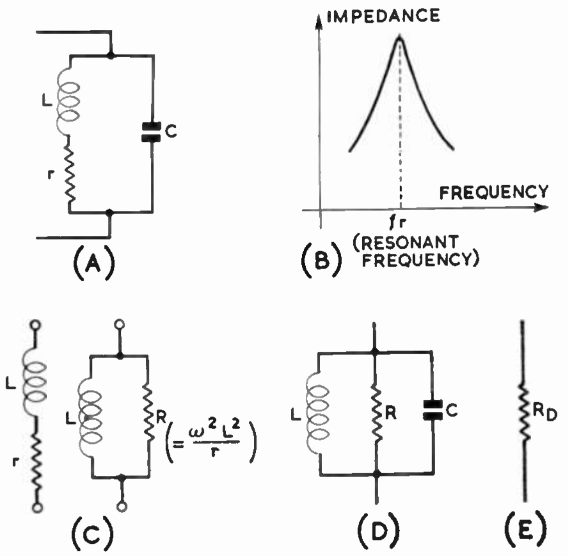


Fig. 1.21. The parallel-resonant circuit. The curve shown at (B) indicates how the impedance varies with frequency in the type of circuit shown at (A). A series-to-parallel transformation to replace the series resistance by an equivalent parallel resistance can be made as shown at (C). The circuit shown at (A) can therefore be replaced by that shown at (D) in which a perfect coil and capacitor are shunted by a parallel resistance. At resonance the impedance of this circuit reduces to a fictitious resistance as shown at (E).

parallel-resonant circuit is used as an anode load), the ratio of L to C should be high and r should be small (i.e., the Q should be high). The gain of a pentode radio-frequency or intermediate-frequency amplifier is roughly equal to the dynamic resistance of the tuned-circuit load multiplied by the mutual conductance of the valve: see Chapter 2 (*Valves*).

L/C Ratio

Examination of the formula for the resonant frequency of a tuned circuit, $f = \frac{1}{2\pi\sqrt{LC}}$ shows that the resonant frequency is determined by the product of the inductance and capacitance, LC. It follows that there is a wide range of L and C values which can be chosen to resonate at a particular frequency. Put another way, for a particular resonant frequency there is flexibility in choice of the L/C ratio, i.e. a large L and small C or, alternatively a small L and a large C may be used provided that the LC product gives the required resonant frequency.

The choice of L/C ratio is determined by practical considerations connected with the particular application of the tuned circuit.

In h.f. receivers it is usual to employ circuits with a high L/C ratio as this leads to circuits with a high dynamic resistance and therefore high stage gain, the minimum possible value of C being determined by the stray circuit capacitance. On the other hand, a low L/C ratio, i.e. a high C value, is usually employed in variable frequency oscillator circuits associated with transmitters in order to swamp out the effect of valve and other stray circuit capacitances which tend to make the frequency of the oscillator unstable.

In the case of h.f. transmitter output tank circuits the L/C ratio is specially chosen so that when damped by the load (or aerial) the effective Q of the tank circuit has a value which gives a good compromise between efficiency and harmonic suppression. This matter is dealt with in Chapter 6 (*H.F. Transmitters*).

Resonance Curves and Selectivity

The curves of Figs. 1.20 (B) and 1.21 (B) show how tuned circuits are more responsive at their resonant frequency than at neighbouring frequencies. Curves of this nature are called *resonance curves* and show how a tuned circuit can be used to pick out, or select, a wanted signal from a number of unwanted signals on other frequencies. **Fig. 1.22** shows a resonance curve A for a single parallel-tuned circuit having a resonant frequency of 465 kc/s and a Q of 100. When such a circuit is connected as the anode load in a simple single-stage intermediate-frequency amplifier circuit the curve shows how the gain varies with frequency. For instance, it will be seen that an unwanted signal at 461 kc/s (4 kc/s off tune) would give only about 50 per cent of the output of the wanted signal (of equal input amplitude) at 465 kc/s. The dashed curve B is for a similar circuit but with a Q of 300 and shows clearly how a higher Q gives greater selectivity.

There is a practical upper limit to the value of Q which can be obtained with a coil of reasonable size, and since much greater selectivity is required in a receiver than can be obtained with a single-tuned circuit several amplifying stages are commonly used in cascade, each with a resonant circuit as its anode load. The dotted curve C in Fig. 1.22 shows the response of three circuits in cascade, each with a Q of 100. The response at resonance has been rated at 100 per cent in each case regardless of the additional gain due to the extra valves. It will be seen that the curve is wider at the peak or " nose " than that of a single circuit with a Q of 300 (a desirable feature for telephony reception) and gives more selectivity down the " skirts " than the single circuit.

Another method of obtaining better selectivity than can be provided by a single tuned circuit is to apply *reaction*, or *positive feedback*, a method commonly used in straight receivers: see Chapter 4 (*H.F. Receivers*). The application of reaction artificially increases the Q of the circuit, a value of several thousand being readily obtainable, with a consequent increase in selectivity and sensitivity.

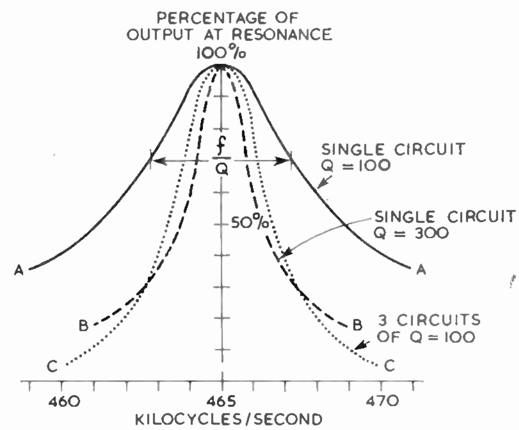


Fig. 1.22. Selectivity curves of single and cascaded tuned circuits.

TABLE 1.5

Selectivity of Tuned Circuits

Percentage of Output at Resonance	Bandwidth (width across response curve)
95%	$f/3Q$
90%	$f/2Q$
70%	f/Q
45%	$2f/Q$
24%	$4f/Q$
12%	$8f/Q$

(f is the resonant frequency of the circuit).

Table 1.5 is useful for deriving approximate selectivity curves from the Q of the circuit and the operating frequency. Thus, as an example, a circuit having a Q of 100 and a resonant frequency of 465 kc/s would have a bandwidth (i.e. width across the resonance curve) of $f/Q = 465/100 = 4{\cdot}65$ kc/s between points where the output had fallen to 70 per cent of the resonant value on either side of resonance, as marked in Fig. 1.22. The approximations given in Table 1.5 may be used with series-tuned circuits or with parallel-tuned circuits whose Q is greater than 10, as is the case in most radio equipment.

Coupled Circuits

For the satisfactory reception of radio telephony it is necessary to amplify equally the carrier frequency and a number of sideband frequencies which occur above and below the carrier frequency and which carry the speech intelligence: see Chapter 9 (*Modulation*). The ideal r.f. response curve would therefore be flat at the peak for a few kilocycles on either side of resonance to pass the sidebands and then fall away rapidly down the " skirts " to provide good adjacent-channel selectivity.

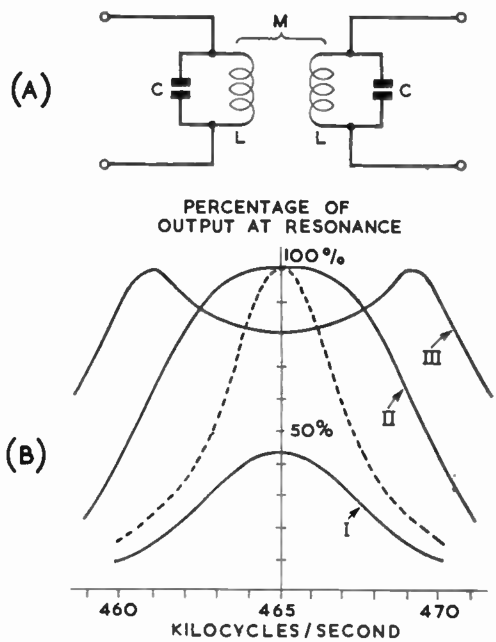


Fig. 1.23. Inductively-coupled tuned circuits. The curves shown at (B) represent the various frequency response characteristics of the coupled circuit shown at (A) for different degrees of coupling.

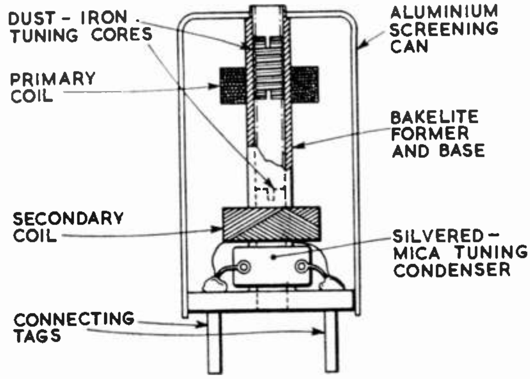


Fig. 1.24. A practical design for inductively-coupled tuned circuits.

This ideal response characteristic is more nearly obtained by using pairs of mutually-coupled tuned circuits than by the cascaded single-tuned circuits mentioned in the previous section.

The most common type of coupled circuit is that shown in **Fig. 1.23 (A)** in which inductive coupling is provided by the presence of mutual inductance between the identical primary and secondary circuits. The shape of the response curve of a pair of coupled circuits depends on the degree of coupling between them as well as on the Q of the parallel circuits. If there is little coupling (usually referred to as *loose coupling*) between the circuits, i.e. with the coils spaced far apart, the shape of the response curve is as for two cascaded circuits of the same Q and resonant frequency, but the output voltage would be small. This is shown in **Fig. 1.23 (B)**, curve I. If the degree of coupling is gradually increased by bringing the coils closer together the output at resonance will increase to a maximum value as shown in curve II. Further increase in coupling is accompanied by a reduction in output at the resonant frequency and the appearance of a double-humped response, curve III.

The degree of coupling corresponding to curve II is called *critical coupling*. For critical coupling the mutual inductance required between the coils is $1/Q$ of the inductance of either coil.

The dotted curve of Fig. 1.23 (B) is the overall response curve of two cascaded single circuits of the same Q as those in the coupled pair, and is included to illustrate the broader peak and steeper "skirts " of the latter.

Fig. 1.24 is a sketch showing a common method of mounting coupled tuned circuits. Two wave-wound coils are mounted on a common cylindrical former which is threaded internally to take a pair of dust-iron cores, one for tuning each coil. Fixed silvered-mica tuning capacitors are used and connections are brought out to soldering tags on the base. The whole unit is enclosed in an aluminium screening can. The degree of coupling between the coils is determined by their spacing, the correct coupling being found experimentally rather than by calculation. This is necessary because allowance must be made for a certain amount of unintentional capacitive coupling between the circuits which is additional to that provided by the mutual inductance.

The amount of coupling used in intermediate frequency transformers in superheterodyne receivers is usually less than the critical value since critically coupled and over-coupled transformers are more difficult to tune correctly.

Other methods of coupling tuned circuits are sometimes used. **Fig. 1.25 (A)** shows a pair of circuits with "top-capacity" coupling. The exact value of the coupling capacity C_c is found experimentally and a good starting value is $1/Q$ of the tuning capacitance C, where Q relates to either of the single-tuned circuits. In "bottom-capacity" or "common-capacity" coupling, shown in **Fig. 1.25 (B)**, the value required will be about Q times the tuning capacitance.

Fig. 1.25 (C) shows another way of employing mutual inductance known as *link coupling* which is useful if the two circuits are separated by an appreciable distance. Current in the primary coil induces a current in the link circuit which in turn induces a current in the secondary coil. The link coils usually have only a few turns of wire and sometimes merely a single turn.

Fig. 1.25 (D) shows a circuit in which the degree of coupling between primary and secondary may be varied by selecting the number of turns in the primary circuit which are arranged to link with the secondary by means of a switch. The switch is shown in the position of least coupling (narrowest bandwidth), and this arrangement may be adopted in a variable-bandwidth i.f. amplifier in a superheterodyne receiver.

FILTERS

Wave filters, usually referred to simply as filters, are networks of reactive components, i.e. capacitors and inductors, which exhibit certain characteristics as the frequency of a signal applied to them is varied. They are classed according to their frequency response and there are three types which are commonly used in Amateur Radio work, as follows:

(a) *Low-pass Filters* which have the characteristic of passing all frequencies below a specified frequency, called the cut-off frequency, fc, and attenuating all frequencies above fc.

(b) *High-pass Filters* which have the characteristic of passing all frequencies above a specified cut-off frequency, fc, and attenuating all frequencies below fc.

(c) *Band-pass Filters* which pass all frequencies between two specified cut-off frequencies f_1 and f_2 and attenuate all frequencies outside these two limits.

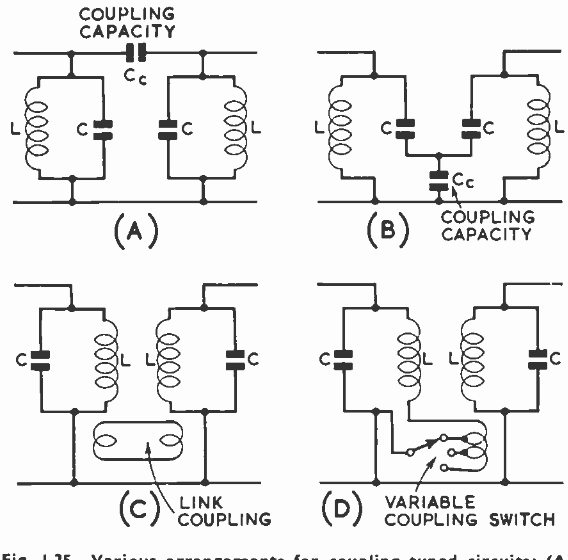


Fig. 1.25. Various arrangements for coupling tuned circuits: (A) "Top-capacity" coupling, (B) "Bottom-capacity" coupling, (C) link coupling, (D) variable inductive coupling, especially suitable for variable-bandwidth i.f. amplifier transformers.

Constant *k* Filters

The simplest type of filter is a constant k filter and examples of filter sections in this class of each of the three types mentioned above are shown in **Fig. 1.26** together with their frequency responses. Constant k filters have the characteristic of having negligible loss in the passband and have a steadily increasing attenuation outside the passband.

The solid line frequency responses of Fig. 1.26 are for ideal filters and show a sudden onset of attenuation at the cut-off frequency. In practical filters, due to losses in the reactive components and the effect of the load resistors terminating the filter, the transition from passband to stopband is rounded as indicated by the dotted lines.

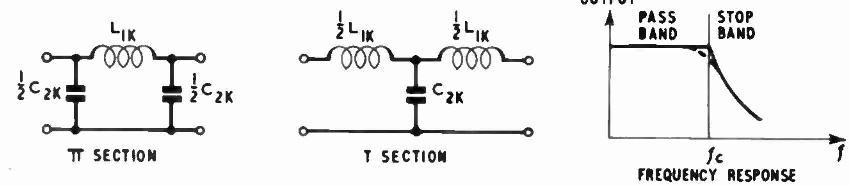
Two types of constant k filter sections shown in Fig. 1.26 are called π sections and T sections because of their circuit configurations—π or T sections may be chosen to make up a filter, the choice being governed by economic considerations and the impedance characteristics desired looking into the filter in the stopband remote from the cut-off frequency. For example, it is usual to make a low-pass filter of π sections and a high-pass filter of T sections, because the number of coils, which are expensive relative to the capacitors, is smaller.

A single section filter will give a certain attenuation outside the passband; for greater attenuation several similar sections may be connected in cascade. Curve A of **Fig. 1.27** shows the theoretical frequency response of an ideal constant k filter section but a practical filter will have a rounded response instead of the sharp corner at the cut-off frequency. The curve shown is useful, however, in assessing how many filter sections will be needed to obtain some specified attenuation at some frequency away from cut-off.

When constructing filters it is important to use coils of high Q in order to minimize loss in the passband. Care should be taken to make sure that there is no stray magnetic coupling between the coils in a filter. In radio frequency filters it is particularly important to see that there is no stray capacitance coupling between sections and for this reason it is usual to mount different sections each in a screened compartment.

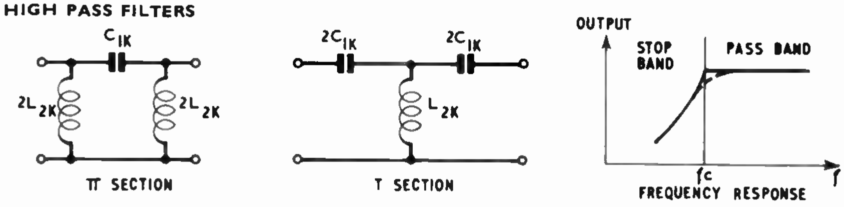
Filters are designed to have a required cut-off frequency f_c and also to operate between a desired generator and load impedance R_o. The design formulae of Fig. 1.26 allow the coil and capacitor values, L_{1K}, C_{1K}, etc., of what are called the prototype constant k sections, to be calculated for chosen f_c and R_o. The actual values to be used in a particular filter configuration are then obtained by substituting these values as shown against the components in the circuit

LOW PASS FILTERS



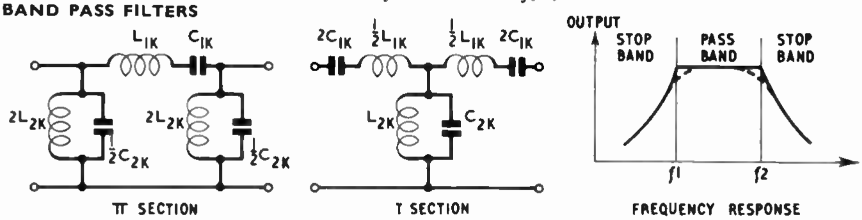
$$L_{1K} = \frac{R_o}{\pi f_c}\ ;\ C_{2K} = \frac{1}{\pi f_c R_o}$$

HIGH PASS FILTERS



$$L_{2K} = \frac{R_o}{4\pi f_c}\ ;\ C_{1K} = \frac{1}{4\pi f_c R_o}$$

BAND PASS FILTERS



$$L_{1K} = \frac{R}{\pi(f_2 - f_1)}\ ;\ C_{1K} = \frac{(f_2 - f_1)}{4\pi R f_1 f_2}$$

$$L_{2K} = \frac{R(f_2 - f_1)}{4\pi f_1 f_2}\ ;\ C_{2K} = \frac{1}{\pi R(f_2 - f_1)}$$

Fig. 1.26. Design formulae for constant k filters.

diagrams. If, after calculation, impracticably large inductors or small capacitors are called for, the filter should be re-designed for a different value of R_o.

M-derived Filters

By modifying the constant k filter sections it is possible to obtain a frequency response which falls off more rapidly beyond the cut-off frequency f_c and at a particular frequency in the stop-band, f_∞, a theoretically infinite attenuation can be obtained. Such filters are called m-derived filters and some of the commonly used types are shown in **Fig. 1.28** together with their frequency responses and design formulae. When using the m-derived filter design formulae first calculate the values of C_{1K}, L_{1K}, etc., as for the prototype constant k filter (Fig. 1.26) and then find the values of $L1$, $C1$, etc. Finally substitute the value of $L1$, $C1$, etc., shown against the components in the circuit diagrams.

The value of m is related to the ratio of the cut-off frequency f_c and the frequency of infinite attenuation f_∞. For example if $f_\infty/f_c = 1{\cdot}25$ for a low-pass filter, then $m = 0{\cdot}6$ (see Fig. 1.28). Curve B of Fig. 1.27 shows the theoretical frequency response curve of m derived low-pass and high-pass filter sections for a value of $m = 0{\cdot}6$. The difference in shape of the response curves of constant k (Curve A) and m derived (Curve B) filter sections is clearly shown.

M-derived sections are not usually used alone because the attenuation drops beyond the frequency f_∞. To overcome this difficulty, practical filters use a combination of m derived and constant k sections connected in series.

M-derived End Sections

An m-derived filter section, with $m = 0{\cdot}6$, has the property of matching well to a resistive load over the passband, giving therein a level frequency response and low attenuation. It is common practice to use m-derived half-sections, with $m = 0{\cdot}6$, at each end of a filter in order to obtain this good match. Suitable end half-sections for use with π section low-pass filters and T section high-pass filters are shown in Fig. 1.28.

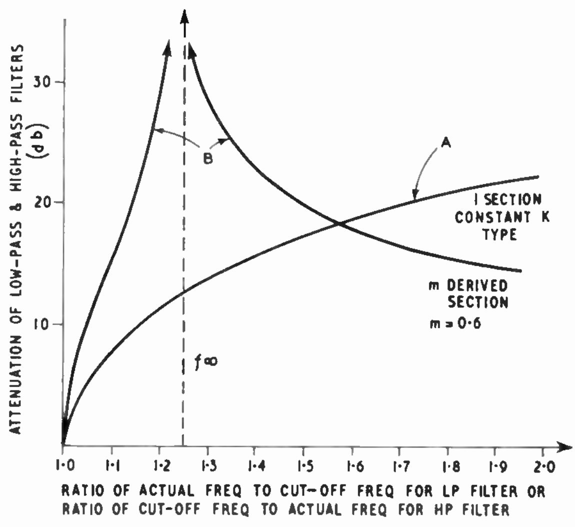
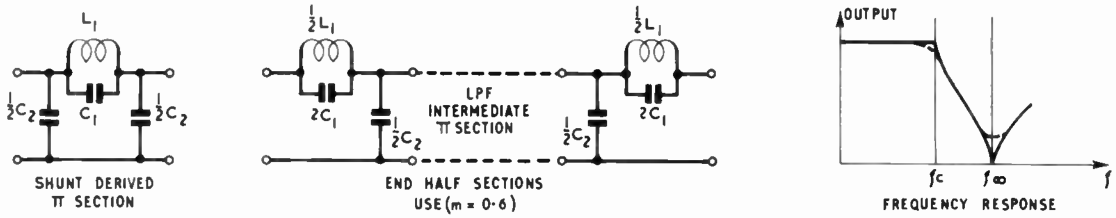


Fig. 1.27. Frequency characteristics of low-pass and high-pass filter sections. Curve A is for constant k and Curve B for m-derived ($m = 0{\cdot}6$) types.

LOW PASS FILTERS

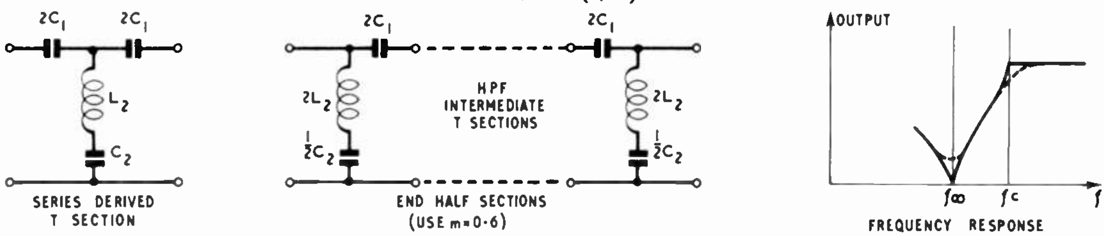
$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2}$$



$$L_1 = m L_{1K}\,;\quad C_1 = \frac{1 - m^2}{4m} C_{2K}\,;\quad C_2 = m C_{2K}.$$

HIGH PASS FILTERS

$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2}$$



$$C_1 = \frac{C_{1K}}{m}\,;\quad C_2 = \frac{4m}{1 - m^2} C_{1K}\,;\quad L_2 = \frac{L_{2K}}{m}$$

Fig. 1.28. Design formulae for m-derived filters.

TRANSFORMERS

The fact that a changing current in one circuit can induce a current in a second circuit, as in Fig. 1.14 (B), is the basis of transformer action.

Transformers are useful for transferring electrical energy from one circuit to another without direct connection; for example from the anode circuit of one valve to the grid circuit of another valve. In the transfer process it is possible to change the relative voltages and impedances of the primary and secondary circuits, examples being the supply of a low voltage to operate a valve heater from the high-voltage supply mains and the impedance matching of a low-resistance loudspeaker to a valve anode circuit.

The coupled circuits referred to in the discussion on Fig. 1.23 are examples of transformers with very loose coupling between the primary and secondary windings. In audio-frequency and power supply transformers, tight coupling is required, and the primary and secondary windings are therefore wound on an iron core with a construction similar to that of the low-frequency choke of Fig. 1.15 (A-C).

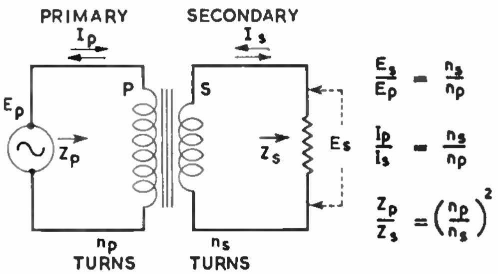


Fig. 1.29. The low-frequency transformer.

The size of core used in the transformer depends on the amount of power to be handled.

Fig. 1.29 shows a simple transformer with a primary winding *P* and a secondary winding *S*. Since both windings are in the same alternating magnetic field, the induced voltages will be in proportion to the number of turns on each coil.

Thus if—

Number of turns on the primary $= n_p$
Number of turns on the secondary $= n_s$
Voltage across the primary $= E_p$
Voltage across the secondary $= E_s$

$$E_s = E_p \frac{n_s}{n_p}$$

The ratio n_s/n_p is called the *turns ratio* and a transformer may step a voltage up or down according to whether n_s/n_p is greater or less than unity.

As an example consider a transformer which supplies valve heaters having 1200 turns in its primary and 32 turns in its secondary. When connected to a 240 volt mains supply the secondary voltage will be—

$$E_s = \frac{240 \times 32}{1200}$$
$$= 6{\cdot}4 \text{ volts.}$$

As long as there is no load connected to a transformer the primary current should be very small. This current is called the *magnetizing current* and can be neglected in most calculations when compared with the primary current due to the secondary load. Since the intensity of the magnetic field set up by the primary and secondary current is the same and since the field intensity is proportional to the number of ampere-turns of each winding it follows that the primary current will be equal to the secondary current multiplied by the turns ratio.

Thus if—

Current in the primary $= I_p$
Current in the secondary $= I_s$

$$I_p = I_s \frac{n_s}{n_p}$$

Taking the previous example again, if the heater current is 6 amperes the primary current would be—

$$I_p = \frac{6 \times 32}{1200} \text{ amperes} = 160 \text{ mA.}$$

A transformer has the property of being able to transform impedances. If the impedance offered by the primary of the transformer of Fig. 1.29 is measured it will be found to be equal to the secondary or load impedance divided by the square of the turns ratio. Thus if—

Primary impedance $= Z_p$
Secondary impedance $= Z_s$

$$Z_p = Z_s \left(\frac{n_p}{n_s}\right)^2$$

It is probably more convenient to remember this result as—

$$\frac{Z_p}{Z_s} = \left(\frac{n_p}{n_s}\right)^2$$

which means that *the impedance ratio is equal to the square of the turns ratio.*

The transformation of impedance is a valuable property and transformers are widely used for matching a load to a source of power, the turns ratio being calculated to give maximum transfer of power to the load.

For example, the load into which an audio frequency amplifier valve delivers its maximum undistorted output may be 6400 ohms. However, the loudspeaker which it is desired to use with the valve may have an average impedance of only 4 ohms and a matching transformer will therefore be required. The transformer turns ratio can be calculated from the formula just mentioned by taking the square root of the impedance ratio, thus—

$$\frac{n_p}{n_s} = \sqrt{\frac{Z_p}{Z_s}} = \sqrt{\frac{6400}{4}} = \frac{40}{1}$$

This means that a transformer with 40 times as many turns on the primary as on the secondary will give the required impedance match.

Transformers are often referred to by their primary-to-secondary winding turns ratio. Thus a 40-to-1 (or 40 : 1) transformer would have 40 turns on the primary for each turn on the secondary, and would be a " step-down " transformer suitable for matching a primary-to-secondary impedance ratio of $(40)^2$: 1 or 1600 : 1. A transformer with a 1 : 2 ratio would have two turns on the secondary for each turn on the primary, and would be a step-up transformer. It would match a primary-to-secondary impedance ratio of 1 : 4.

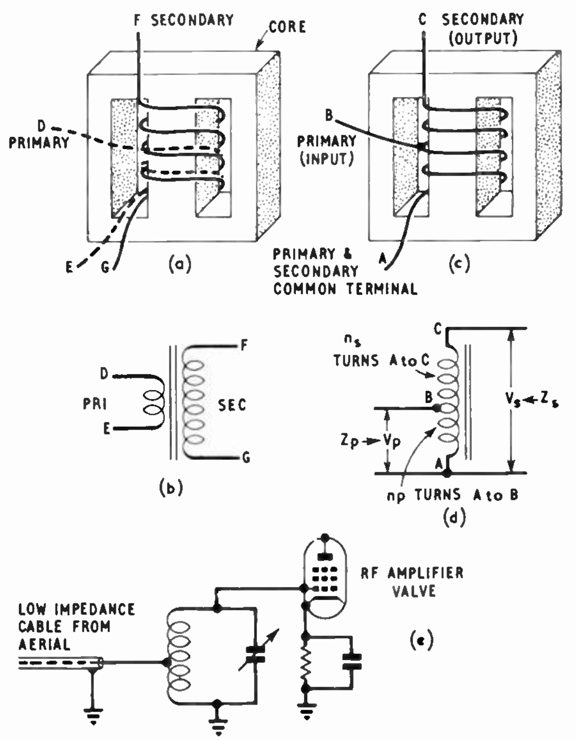


Fig. 1.30. Derivation of the autotransformer is shown in (a), (b) and (c). The circuit symbol is shown in (d) and a typical use at radio frequency in (e).

Auto-transformers

It is not always necessary to have d.c. isolation between the primary and secondary circuits of a transformer and in this case an *auto-transformer* can be used.

An auto-transformer can be looked upon as an ordinary transformer in which the primary or secondary winding is common to the other. **Fig. 1.30 (A)** shows the magnetic core of an ordinary step-up transformer with a two turn primary, shown dotted, and a four turn secondary. The primary is shown lying alongside the corresponding turns of the secondary; such a transformer would have a voltage step-up of 2 : 1 and its circuit symbol is as shown in **Fig. 1.30 (B).**

In **Fig. 1.30 (C)** the two turn primary has been merged (or made common) with the secondary to form an auto-transformer. The two turns between the common point A and the tap B form the primary and the whole four turns from A to C form the secondary. The circuit symbol for such an auto-transformer is shown in **Fig. 1.30 (D).**

The voltage ratio of an auto-transformer (Fig. 1.30 (D)) is calculated from the ratio of the number of turns n_p between the common point A and the tap B to the total number of turns between A and C, thus—

$$\frac{V_p}{V_s} = \frac{n_p}{n_s}$$

As with ordinary transformers, auto-transformers can be used to provide impedance transformation or matching, the impedance ratio being equal to the square of the turns ratio. Thus in Fig. 1.30 (D)—

$$\frac{Z_p}{Z_s} = \left(\frac{n_p}{n_s}\right)^2$$

Auto-transformers can be used to step up or down depending on whether the tap is the primary or secondary connection.

For a.c. power work, auto-transformers have the advantage that the windings and their insulation take up less space than in an ordinary transformer. They are commonly used to step-down or boost an a.c. supply voltage to some other value. For example a step-down auto-transformer, with a suitable tap, could be used to run 110 volt a.c. equipment from 240 volt a.c. mains.

The tapped tuned circuit is an example of a tuned auto-transformer which is often used at radio frequency. A typical application is shown in **Fig. 1.30 (E)** in which a tapped tuned circuit is used to match and provide a voltage step-up between a low impedance aerial feeder and the input of a r.f. amplifier valve.

SCREENING

When two circuits are near one another, unwanted coupling may exist between them due to stray capacitance between them or due to stray magnetic coupling.

Stray capacitance coupling can be eliminated by placing an earthed screen of good conductivity between the two circuits in question as shown in **Fig. 1.31 (B)**. There is then only stray capacitance from each circuit to earth and no direct capacitance between them. A useful practical rule is to position screens so that the two circuits are not visible from one another.

Stray magnetic coupling can occur between coils and wires due to the magnetic field of one coil or wire embracing the other. At radio frequency, coils can be screened by placing them in closed boxes or cans made from material of high conductivity such as copper, brass or aluminium. In practice, eddy currents are induced in the screening can which in turn set up a field which opposes and practically cancels the field due to the coil beyond the confines of the can.

If a screening can is too close to a coil the performance of the coil, i.e. its Q and also its inductance, will be considerably reduced. A useful working rule is to use a screening can of such a size that nowhere is it nearer to the coil than a distance equal to the diameter of the coil; see Fig. 1.31 (C).

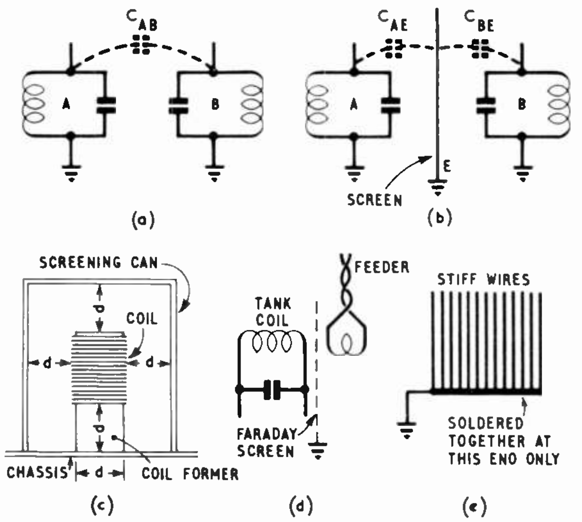


Fig. 1.31. (a) Stray capacity coupling C_{AB} between two circuits A and B. The introduction of an earthed screen E in (b) eliminates direct capacity coupling, there being now only stray capacitance to earth from each circuit C_{AE} and C_{BE}. A screening can (c) should be of such dimensions that it is nowhere nearer to the coil it contains than a distance equal to the diameter of the coil d. A Faraday screen between two circuits (d) allows magnetic coupling between them but eliminates stray capacitance coupling. The Faraday screen is made of wires as shown at (e).

At low frequencies where screening due to eddy currents is not so effective it may be necessary to enclose a low frequency choke or transformer in a box of high permeability magnetic material such as Mumetal in order to obtain satisfactory magnetic screening. Such measures are not often required but a sensitive component such as a microphone transformer may be enclosed in such a screen in order to make it immune from hum pick-up.

It is sometimes desirable to have pure inductive coupling between two circuits with no stray capacitance coupling. In this case a *Faraday Screen* can be employed between the two coils in question as shown in **Fig. 1.31 (D)**. This arrangement is sometimes used between an aerial and a receiver input circuit or between a transmitter tank circuit and an aerial. The Faraday Screen is made of stiff wires (Fig. 1.31 (e)) connected together at one end only, rather like a comb. The screen is transparent to magnetic fields because by its very construction there is no continuous conducting surface in which eddy currents can flow and thereby give magnetic screening. However, because the screen is connected to earth it acts very effectively as an electrostatic screen eliminating stray capacitance coupling between the circuits.

UNBALANCED AND BALANCED CIRCUITS

Unbalanced Circuits

An unbalanced circuit is shown in **Fig. 1.32 (A)**. A generator, or signal source is shown connected to a load resistor through a single conductor; the return path for current is through what is termed an *earth return*. In practice, although the term an earth return is used, the return path is usually through the chassis of the equipment.

Unbalanced circuits of this type are very commonly used in radio equipment and are perfectly satisfactory provided leads are kept short and are spaced well away from other leads. The circuit is, however, prone to the pick-up of extraneous noise and signals from neighbouring circuits by three means: Inductive pick-up, Capacitive pick-up and through a common earth return path.

Inductive pick-up, **Fig. 1.32 (B)**, can take place due to transformer action between the unbalanced circuit wire and another nearby wire carrying an alternating current; an example is hum pick-up in the grid circuit wiring of an amplifier due to a.c. in the heater wiring.

Capacitive pick-up, **Fig. 1.32 (C)**, takes place through the stray capacitance between the unbalanced circuit lead and a neighbouring wire. Such pick-up can usually be eliminated by introducing an earthed metal screen around the connecting wire.

If the unbalanced circuit has an earth return path which is common to another circuit, **Fig. 1.32 (D)**, unwanted signals or noise may be injected into the unbalanced circuit. The extraneous signal current, flowing in the common earth return path, can cause a voltage to appear between the two earth return points of the unbalanced circuit. This voltage will, in turn, cause an unwanted current to flow around the unbalanced circuit. Interference of this type can be minimized by using a low resistance chassis and avoiding common earth paths as far as possible.

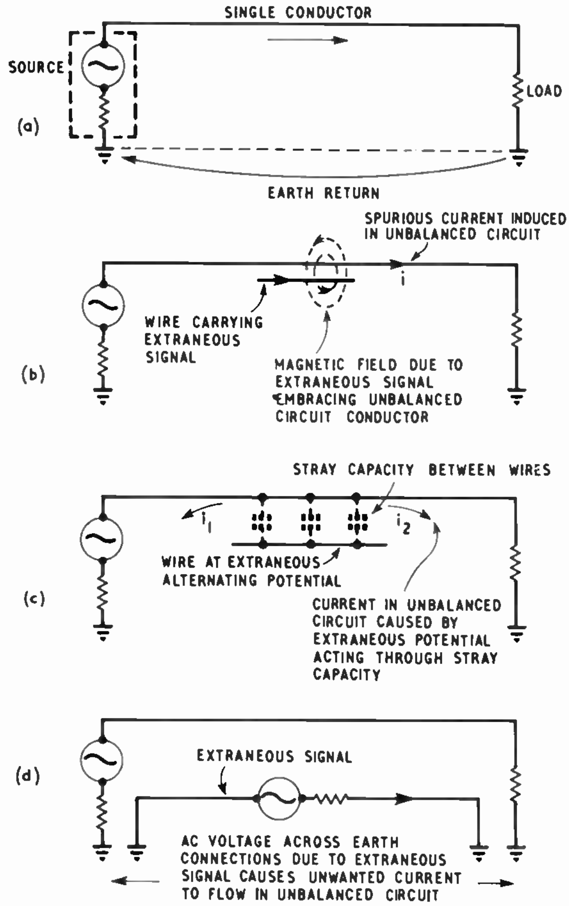


Fig. 1.32. The unbalanced circuit. (a) The basic unbalanced circuit showing earth return path. (b) How extraneous signals and noise can be induced in an unbalanced circuit by magnetic induction. (c) Showing how extraneous signals can be induced by stray capacitance coupling. (d) Diagrams showing how extraneous signals can be induced due to a common earth return path.

Balanced Circuits

A balanced circuit is shown in **Fig. 1.33.** As many signal sources, and often loads as well, are inherently unbalanced (i.e. one side is earthed) it is usual to use transformers to connect a source of signal to a remote load via a balanced circuit. In the balanced circuit, separate wires are used to conduct current to and back from the load; no current passes through a chassis or earth return path.

The circuit is said to be balanced because the impedances from each of the pair of connecting wires to earth are equal. It is usual to use twisted wire between the two transformers as shown in Fig. 1.33. For a high degree of balance, and therefore immunity to extraneous noise and signals, transformers with an earthed screen between primary and secondary windings are used; sometimes the centre taps of the balanced sides of the transformers are earthed as shown dotted in Fig. 1.33.

The balanced circuit overcomes the three disadvantages of the unbalanced circuit. Inductive and capacitive pick-up are eliminated since equal and opposite currents are induced in each of the two wires of the balanced circuit and these cancel out. The same applies to interfering currents in the common earth connection in the case where the centre taps of the windings are earthed.

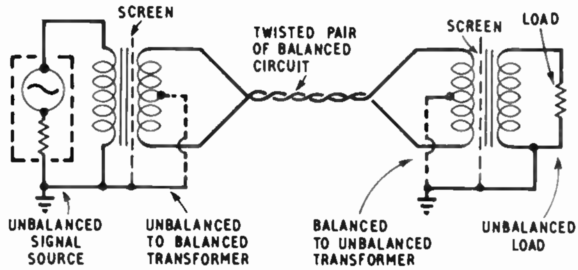


Fig. 1.33. The balanced circuit.

THE ELECTROMAGNETIC SPECTRUM

Radio waves are electromagnetic waves forming part of the electromagnetic spectrum which comprises radio, heat, light, ultra-violet, gamma rays, and X-rays: see **Fig. 1.34**. The various forms of electromagnetic radiation are all in the form of oscillatory waves, but differ from each other in frequency and wavelength. Notwithstanding these differences they all travel through space with the same speed, namely, 3×10^{10} cm per second. This is equivalent to about 186,000 miles per second (i.e., once round the world in one-seventh of a second).

Frequency and Wavelength

The distance travelled by a wave in the time taken to complete one cycle of oscillation is called the *wavelength*. It follows that wavelength, frequency and velocity of propagation are related by the formula—

$$\textit{Velocity} = \textit{Frequency} \times \textit{Wavelength}$$

or $c = f.\lambda$

where c = velocity of propagation (3×10^{10} cm/sec.)

f = frequency of oscillation (c/s)

λ = wavelength (cm).

This formula enables the wavelength to be calculated if its frequency is known, and vice versa.

The following example is taken from a Radio Amateurs' Examination paper.

Example. *What are the frequencies corresponding to wavelengths of (i) 150m, (ii) 2m and (iii) 75cm.*

From the formula $c = f\lambda$, the frequency is given by—

$$f = \frac{c}{\lambda}$$

Case (i): $\lambda = 150\text{m} = 1{\cdot}5 \times 10^4$ cm
$c = 3 \times 10^{10}$ cm/sec.

$$\text{Therefore } f = \frac{3 \times 10^{10}}{1{\cdot}5 \times 10^4} = 2 \times 10^6 \text{ c/s}$$

$$= 2{\cdot}0 \text{ Mc/s}$$

Case (ii): $\lambda = 2\text{m} = 2 \times 10^2$ cm

$$\text{Therefore } f = \frac{3 \times 10^{10}}{2 \times 10^2} = 1{\cdot}5 \times 10^8 \text{ c/s}$$

$$= 150 \text{ Mc/s}$$

Case (iii): $\lambda = 75$cm

$$\text{Therefore } f = \frac{3 \times 10^{10}}{75} = 4 \times 10^8 \text{ c/s}$$

$$= 400 \text{ Mc/s.}$$

Fig. 1.34 shows how the radio spectrum may be divided up into various bands of frequencies, the properties of each

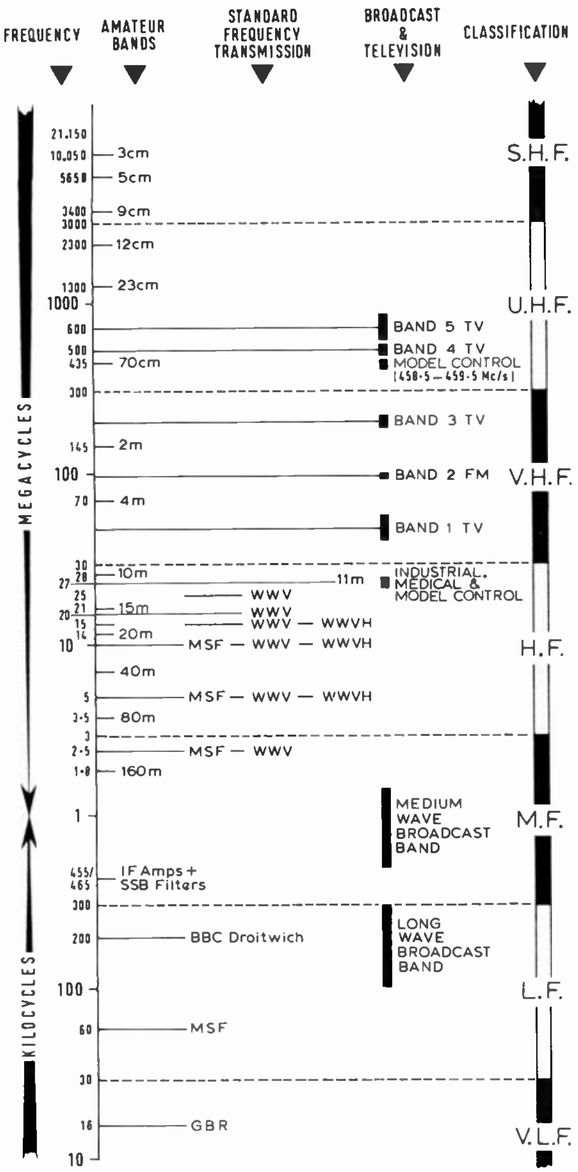


Fig. 1.34. The radio-frequency spectrum.

making them suitable for specific purposes. Amateur transmission is permitted on certain frequency bands in the h.f., v.h.f. and u.h.f. ranges.

AERIALS

An aerial is used to launch electromagnetic waves into space or conversely to pick up energy from such a wave travelling through space. Any wire carrying a high frequency alternating current will radiate electromagnetic waves and conversely an electromagnetic wave will induce a voltage in a length of wire. The problem in aerial design is to radiate as much transmitter power as possible in the required direction, or in the case of a receiver, to pick up as strong a signal as possible, very often in the presence of local interference.

An aerial may be considered as a tuned circuit consisting of inductance, capacitance and resistance, and for maximum radiation an aerial is usually operated so that it is naturally resonant at the operating frequency, maximum radiation corresponding to maximum high-frequency current flowing in the aerial.

There are two basic types of aerial used by amateurs, (*a*) the Hertzian aerial, known as a *dipole* or *doublet*, which is self-resonant by virtue of its length in relation to the wavelength of the signal, and (*b*) the Marconi aerial which may be of any convenient length and relies on an earth connection and is tuned to resonance by means of a coil or capacitor. The simple Hertzian aerial is one half-wavelength long and is therefore well suited to the shorter waves (higher frequencies) whereas the Marconi aerial is mainly used at longer wavelengths, e.g., 160m and 80m where it is often impracticable to erect a Hertzian dipole. More detailed explanations of aerial theory and practical design will be found in Chapters 13 and 14. A typical examination question on the tuning of a Marconi aerial is worked out in Example III on page 1.17.

PROPAGATION OF RADIO WAVES

Radio waves, on leaving an aerial, travel through space in straight lines with the velocity of light. The waves consist of interdependent electric and magnetic fields which act in directions mutually at right angles and also at right angles to the direction of propagation of the wave.

If the electric field acts in a vertical direction the wave is said to be *vertically polarized* and such waves are launched and best picked up by vertical aerials, **Fig. 1.35 (A)**. If the electric field is horizontal the waves are said to be *horizontally polarized* and horizontal aerials are then used: see **Fig. 1.35 (B)**.

Horizontal polarization is most commonly used by amateurs on the h.f. bands for long-distance communication. Vertical polarization is frequently used for mobile work and local work on v.h.f. and u.h.f. bands where the "all-round" or omni-directional characteristics of simple vertical aerials are an advantage.

The propagation of radio waves to remote parts of the earth depends on reflection of the waves from layers of ionized gas situated between 70 and 200 miles or so above the earth in a region called the *ionosphere*. As the reflecting properties of this region depend on the frequency used and are continually changing with night and day, the time of year, and sunspot activity, it can be imagined that the

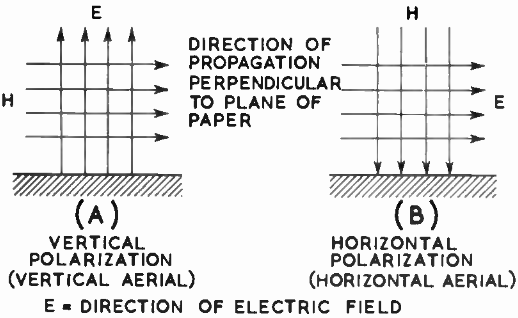


Fig. 1.35. Diagram illustrating the polarization of electromagnetic waves: (A) represents vertical polarization and (B) represents horizontal polarization, the distinction being in terms of the electric vector rather than the magnetic vector.

mechanism of long-distance short-wave propagation is complex and very variable. However, transmission conditions to distant points can be forecast with some accuracy. The propagation of v.h.f. and u.h.f. signals rarely depends on the conditions in the ionosphere but on meteorological conditions in the lower levels of the atmosphere called the *troposphere*.

The subject of radio-wave propagation is treated in detail in Chapter 12 (*Propagation*).

Fading

It is common for h.f. signals which have arrived at a receiver from a distant transmitter to vary continually in strength. This phenomenon is known as *fading* and is caused by the varying relationship between signals arriving at the receiver by different paths whose relative lengths are changing.

The strength of the signal at the receiving aerial terminal is the sum of signals arriving from a transmitter by perhaps two different paths. If the two paths differ in length by an exact number of wavelengths, corresponding to the frequencies in use the signals arriving by the two paths will arrive in phase and the total strength of received signal will be a maximum. If, however, the path lengths differ by an odd number of half-wavelengths the two signals will be out of phase and the total received signal strength will be a minimum. Thus if a signal is being received both by ground-wave and sky-wave paths, or by two different sky-wave paths, fading will be experienced when the effective height of the ionized layer which reflects the sky-wave varies since this has the effect of altering the relative path lengths between transmitter and receiver.

Fading may be slow or very rapid: if it is rapid, the fading appears as an audible low-frequency flutter. During fading, speech transmitted by a.m. sometimes becomes badly distorted because the carrier and various sideband frequencies do not all fade equally at the same time.

The effect of fading is combated by applying automatic volume control (better referred to as *automatic gain control*) to receivers so that the overall gain of the set varies inversely with the strength of the received signal. Other methods used are (*a*) *highly directive aerial systems*, particularly in the vertical plane to restrict the possibility of multi-path transmission, and (*b*) *diversity reception* in which two or more spaced aerials or two or more different frequencies are used for a transmission. Diversity reception depends for its operation on the fact that, statistically, fading is unlikely to occur at two different points or on two different frequencies at the same time. Two receivers are used with an automatic change-over arrangement for selecting the stronger of the two signals, a considerable reduction in fading being achieved.

VALVES

The successful development of radio communication has depended overwhelmingly on the thermionic valve. Valves are used as rectifiers for converting alternating current into direct current both at power-supply frequencies and at radio frequency where the process is more commonly called *detection*. Other common applications are the generation of radio frequency power in transmitters and the amplification of signals in receivers.

The earliest valves, called *diodes*, had two electrodes enclosed in an evacuated envelope. This is shown diagrammatically in **Fig. 1.36** (A). The filament electrode or *cathode*, was made of tungsten wire and could be heated to incandescence by the passage of a current through it supplied by a low-tension or " A " battery. When incandescent the filament has the property of emitting free electrons which, being negatively charged, may be attracted to the second electrode called an *anode* or *plate* if the voltage of the anode is made positive with respect to the filament by the use of a high-tension (h.t.) battery " B," as shown in **Fig. 1.36 (B)**. Reversal of the polarity between anode and cathode, as in **Fig. 1.36** (C), will repel the electrons and prevent the passage of current. The actual flow of current through a valve is in the form of a flow of electrons from filament to anode; nevertheless the conventional flow of current from the positive to the negative battery terminal, i.e., from anode to cathode, is usually assumed.

More efficient emitter materials are now used in valves than the original tungsten filament. For example, thoriated tungsten (a mixture of tungsten and 1–2 per cent of thorium) is up to ten times as efficient as pure tungsten. An even more

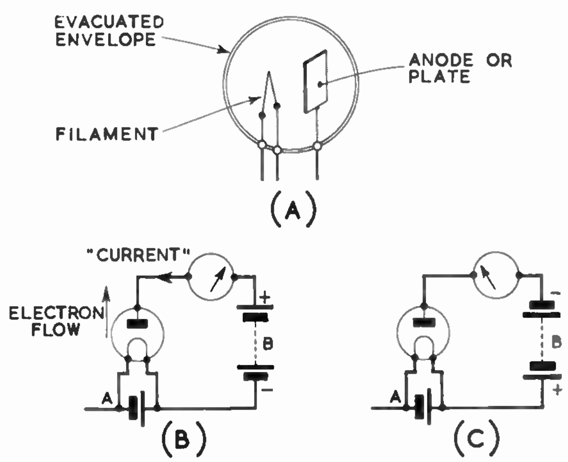


Fig. 1.36. A simple diode valve: (A) shows the essential construction comprising a filament and an anode: (B) illustrates how current flows through the valve when the anode is made positive with respect to the filament, and (C) shows the cessation of current when the polarity is reversed.

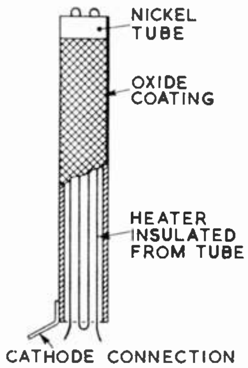


Fig. 1.37. Partly sectioned view of an indirectly-heated cathode.

efficient emitter very commonly used in valves is the oxide-coated type in which a mixture of barium, strontium and calcium oxides forms a coating on a filament. Such a composite filament gives a copious emission at relatively low temperature.

Valves using a filament as a source of electrons are called *directly-heated* valves. It is an advantage, particularly in equipment operating on a.c. supplies, to use valves of the types having what is called an *indirectly-heated cathode*. An indirectly-heated cathode consists of a nickel tube coated with the rare-earth oxides mentioned above to form the emitting surface which is heated to the required temperature by a filament or heater inside the tube but insulated from it as shown in the sketch in Fig. 1.37.

The introduction of an extra electrode, called a *grid*, between the cathode and anode forms a three-electrode valve which is called a *triode* and is shown diagrammatically in Fig. 1.38 (A). The grid has the property of controlling the flow of electrons from the cathode to the anode, and the more negative the grid voltage is relative to the cathode, the smaller the anode current becomes until it is finally cut off altogether. A small change in grid voltage may cause a considerable change in anode current and this effect forms the basis of the valve amplifier.

Fig. 1.38 (B) shows the construction of an indirectly-heated triode valve. The cathode, grid and anode are mounted between mica supports and are connected to appropriate pins fitted in a glass base. The electrode assembly is surrounded by a glass bulb sealed to the base, and at the top of the bulb is the seal through which the air is exhausted during manufacture. The purpose of the *getter*, which is fired during the evacuation process, is to deposit a gas-absorbing film of barium metal on the inside of the bulb. It remains active after manufacture and maintains a very high vacuum.

More complex valves using more electrodes than the triode are in common use. Detailed information on all types of valve and their applications will be found in Chapter 2 (*Valves*).

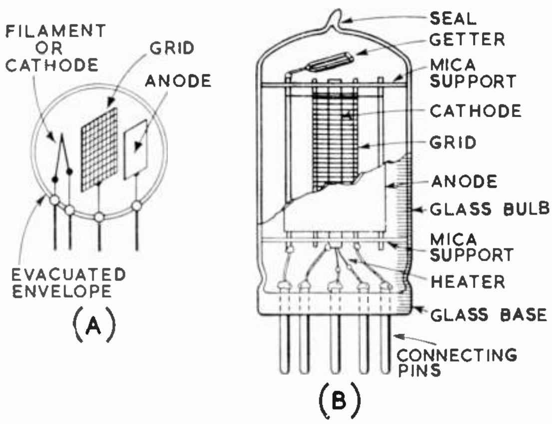


Fig. 1.38. The triode valve: (A) indicates the basic arrangement of the electrodes, the grid being interposed between the cathode and the anode, and (B) shows the construction of a typical indirectly-heated triode valve.

Rectification and Detection

The operation of a diode as a rectifier is illustrated in Fig. 1.39. The circuit shows a mains transformer with two secondary windings, one to heat the cathode of the rectifier valve and the other to step the mains voltage up or down according to the output voltage required. The diode will conduct only during the positive half-cycle of the a.c. waveform V_a applied to its anode so that the voltage V_L across the load resistance (and the current through it) will be uni-directional as shown. Practical applications of this circuit to power supplies will be found in Chapter 17 (*Power Supplies*).

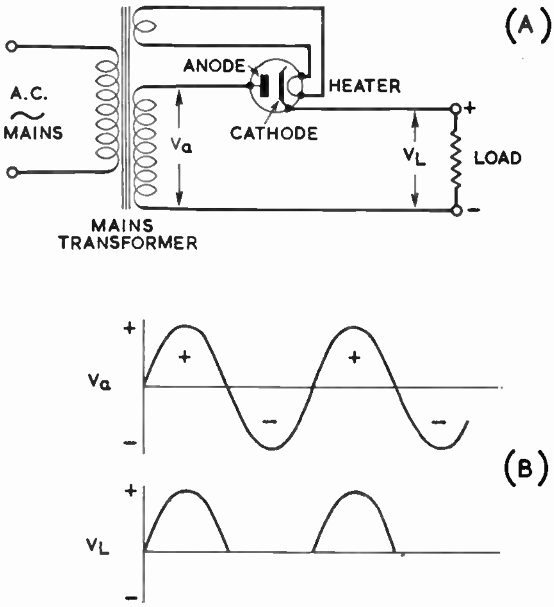


Fig. 1.39. The operation of the diode as a rectifier: (A) shows an elementary circuit in which an a.c. supply is rectified, producing uni-directional current pulses in the resistive load, (B) represents the waveform of the a.c. input voltage and the rectified voltage across the load.

The uni-directional conduction of a diode is used in a similar way but at a higher frequency to rectify or "detect" radio-frequency signals, the most common applications being the second detector and the a.g.c. rectifier in superheterodyne receivers.

Amplification and Oscillation

The basic amplifier circuit is shown in Fig. 1.40. The grid

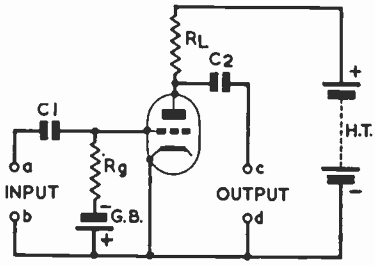


Fig. 1.40. Basic amplifier circuit (untuned).

is biased slightly negative with respect to the cathode by the bias battery *GB* in series with the high grid resistance R_g (usually 0·1–1 Megohm). A steady anode current will flow through the valve and the load resistance R_L.

An a.c. signal applied between the input terminals *a* and *b* will be applied through the coupling capacitor *C*1 to cause the grid-to-cathode voltage to vary above and below its steady bias voltage in sympathy with the a.c. waveform applied. The changing grid-to-cathode voltage will cause a corresponding changing anode current to flow through the load resistance R_L and the anode-to-cathode path in the valve, this variation being superimposed on the steady anode current. The alternating component of the anode current will produce an amplified replica of the input waveform at the output terminals *c* and *d*. It should be noted that the amplified output waveform will be 180° out-of-phase with the input waveform since a positive-going grid voltage causes an increase in anode current which in turn causes a decrease in voltage between the anode and the cathode.

The circuit shown is applicable to audio frequency amplification. At radio frequencies a frequency-selective amplifier is usually required and selectivity is obtained by using a parallel-tuned circuit, *LC*, as an anode load instead of the resistor R_L shown in **Fig. 1.41**. Except with special constructional or circuit arrangements triodes are not suitable for use as r.f. amplifiers since the unavoidable capacitance between grid and anode inside the valve results in unwanted oscillation. This effect is avoided by the use of a pentode valve, as shown in this circuit. Since the output voltage from such an amplifier is, within limits, proportional to the load impedance it follows that the output voltage will be a maximum at the resonant frequency of the tuned circuit where its impedance is a maximum, falling away on either side of the resonance where the impedance decreases.

The voltage amplification of an r.f. amplifier may be as

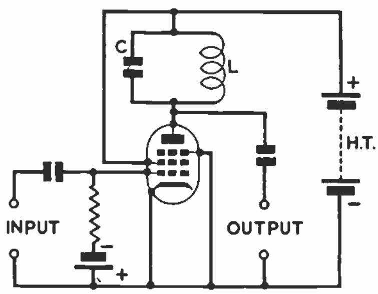


Fig. 1.41. Basic radio frequency amplifier circuit (tuned). A pentode is used instead of a triode to simplify the problem of eliminating unwanted feedback.

much as 100 or even higher. If a secondary winding of a few turns is wound round the coil in the anode tuned circuit *LC*, a voltage will be obtained which can be fed back to the grid circuit as shown in **Fig. 1.42**. The arrangement then really becomes an amplifier capable of supplying its own input and, provided that the sense of the feedback winding is correct, continuous oscillations will be produced at a resonant frequency determined by the values of *L* and *C*. Such a tuned amplifier with suitable feedback from output to input forms the basis of most oscillators used in transmitters and receivers.

Negative Feedback

The type of feedback mentioned in the previous paragraph is called *positive feedback* because the signal fed back from output to input is in phase with the input signal. If an amplifier circuit is arranged so that some of the output is fed back to the input in such a way that it is 180° out of phase with the input signal, *negative feedback* is said to have been

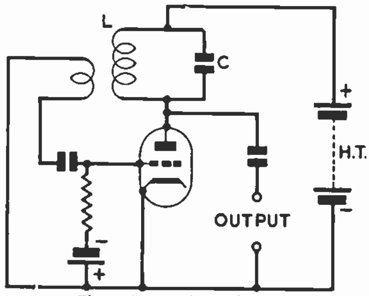


Fig. 1.42. Basic oscillator.

applied and a reduction in the overall gain of the amplifier ensues.

At first sight such a reduction of gain appears to be a disadvantage; however, the application of negative feedback yields several advantages which more than offset the reduction in gain. These advantages are—

(a) A reduction in harmonic distortion caused in the amplifier.

(b) A flattening of the amplifier response curve.

(c) A reduction in hum and other amplifier noise.

(d) A stabilization of amplifier performance against variations in supply voltages, component values and in valve characteristics.

The way in which negative feedback reduces harmonic distortion may be explained simply as follows. Suppose an amplifier is being fed with an input signal which is a pure tone (i.e. an undistorted sine wave). Due to imperfections in the amplifier, harmonics of the input signal will be produced which will appear in the output as harmonic distortion. Clearly, by feeding the correct amount of each harmonic, in the correct phase, into the input terminals of the amplifier from some auxiliary source, it would be possible to cancel out the distortion produced in the amplifier. Negative feedback from the output circuit provides just such a type of distortion-cancelling signal and, although complete elimination of distortion cannot be achieved, provided that the gain of the amplifier is high and that a sufficient fraction of the output is fed back, a very worthwhile reduction in distortion may be obtained.

A further reference to negative feedback is to be found in Chapter 9 (*Modulation*).

PIEZOELECTRIC EFFECT

Certain crystalline substances have the property of developing an electric charge on their surfaces when subjected to mechanical stress and, conversely, of exhibiting mechanical strain when their surfaces are electrically charged. This is known as the *piezoelectric effect*.

For radio purposes the most widely applied piezoelectric substance is quartz. This is used in high-stability radio-frequency oscillators and in sharply tuned frequency filters. Another piezoelectric material is Rochelle salt (sodium potassium tartrate): this has somewhat different characteristics from quartz, and is used in crystal microphones and in crystal headphones and loudspeakers. In amateur practice, however, these items seldom if ever require any attention and it is only the quartz crystal that is likely to merit a detailed study.

Quartz Crystals

Apart from the imperfections which normally occur in nature, a crystal of quartz is shaped like a hexagonal prism with a hexagonal pyramid at each end, as shown in **Fig. 1.43 (A)**. Typical crystals may be an inch or two in diameter and several inches long, and from these a large number of thin slabs or plates having the required electromechanical characteristics can be cut.

If one of these plates is made to vibrate by mechanical means, it will develop an alternating voltage between its opposite faces, and vice versa if an oscillatory voltage is applied to a pair of electrodes on opposite sides of the plate it will vibrate at the same frequency as the applied voltage. If this frequency happens to coincide with a mechanical resonance of the plate, the vibration will reach a considerable amplitude and in extreme cases it may be sufficient to cause the plate to fracture, thereby rendering it permanently worthless. The actual amplitude of vibration is always microscopically small, but the corresponding voltage developed may be of the order of several volts.

The characteristics of such a quartz plate depend on the way in which it has been cut from the natural crystal in relation to the crystallographic axes. These are indicated in **Fig. 1.43 (B)**. Axes such as *XX* which bisect the angle between any pair of adjacent faces of the hexagonal section are called the *electric axes*, and those such as *YY* which are perpendicular to any of the faces of the crystal are called the *mechanical axes*, while the axis *ZZ* is called the *optical axis*. If a plate is cut so that its faces are perpendicular to an *XX*-axis or to a *YY*-axis, it is said to be an *X*-cut or a *Y*-cut crystal respectively. Innumerable other cuts can be chosen, simply by tilting the plane of cutting so as to make various angles with the axes, but only a few particular orientations are used in practice and each of these has its own special advantage. An example is the *AT*-cut crystal: in this type the plane of the cut is rotated 35° from the *Z*-axis, as shown in **Fig. 1.44**.

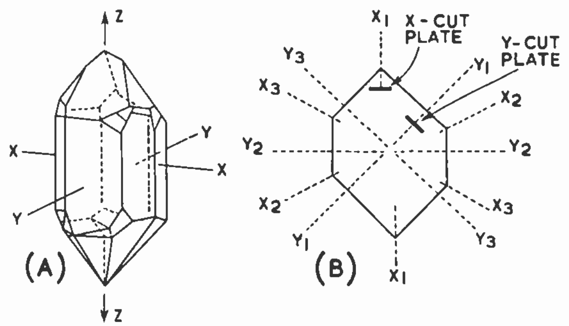


Fig. 1.43. The sketch at (A) shows a double-terminated natural quartz crystal and indicates the relationship between the X, Y and Z axes. Usually the crystal is found to have one pyramidal termination and one rough end where it has been broken from the parent rock. The sectional drawing (B) shows the various X and Y axes as viewed along a direction parallel to the Z-axis and includes examples of positions of X-cut and Y-cut plates.

Any quartz plate will have several different mechanical resonances, depending on the mode of vibration. In the ordinary type of crystal used in h.f. transmitters, the most important mode is the thickness vibration, since this gives the highest frequency for a reasonable set of dimensions. Even so, the plate has to be made quite thin to achieve a usefully high frequency. For instance, a crystal designed for 7 Mc/s might have a thickness of only 15 thousandths of an inch. A slight variation in thickness corresponds to a large change in frequency, and therefore to obtain a strong piezoelectric response the plate must be uniformly thick within very fine limits over its entire area.

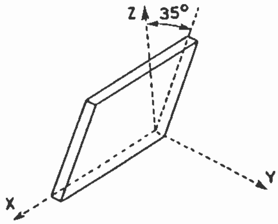


Fig. 1.44. An AT-cut crystal has its major surfaces at an angle of 35° to the Z-axis. This type of crystal has a low temperature coefficient and is widely used in r.f. oscillators.

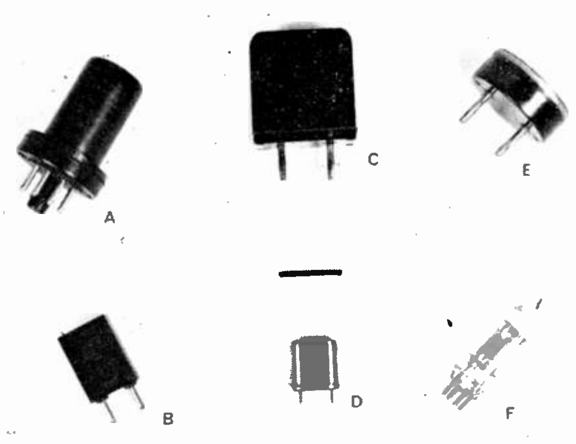
With this mode of vibration, the main resonant frequency is inversely proportional to the thickness. The precise relationship depends on the angle of cut, as shown in the few examples given here:

Type of Cut	*Frequency (kc/s) × Thickness (mm)*
X	2860
AT	1675
BT	2500

Other factors to be considered include the temperature coefficient of frequency, which is a measure of how much the frequency varies with temperature. This is of great importance in high-precision frequency control, and a discussion of it will be found in Chapter 6 (*H.F. Transmitters*).

Quartz is an exceptionally hard substance and the necessary cutting, grinding and polishing operations are very tedious. The usual materials employed are diamond, carborundum, emery and jeweller's rouge. Great accuracy is essential in the dimensioning and finishing processes, similar to that found in the highest class of optical work.

The grinding of the crystal by these methods results in some damage to the crystalline structure of the surface. Much improved stability and performance can be achieved by using a final etching process, whereby the damaged surface is dissolved away by means of a bath of hydrofluoric acid solution or of ammonium bifluoride. The final adjustment of the crystal to the required frequency is invariably carried



Six types of quartz crystal holders. A—International Octal. B—FT243. C—10X. D—HC-6/U. E—P5. F—B7G.

out today by this method, which also leaves the crystal perfectly clean.

The acids used for etching are *extremely dangerous* and the method should only be used by those experienced in their handling. In particular, the acid must not be allowed to come in contact with the skin. If it is, medical attention should be obtained *immediately*.

The quartz plate is usually clamped between two flat metal electrodes, and there may or may not be a small air gap adjacent to one of the crystal surfaces. When it is desired to operate with a gap, the plate is clamped at the periphery if it is circular or at the corners if it is rectangular or square. Alternatively, thin film electrodes of gold or silver may be deposited on the faces of the crystal plate and connecting wires can then be soldered to the metallic films at suitable points (the *nodal points*): these wires also serve as supports. The assembly is sometimes enclosed in an evacuated glass envelope to exclude dirt and moisture, both of which have a deleterious effect on the performance of the crystal.

Although such a quartz plate may appear quite similar in constructional features to a small capacitor, its electrical behaviour is very different. Because of the combination of the piezoelectric properties and the mechanical vibration characteristics it behaves as a series-tuned circuit having a very high L/C ratio and a very high Q value. The internal frictional losses in the crystal are remarkably small and consequently the equivalent resistance is very low. This is of course the effective resistance, considered as part of the equivalent circuit, to r.f. currents: the d.c. resistance is

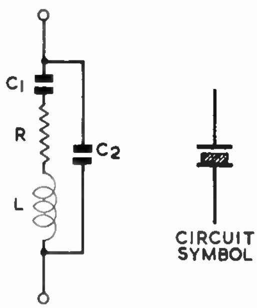


Fig. 1.45. Equivalent circuit of a piezoelectric crystal as used in r.f. oscillators or in filter circuits. Typical values for a 400 kc/s plate are—
C1 = 0·04 pF.
C2 = 6 pF.
L = 3·5 H.
Q = 25,000 or higher.
C2 represents the capacitance of the electrodes and the holder. The symbol used in circuit diagrams to represent a piezoelectric crystal is shown on the right.

extremely high. Fig. 1.45 shows the equivalent circuit of a simple quartz plate: the shunt capacitor C2 represents the capacitance of the electrodes and the connecting leads.

A circuit of this kind has two resonant frequencies, one corresponding to a high impedance and the other to a low impedance. They are indicated in Fig. 1.46 which shows how the r.f. current flowing through a typical crystal changes when the frequency of the applied r.f. voltage is varied. The resonant (or series) frequency corresponds to resonance in the series combination, LC1, which behaves as a low-impedance acceptor circuit. The so-called anti-resonant (or parallel) frequency corresponds to resonance in the high-impedance rejector circuit formed by the LC1 combination in parallel with C2.

The difference between the two frequencies is normally a small fraction of one per cent of the nominal frequency. Although there are circuits which have been devised for operating such a crystal on one or other of these frequencies, the majority of circuits used for frequency control in amateur transmitters utilize the parallel or anti-resonant high-impedance condition.

The most important feature with regard to frequency is that for either the resonant or the anti-resonant condition

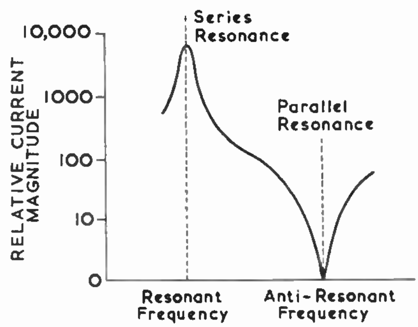


Fig. 1.46. Typical variation of current through a quartz crystal with frequency. The resonant and anti-resonant frequencies are normally separated by about 0·01 per cent of the frequencies themselves (e.g. a few hundred cycles for a 7 Mc/s crystal).

the impedance variation is extremely sharply defined. Thus the impedance may be found to vary by a ratio of more than 10,000 : 1 when the frequency is changed by merely a few hundred cycles. This may be expressed in terms of the equivalent circuit by saying that the Q-value is very high. In fact, a quartz crystal behaves as a tuned circuit with a Q-value many times greater than can be achieved even with the best inductors and capacitors.

By using a quartz crystal in place of the normal tuned circuit in an ordinary feedback type of r.f. oscillator, the frequency of oscillation is automatically stabilized within very close limits. Such an oscillator is known as a *crystal-controlled oscillator* (or more simply, a *crystal oscillator*), and a typical circuit is shown in Fig. 1.47.

Besides the thin flat plate type of crystal which is preferred for use in h.f. oscillators, a bar type is often used for resonators in filters operating at relatively low frequencies (below about 500 kc/s): an example of this application is the *crystal filter* (sometimes known as a *crystal gate*) in a high-selectivity intermediate-frequency amplifier: see Chapter 4 (*H.F. Receivers*).

Certain crystals have a tendency to operate at unwanted

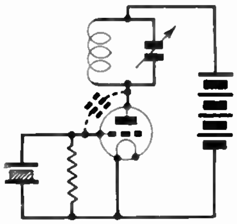


Fig. 1.47. Basic crystal-controlled oscillator circuit using a quartz crystal. The necessary feedback is provided by the grid-to-anode capacitance (shown dotted).

or spurious frequencies, owing to mechanical coupling effects in the crystals themselves, and the circuits in which they are used may have to be designed to prevent any possible operation at a frequency other than the wanted frequency.

Where a crystal is required to oscillate at a frequency which is above the normally attainable maximum (usually determined by the minimum practicable thickness), it can be made to vibrate at an overtone frequency if the circuit is arranged to stimulate this action. An overtone is approximately (never exactly) a multiple of the fundamental frequency, and in most cases this type of operation is limited to the odd multiples since the even multiples are associated with modes of vibration which present difficult problems in mounting.

SEMICONDUCTOR DEVICES

Since the Second World War a number of electronic devices using semiconductor materials such as germanium and silicon have been introduced. These devices have brought about a revolution in communication engineering comparable with that which followed the invention of the thermionic valve at the beginning of the century.

The range of devices includes semiconductor diodes and also semiconductor triodes, known as *transistors*, which have the property of being able to amplify signals. All these devices have the important advantage of consuming less power than their thermionic valve counterparts, much of this saving arising from the fact that no cathode-heater power is required in the semiconductor devices. A practical example which illustrates this overall saving in power is a small gramophone amplifier delivering an audio output of two watts. The transistor version consumes 3·5 watts representing an efficiency of 60 per cent whereas an equivalent valve amplifier consumes 12 watts corresponding to an efficiency of 16·66 per cent. Quite apart from the obvious saving in power consumption, transistors offer the advantage of small size and weight which makes them very suitable for use in portable equipment.

A common example of a semiconductor device is the point-contact *germanium diode*. It consists of a small wafer of germanium, containing a very small but controlled amount of arsenic or antimony, in contact with a "cat's whisker" of wire which may be of phosphor bronze. The general arrangement is shown in Fig. 1.48, the completed diode being sealed against moisture in a glass envelope. Diodes of this type may have a forward resistance of a few

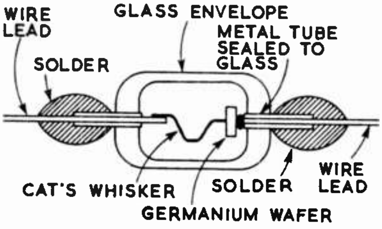


Fig. 1.48. Point-contact germanium diode.

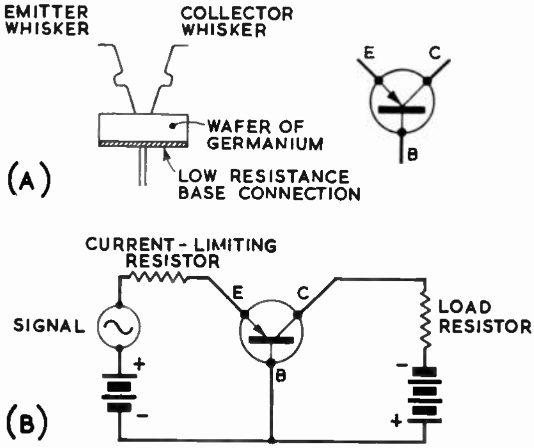


Fig. 1.49. Point-contact transistor. The basic arrangement of the electrodes is shown at (A): when connected in the circuit (B) it acts as an amplifier giving a current gain.

hundred ohms and a reverse resistance of several hundreds of thousands of ohms and are often used as radio-frequency rectifiers or detectors.

The earliest forms of *transistors* were "point-contact" types comprising a thin wafer of germanium, to one side of which a low-resistance or base connection was made, and a pair of closely spaced "cat's whisker" electrodes pressing on the other side of the wafer and called the *emitter* and *collector* electrodes. The general arrangement of the electrodes is shown in Fig. 1.49 (A). When connected in the circuit of Fig. 1.49 (B) the point-contact transistor has the property of giving a current gain; i.e., for a certain change in current in the emitter-to-base circuit two to three times that change of current will occur in the collector-to-base circuit. In a typical transistor the emitter circuit may have a resistance of a few hundred ohms whereas the collector circuit may have a resistance of several tens of thousands of ohms, and the overall power gain is therefore considerable.

Point-contact transistors have been superseded by junction transistors because of their superior characteristics, such as lower noise level, higher efficiency and greater

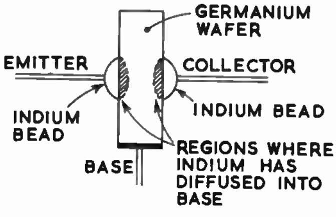


Fig. 1.50. Basic arrangement of a junction transistor.

power-handling ability. **Fig. 1.50** illustrates one method of junction-transistor construction. Two small beads of indium are placed on either side of a thin wafer of germanium and then heated sufficiently to cause the indium to diffuse into the germanium as indicated by the shaded areas. This process can be controlled so that there is only a very small gap between the indium-diffused areas. The junctions which form the transistor are (*a*) between one indium-diffused area and the wafer and (*b*) between the wafer and the other indium-diffused area. Emitter and collector connections are made to the indium beads and a low-resistance base connection is made to the germanium wafer. The transistor is then sealed in a metal or glass envelope to prevent ingress of moisture or other contaminants.

Junction transistors exhibit a current gain of slightly less than unity but the power gain available is greater than for the point-contact type since the ratio of output to input resistance is higher.

Low-power transistors are available which operate at frequencies of hundreds of megacycles per second. High-power types capable of handling powers of several watts are also manufactured. Detailed information on transistors will be found in Chapter 3 (*Semiconductors*).

THE AMATEUR TRANSMITTER

The purpose of the transmitter is to provide radio frequency energy which can be fed into an aerial for transmission to a distant point on a chosen frequency.

The simplest type of transmitter is one which will send Morse signals in the form of long and short bursts of r.f. energy corresponding to the dashes and dots of the code. A rather more elaborate arrangement is required when the transmission is to be in the form of speech or vision signals. An outline of the function of each stage is given here, but for further details the reader is referred to Chapter 6 (*H.F. Transmitters*).

Modern transmitters are usually made up of several distinct circuits or stages consisting of

(a) a master oscillator,
(b) one or more frequency-multiplying stages,
(c) a power amplifier, or output stage, and
(d) an aerial coupler.

For speech or television transmission the necessary modulation is introduced by a further distinct unit,

(e) a modulator.

A typical telegraphy and telephony transmitter system is shown in block diagram form in **Fig. 1.51**.

The purpose of the master oscillator is to generate a stable high-frequency oscillation which will control the frequency of the emissions from the transmitter. A high degree of frequency stability is required not only to avoid interference with other stations operating on nearby frequencies in the crowded amateur bands but also to ensure that the signal remains within the frequency bandwidth, or passband, of modern highly selective receivers and shows no tendency to deviate from it. A high degree of frequency stability is usually obtained by operating the master oscillator at low power to avoid excessive generation of heat and by selecting good-quality components of high electrical and mechanical stability and by using a master-oscillator tuned circuit of high Q.

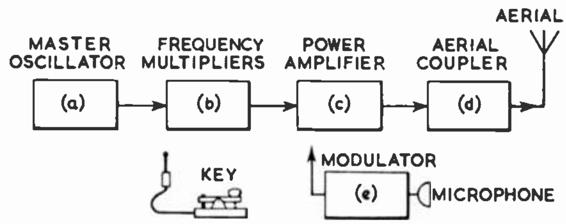


Fig. 1.51. Block diagram of a typical amateur transmitter for telegraphy and telephony. A single power supply unit is sometimes relied upon to feed the entire transmitter but there are usually important advantages to be gained by providing separate power supply units especially for the power amplifier and the modulator.

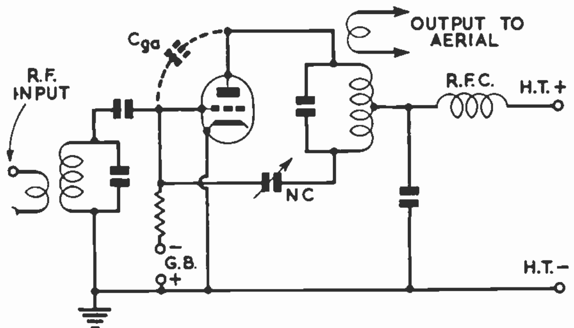


Fig. 1.52. Triode power amplifier circuit showing the use of a neutralizing capacitor NC to balance the grid-to-anode capacitance.

As the amateur bands are for the main part harmonically related it is common practice to operate the master oscillator in the lowest frequency amateur band and obtain sufficient power to drive the power amplifier or output stage from a succession of frequency-multiplying stages. For example, an oscillator capable of being varied between 1·75 and 2 Mc/s will cover not only the 160m amateur band but also the 80, 40, 20, 15 and 10m bands when multiplied by 2, 4, 8, 12 and 16 respectively. Frequency-multiplier stages use heavily-biased valves which are driven with large input voltages. This has the effect of causing the anode current to flow in short pulses instead of following the sinusoidal waveform of the drive voltage. The pulse-like anode current contains components of frequencies which are multiples of the frequency of the input waveform. A voltage at any one of these frequencies may be developed at the anode by means of a resonant circuit tuned to the required frequency. It is sometimes practicable to select up to the fifth harmonic in one stage; for higher orders of frequency multiplication several stages may be used in cascade.

The power-amplifier stage is driven by a low-power stage at the desired output frequency, a watt or two being sufficient to drive almost any of the pentode or tetrode amplifier valves likely to be used in amateur transmitters. For high-efficiency, a radio-frequency power amplifier is run under class C conditions: see Chapter 2 (*Valves*).

Triode amplifiers are not as often used as tetrodes or pentodes because they require more drive power and need *neutralizing* to avoid self-oscillation. The self-oscillation in a triode amplifier stage occurs as a result of feedback from the anode to the grid circuit through the capacitance which exists between grid and anode inside the valve. The feedback may be balanced out, or neutralized, by arranging for an equal feedback voltage of opposite phase to be applied to the grid circuit. This is usually achieved in amateur transmitters by centre-tapping the tuned circuit of the output amplifier as shown in **Fig. 1.52** to obtain an antiphase

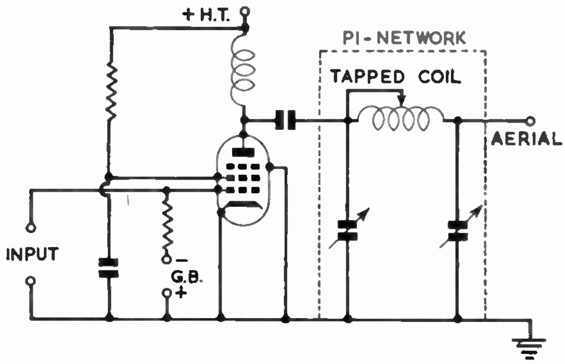


Fig. 1.53. A pentode power amplifier with a pi-network coupler in the anode circuit.

neutralizing voltage which is fed back to the grid circuit through the small neutralizing capacitor NC.

There are many types of aerial which can be used with a transmitter, and an aerial coupling or tuning unit may be necessary to ensure that the power generated by the amplifier stage is efficiently transferred to the aerial. Generally speaking aerial coupling only becomes a difficult problem if a transmitter has to operate and supply power to an aerial over a wide and continuous range of frequencies; since amateur transmitters operate only in narrow specified bands of frequencies the coupling circuit in practice can often be reduced to merely an inductive link comprising one or two turns of wire wound round the power-amplifier output coil as indicated in the circuit of Fig. 1.52.

A useful combined aerial coupling and power-amplifier output circuit is the pi-network coupler. It has the advantage of operating over a considerable frequency range and feeding efficiently into a wide range of load impedances. It also has the merit of helping to reduce television interference. The circuit is shown in **Fig. 1.53**. The proper tuning of the aerial and loading of the transmitter output stage is obtained by adjusting the two variable capacitors and the tapped coil.

A radio frequency power amplifier may be considered as a generator which converts d.c. power taken from the h.t. supply into radio frequency power to be fed to the aerial. Important factors in connection with an amplifier are:

(a) The d.c. power input.
(b) The r.f. power output.
(c) The efficiency of the amplifier.
(d) The anode dissipation.

The d.c. input power W_{in} is equal to the product of the h.t. supply voltage V_a and the anode current I_a of the valve as shown in **Fig. 1.54** thus:

$$\text{Power input } W_{in} = V_a \times I_a$$

The r.f. output power is often calculated by measuring the r.f. current flowing in a load resistor sometimes called a *dummy load.* This load resistor should be non-inductive and may conveniently be of the high-power carbon type. The output power W_{out} may be calculated by finding the product of the load resistance R and the square of the load current I_{rf} thus:

$$\text{Power output } W_{out} = R \times I_{rf}^2$$

The efficiency η of the amplifier gives a figure of merit for the stage as a converter of d.c. into r.f. power. The efficiency is calculated by dividing the power output by the power input and multiplying by 100 to express the result as a percentage: thus

$$\text{Efficiency } \eta = \frac{W_{out}}{W_{in}} \times 100 \text{ per cent}$$

It is clear that since no amplifier can be 100 per cent efficient more power is fed into the amplifier than appears at the output as r.f. power. Most of this excess of power is spent or dissipated as heat at the anode of the amplifier valve. The anode dissipation W_{dis} may be calculated by taking the difference between the input and output power:

$$\text{Anode Dissipation } W_{dis} = W_{in} - W_{out}$$

As an example of the preceding paragraphs consider the case of a valve operating with an h.t. supply of 750 volts and an anode current of 100 mA. If the amplifier is producing a current of 0·71 ampere in a load resistor of 100 ohms

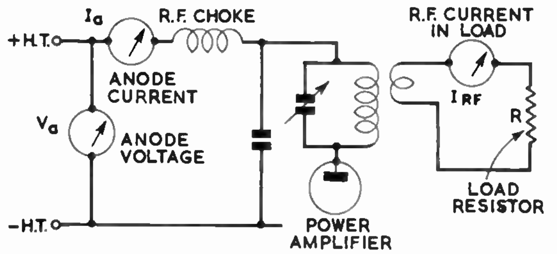


Fig. 1.54. Diagram showing how the d.c. power input to a radio frequency amplifier and the r.f. power which it produces can be measured.

calculate (*a*) the power input, (*b*) the power output, (*c*) the efficiency and (*d*) the anode dissipation of the stage.

(a) Power input $W_{in} = 750 \times 0{\cdot}1 = 75$ watts.

(b) Power output $W_{out} = 100 \times (0{\cdot}71)^2 = 50$ watts.

(c) Efficiency $\eta = \frac{50}{75} \times 100 = 66{\cdot}66$ per cent.

(d) Anode dissipation $W_{dis} = 75 - 50 = 25$ watts.

As so far described the transmitter is capable of generating a steady radio-frequency signal called a *carrier wave.* This in itself conveys no information and for messages to be transmitted by telegraphy the carrier wave must be broken up into the characters of the Morse code as indicated in **Fig. 1.55.** This process is called *keying* and various methods which may be adopted are described in Chapter 8 (*Keying & Break-in*).

The transmission of voice or music (audio) waveforms is achieved by modulating the carrier wave, i.e. causing its amplitude, frequency or phase to vary in sympathy with the audio. Modulation produces new frequencies called

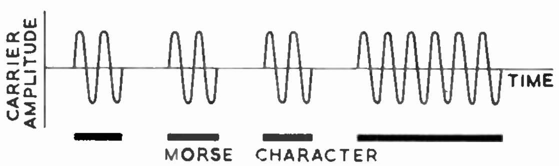


Fig. 1.55. Morse code signals are produced by breaking the carrier wave into long and short characters. The illustration represents the letter V.

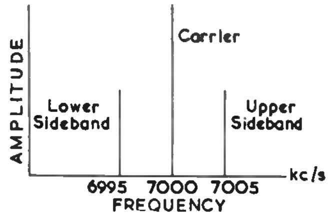


Fig. 1.56. Sideband spectrum of a 7 Mc/s carrier amplitude modulated by a 5 kc/s audio tone.

sidebands, extending above and below the carrier frequency by at least the highest frequency component of the audio wave.

The most commonly used type of modulation is amplitude modulation (a.m.) in which the amplitude of the transmitter output is varied up and down in sympathy with the audio modulating waveform, the carrier frequency being constant. In practice this variation of amplitude is usually achieved by varying the anode, screen or grid voltage of a Class C amplifier stage according to the audio waveform it is required to transmit. **Fig. 1.57** shows the unmodulated carrier wave at the top, the audio modulating signal in the centre and the composite amplitude modulated wave at the bottom. If the output from a 100 per cent amplitude modulated transmitter is displayed on an oscilloscope whose timebase is locked to the audio signal it will appear as shown at the foot of Fig. 1.57.

An amplitude modulated radio signal actually consists of three distinct components; the carrier wave of constant amplitude, and two sidebands, or side frequencies, spaced on either side of the carrier. These sidebands carry the audio intelligence and vary in amplitude as the audio signal changes in intensity. The frequency spacing between the sidebands and the carrier is equal to the frequency of the modulating signal.

The frequency spectrum corresponding to a 7 Mc/s carrier wave modulated fully, or 100 per cent, by an audio signal of 5 kc/s is shown in Fig. 1.56. At full modulation the amplitude of each of the two sidebands is one half that of the carrier, the power in each sideband being one quarter of that in the carrier. The power in the two sidebands taken together is therefore one half that of the carrier when it is fully modulated. When modulated by speech the sidebands are complex and vary continually in frequency spacing and amplitude following the rapidly changing speech waveform. The sidebands extend on either side of the carrier up to a frequency spacing corresponding to the highest frequency component present in the speech waveform. In practice good telephone quality speech contains frequencies up to about 3 kc/s so that in an a.m. signal of that quality the sidebands occupy 3 kc/s of the frequency spectrum above and below the carrier.

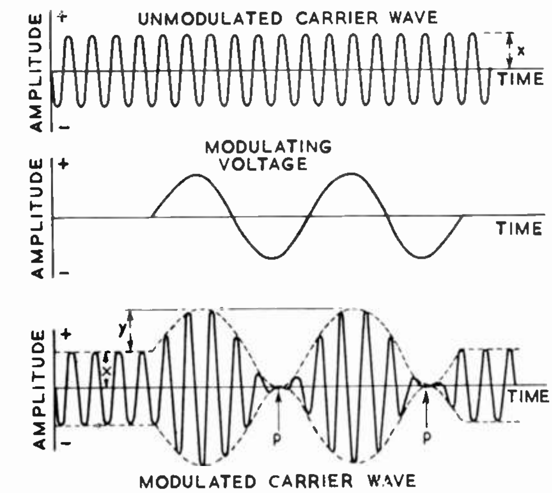


Fig. 1.57. Amplitude modulation of a carrier. The modulation envelope may be displayed on an oscilloscope and represents 100 per cent modulation by a sinusoidal signal. It shows graphically how the sum of the carrier and sideband voltages varies with time.

When the combined sideband voltages are equal to the carrier voltage, the transmitter is said to be modulated 100 per cent. If the amplitude of the modulating voltage applied to the power amplifier valve exceeds that required for 100 per cent modulation, the valve will cease to conduct at the negative peak of the audio modulating voltage. This interruption of the carrier will result in severe distortion of the transmitted signal and must be avoided as it would cause serious interference to signals from other stations operating on adjacent frequencies.

Intermediate depths of modulation between zero and maximum values are usually expressed as a percentage. If x is the unmodulated amplitude of the carrier envelope and y is the peak increase (or decrease) in amplitude of the envelope during modulation, the depth of modulation may be obtained by dividing y by x and multiplying by 100 to express the result as a percentage. Expressed as a formula,

$$M = \frac{y}{x} \times 100\%$$

where M is the modulation percentage.

The simple a.m. signal is wasteful in two ways. First a channel of bandwidth equal to twice the highest frequency of the audio signal used is required, and second at least two-thirds of the radiated power is in the carrier which conveys no actual information and could equally well be omitted and reintroduced at the receiver. Both of these disadvantages are eliminated in the single sideband suppressed carrier method of transmission in which only one set of sideband frequencies and no carrier is transmitted. Full details of this system will be found in Chapter 10 (*Single Sideband Techniques*). The a.m. system has, however, the advantage of simplicity, both at the transmitter and in the receiver.

Ordinary amplitude modulation is usually obtained by varying the h.t. feed to a class C radio frequency amplifier, according to the audio waveform to be transmitted, by applying the modulating voltage in series with the h.t. supply through the medium of a modulation transformer as shown in **Fig. 1.58.** In the method just described, called *anode modulation* or *plate modulation*, the amount of audio power required to modulate the carrier wave fully is equal to one-half of the d.c. power input to the anode and screen circuits of the p.a. stage. For example, a transmitter power amplifier drawing an anode current of 100 mA from an h.t. supply of 600 volts has an input of 60 watts. A maximum audio power of 30 watts is therefore required to modulate this amplifier fully.

In frequency modulation (f.m.) the amplitude of the carrier wave is kept constant but its frequency is varied, or swung to and fro on either side of the nominal value, at a rate corresponding to the modulating frequency and to an extent corresponding to the amplitude of the modulating signal. An f.m. signal has a greater bandwidth than an a.m.

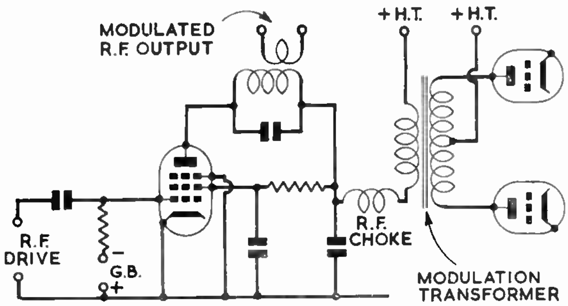


Fig. 1.58. A popular method of applying amplitude modulation to a radio frequency power amplifier.

signal, but on the other hand an f.m. system offers the advantage of a better signal-to-noise ratio provided that the strength of the received carrier is above a certain threshold value. Frequency modulation is thus well suited for the v.h.f. bands where ample bandwidth is available and a good signal-to-noise ratio is desirable.

Further information on methods of amplitude and frequency modulation will be found in Chapter 9 (*Modulation*).

Although a valve used in an amplitude-modulated power amplifier must be run at a reduced input compared with the telegraphy condition to avoid excessive dissipation and possible damage, an advantage of frequency modulation is that the power-amplifier stage can be operated at the full telegraphy rating thus allowing a greater output to be obtained from a given valve than under a.m. conditions.

THE AMATEUR RECEIVER

The purpose of a receiver is to select a desired radio signal from amongst the many others that may be receivable and to recover from it the intelligence that it conveys. Usually the power arriving at the aerial terminals is so minute (perhaps only a small fraction of a micro-microwatt) that the receiver also needs to contain an amplifier so that the received signal can be made powerful enough to actuate the headphones or a loudspeaker. A receiver therefore normally has three functions—(i) the selection of the desired signal by a selective tuning circuit, (ii) the conversion of the radio-frequency signals into audio-frequency signals by the process of demodulation or "detection," and (iii) the amplification of the signals to a useful power level.

There are two main types of receiver—the straight, or tuned-radio-frequency (t.r.f.) receiver, and the superheterodyne receiver or "superhet." Of these the more commonly used today is the superheterodyne, because of its superior

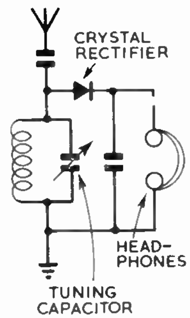


Fig. 1.59. Simple crystal receiver.

selectivity and ease of handling. The t.r.f. receiver offers the advantage of simplicity and cheapness and therefore lends itself more readily to home construction.

The simplest possible receiver comprises a single tuned circuit and a crystal detector, as shown in **Fig. 1.59**. Such a circuit is not suitable for short-wave work since it lacks both selectivity and sensitivity. A triode (or tetrode or pentode) valve may be used as a detector as shown in **Fig. 1.60**, the grid and cathode acting as a diode detector and

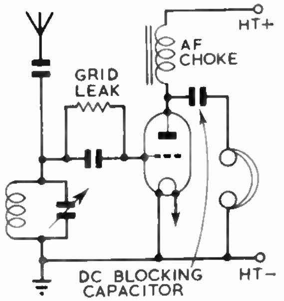


Fig. 1.60. Simple triode valve receiver.

the amplifying properties of the valve providing a certain amount of signal magnification. Although this circuit is more sensitive than that of the crystal set it still lacks selectivity.

Both the sensitivity and selectivity of a valve detector may be increased enormously by applying reaction as shown in

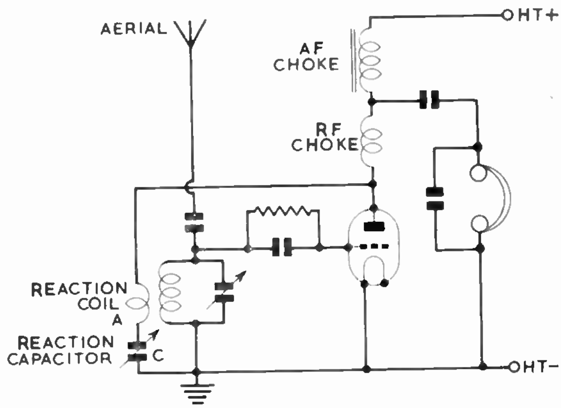


Fig. 1.61. Simple receiver using a triode valve with reaction.

Fig. 1.61. Radio frequency energy in the anode circuit of the valve is fed back to the grid circuit by the reaction coil *A*. The series capacitor *C* is inserted to control the amount of feedback, or reaction. By increasing the value of *C*, the reaction is increased and the sensitivity of the detector is increased: eventually a point is reached at which self-oscillation begins. This is marked by an increase in the background noise in the receiver, a characteristic rushing sound, and by the ability of the receiver to produce a heterodyne beat note with a continuous-wave telegraphy signal. For telephony reception this circuit is used with the detector in a condition just short of oscillation.

The amount of audio power available from a single detector stage even with reaction is small, and further audio

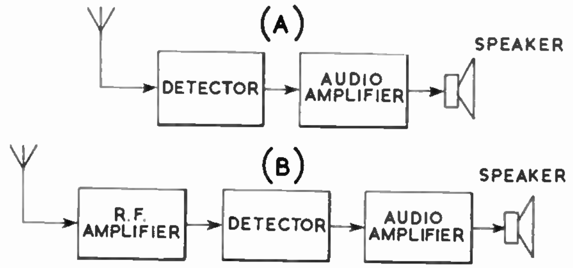


Fig. 1.62. Methods of providing amplification in a receiver: in (A) audio frequency amplification is applied to the output from the detector: in (B) the radio frequency signal is amplified before it is fed into the detector and further amplification is provided for the audio frequency output from the detector.

amplification may be added so that a loudspeaker can be used. This is illustrated in the block diagram **Fig. 1.62 (A)**.

It is not good practice to connect an aerial directly to a detector valve for two reasons. First, when the detector is oscillating there will be radiation from the aerial which will cause interference with other neighbouring receivers and, second, the optimum setting of the reaction control, a somewhat critical adjustment, will be adversely influenced by variations in the impedance of the aerial when the receiver is tuned over a band of frequencies. To eliminate these disadvantages it is common practice to use a stage of radio frequency amplification ahead of the detector as shown in **Fig. 1.62 (B)**.

A code is sometimes used to describe the make-up of a t.r.f. receiver. It takes the form of a numeral representing the number of radio frequency amplifier stages, followed by the letter *V*, to represent the detector stage, and this is followed by a further numeral to represent the number of audio amplifier stages after the detector. Thus 1–*V*–2 means a t.r.f. receiver with one r.f. stage, a valve detector and two a.f. stages; the receiver shown in Fig. 1.61 would be designated 0–*V*–0.

The best t.r.f. receivers are not really selective enough for use on the higher frequency bands. The superheterodyne receiver overcomes this drawback by changing the frequency of incoming signals to a lower fixed frequency called an *intermediate frequency*, usually abbreviated to *i.f.* At this lower fixed frequency, commonly around 465 kc/s, it is easy to obtain high selectivity and high gain with a small number of stages.

A special mixer valve is commonly used for frequency changing. Into it are fed (i) the radio frequency signal and (ii) a locally-generated oscillator voltage whose frequency is higher than the radio signal frequency by an amount equal

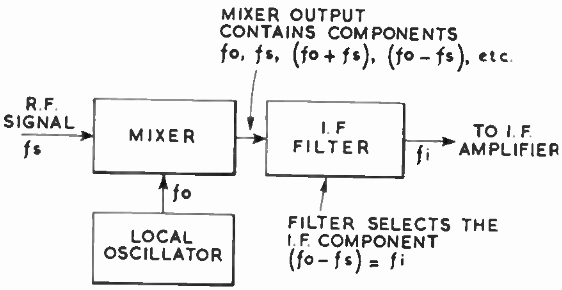


Fig. 1.63. The principle of frequency-changing.

to the intermediate frequency. The anode current of the mixer valve contains components at the signal frequency f_s the oscillator frequency f_o and the sum and difference frequencies $f_o + f_s$ and $f_o - f_s$: it also contains other components, but these are of no importance in the operation of the receiver. The component $f_o - f_s$ is the intermediate frequency and it is selected from the others by an i.f. transformer which usually comprises a pair of coupled tuned circuits tuned to the intermediate frequency, and is then passed on to an i.f. amplifier. The frequency-changing action is illustrated in the form of a block diagram in **Fig. 1.63** while **Fig. 1.64** shows a mixer circuit using a triode-hexode valve in which the various parts of the circuit are laid out to correspond with the block diagram.

The block diagram of a complete popular type of superheterodyne receiver is shown in **Fig. 1.65**. The tuning of the r.f. stage and the mixer and local oscillator stages is usually performed by a three-section ganged variable capacitor. A technical difficulty with this arrangement is that the oscillator

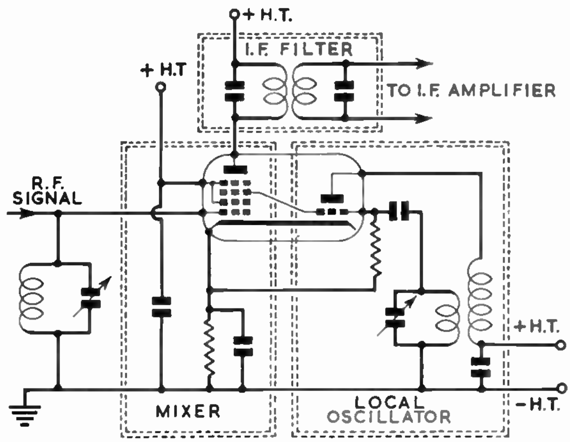


Fig. 1.64. A typical triode-hexode mixer circuit used for frequency-changing in accordance with the principle illustrated in Fig. 1.63.

must always be tuned to a frequency which is equal to the sum of the signal frequency and the intermediate frequency, irrespective of the setting of the tuning. The signal and oscillator circuits are said to *track* correctly if the above requirement is met. In order to obtain correct oscillator tracking it is necessary to modify the inductance of the oscillator coil and add two extra components, a trimmer C_T and a padder C_p as shown in **Fig. 1.66.** With this arrangement perfect tracking can be obtained at three points with only small errors elsewhere over quite a wide tuning range. Further details will be found in Chapter 4 (*H.F. Receivers*). It should be noted that special tracking circuits can be dispensed with if the frequency range covered by the receiver is small compared with the frequency itself, as for instance in a receiver designed exclusively for the amateur bands.

Since most of the gain and selectivity in a superheterodyne is obtained at the intermediate frequency it may appear unnecessary to use an r.f. amplifier ahead of the mixer. It is, however, desirable for two reasons, first to reduce or eliminate what is called *second channel* or *image interference*,

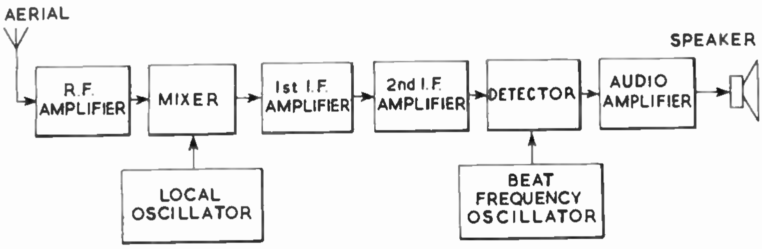


Fig. 1.65. Block diagram of a typical amateur communications superheterodyne receiver.

and second, to improve the signal-to-noise ratio of the receiver.

Second channel interference arises from the fact that there are two radio frequencies at which signals, if applied to the mixer, will give rise to an i.f. signal at its output. These are the frequency of the wanted signal which by design is equal to the *difference* between the oscillator frequency and the intermediate frequency and the so-called second channel or image frequency which is equal to the oscillator frequency *plus* the intermediate frequency. The second channel frequency is thus twice the i.f. below the frequency of the wanted signal and unless there is adequate selectivity in the r.f. tuning circuits to exclude signals on the second channel frequency, both will be received. A similar condition arises when the oscillator is on the low frequency side of the signal. The degree of r.f. selectivity required is not so high in a superheterodyne as in a straight receiver since it is merely a matter of rejecting a spurious signal differing in frequency from the wanted signal by twice the intermediate frequency instead of a spurious signal only a few kilocycles away.

Mixer valves introduce noise and a weak signal applied direct from an aerial to a mixer would be lost in the noisy background. An r.f. amplifier stage is much less noisy than a mixer, and by using such a stage ahead of the mixer an improved signal-to-noise ratio can be obtained.

In many superheterodyne receivers the i.f. amplifier comprises two amplifier stages and three associated i.f. transformers between the mixer and the detector. If greater selectivity is required than can be obtained with high Q coil-and-capacitor tuned transformers a crystal filter may be added to the i.f. amplifier, the crystal acting as another tuned circuit of extremely high Q. Such filters are commonly used for telegraphy reception and are described in Chapter 4 (*H.F. Receivers*).

The choice of the intermediate frequency for a receiver is a compromise. A low i.f. will give good selectivity and high stage gain but it is very difficult to obtain adequate image rejection particularly at higher signal frequencies. On the other hand a high i.f., while allowing good image rejection with few signal frequency tuned circuits ahead of the mixer tends to give poor selectivity, unless expensive crystal filters are used, and lower stage gain.

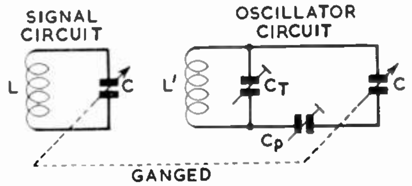


Fig. 1.66. In the oscillator tuning circuit of a superheterodyne receiver modifications are necessary to ensure proper tracking with the tuning adjustment of the signal-frequency circuit over a wide frequency range. Where the tuning range is strictly limited such modifications are not necessary.

The double superhet principle offers a solution to the dilemma just described. In this type of receiver the signal frequency is first changed to a relatively high first i.f. in order to get good image rejection and then changed again from that frequency to a much lower second i.f. in order to obtain good selectivity. For a receiver covering the amateur bands from 1·8 to 28 Mc/s the first i.f. might be chosen at about 1·6 Mc/s and the second i.f. at 100 kc/s.

When operating on v.h.f., triple conversion superhets are commonly used. Particularly at the higher frequencies it is common practice to use a crystal controlled first oscillator in a double superhet and tune the second oscillator and first i.f. circuits over a restricted band of frequencies (typically 500 kc/s). This arrangement has the advantage of good frequency stability, provided by the crystal oscillator, and constant bandspread on all bands, provided by the tunable second oscillator.

For c.w. telegraphy reception it is necessary to introduce an additional oscillator (known as the *beat frequency oscillator* or *b.f.o.*) to beat with the i.f. signal in the detector stage and produce an audible tone which can be read as Morse code.

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CHAPTER 2

VALVES

THE simplest form of valve is the *diode*, which consists of two electrodes, a *cathode* and an *anode*. The two electrodes are contained in an evacuated bulb and connections are made to them through external pins or contacts. If the cathode is heated, the molecules of which the cathode is composed become increasingly agitated, and if the temperature is high enough some of the electrons composing the molecules will be ejected from the cathode.

Electron Flow

When an electron leaves the parent molecule the latter becomes positively charged because the number of electrons remaining are insufficient to neutralize the positive charge in the molecule. Because the electrons are negatively charged, there is a force tending to pull them back to the cathode. The anode, which is positively charged to a higher potential, is placed near the hot cathode (usually surrounding it more or less completely) in order to attract these electrons. As they travel through the space from the cathode to the anode they may encounter molecules of gas (since the vacuum cannot be a perfect vacuum) and such collisions will impede their progress. Consequently, the amount of gas in the valve must be as small as possible. A valve which has been properly evacuated is described as *hard*. If an appreciable quantity of gas is present, the collision between the electrons and the molecules of gas will ionize the gas and a blue haze will become visible between the electrodes; the valve is then said to be *soft*. Such a blue haze should not be confused with a *blue glow* on the inside surface of the envelope.

Space Charge

In travelling from the cathode to the anode, the electrons form a cloud in the intervening space, and the electric charge associated with this cloud is known as the *space charge*. It tends to repel the electrons leaving the cathode because it is of the same polarity, but if the potential applied to the anode is sufficiently high, the effect of the space charge will be overcome and electrons will travel to it from the cathode, the current flow is completed through the external circuit back to the cathode. This means that there is an external electron flow from anode to cathode. In accordance with established convention, however, that the flow of electric current is " from positive to negative," i.e. in the opposite direction to electron flow, a meter will show a " current " flowing from the positive terminal of the high-tension supply towards the anode.

As the anode potential is increased the electron flow or current will increase up to a point where the space charge is completely neutralized and the total emission of the cathode is reached. The total emission can only be increased by raising the cathode temperature.

Obviously if the anode potential is reduced to zero or is made negative, there will be no electron flow because the space charge remains unneutralized. Hence the valve is able to conduct current in one direction only, and in fact the principal use of a diode is as a rectifier.

Cathodes

Although several types of cathode are used in modern valves, the differences are only in the method of producing thermionic emission. The earliest type is the *bright emitter* in which a pure tungsten wire is heated to a temperature in the region of 2500–2600° K. At such a temperature emission of 4 to 40 mA per watt of heating power may be obtained.

Bright emitters are still employed in high power transmitting valves for broadcasting but the only common amateur use is in diodes for applications such as noise generators. The life of a pure tungsten filament at full operating temperature is limited by evaporation of the tungsten, failure occurring when about 10 per cent has been evaporated.

Dull emitters are directly heated thoriated tungsten cathodes which produce greater emission than bright emitters but require less heating power. In a dull emitter, a small quantity of thorium oxide is introduced into the pure tungsten wire, a process known as carburization. This process creates an outer skin of tungsten carbide on the wire and facilitates the reduction of the thorium oxide to metallic thoria, stabilizes the emission and increases the surface resistance of the cathode to gas poisoning. Typical emission efficiency is in the region of 30–100 mA per watt of heating power at an operating temperature of 1900–2100° K. This type of cathode is relatively fragile and valves should not be subjected to shocks or sharp blows.

Provided the operating temperature is correctly maintained long life may be expected. In particular, the rated voltage should be closely controlled.

Oxide coated cathodes are the most common type of thermionic emitter found in both directly and indirectly heated valves. In this type, the emittive material is usually some form of nickel ribbon, tube or thimble coated with a mixture of barium and strontium carbonate, often with a small percentage of calcium. During manufacture, the coating is reduced to its metallic form and the products of decomposition removed during the exhaustion process. The active ingredient is the barium which provides much greater emission than thoriated tungsten at lower heating powers.

Typically, 50–150 mA per watt is obtained at temperatures of 950–1050° K.

Although the emission efficiency of oxide coated cathodes is high and large currents may be drawn, they are less able to resist the poisoning affects of gas or ion bombardment. This type must not be operated under temperature limited conditions.

An *indirectly-heated cathode* is a hollow metal tube or sleeve, or in some cases is of thimble shape, having a coating of emissive material on the outer surface. The cathode is heated by radiation from a metal filament, called the *heater*, which is mounted inside the cathode, and the heater is electrically insulated from the cathode. The emissive material is generally the same as that employed for filamentary oxide-coated cathodes and operates at about the same temperature. The cathode may be made of pure nickel or copper or of special alloys, depending upon the purpose of the valve. The heater is normally made of tungsten.

The life of valves with oxide coated cathodes is generally good provided the ratings are not exceeded. Occasionally there is some apparent reduction in anode current due to the formation of a resistance layer, between the oxide coating and the base metal, which operates as a bias resistor.

In *cold cathode* valves, such as gas stabilizers, the cathode is an activated metal surface.

Anodes

A valve anode is generally in the form of a hollow cylinder, which surrounds the cathode and other electrodes and is intended to collect as many as possible of the electrons ejected from the cathode. In small valves the anode is made of nickel or nickel-plated steel. When it is necessary to dissipate more heat, the nickel may be carbonized to give a matt black finish and thus improve its thermal radiation. In valves with higher anode dissipations the anode must be made of a material which can operate at high temperatures such as molybdenum, carbon, tantalum or zirconium; alternatively, radiating fins are attached to the anode or the anode may form part of the external envelope of the valve and then it can be readily cooled by thermal conduction to a mass of metal forming part of an external circuit or by a circulating-water jacket or an air blast. With the aid of these cooling methods a valve of relatively small physical size can be made to handle very large amounts of power.

Grids

The electron flow from cathode to anode can be controlled in various ways and for various purposes by causing it to pass through one or more *grid* electrodes; in some types of valve there may be as many as four or five grids in succession. A simple system of designation has been generally adopted whereby the generic name given to a valve indicates the total number of electrodes associated with the electron flow, starting from the cathode and ending at the anode: this is shown in Table 2.1.

TABLE 2.1

Number of Grids	Total Number of Electrodes	Generic Name
1	3	*Triode*
2	4	*Tetrode*
3	5	*Pentode*
4	6	*Hexode*
5	7	*Heptode*
6	8	*Octode*

Grids are usually made in the form of a wire helix but are sometimes composed of square-mesh gauze. The wire helix may be of nickel or molybdenum or an alloy wound on support rods of nickel or copper, and it may be circular, oval or rectangular in section. In some v.h.f. valves the grid is made of parallel wires tightly stretched across a hole in a disc or in the form of a squirrel cage. Many modern high performance valves use a form of construction known as *frame grid* which permits them to have a greater slope and shorter electron transit time than is possible with normal construction. One satisfactory method shown in Fig. 2.1 is to wind the grid wire under tension (about half its breaking stress) on a stiff rectangular frame to which it is firmly fixed by glazing or gold brazing. This produces a very rigid grid which can be spaced only a few thousandths of an inch from the cathode or other grids. The cathode must have a flat surface, the coating being ground flat if necessary.

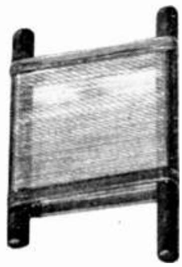


Fig. 2.1. Construction of a frame grid.

In beam valves, such as cathode ray tubes, travelling wave tubes and klystrons, the grid is in the form of a single hole in a plate through which the beam passes. A voltage applied to the plate controls the beam current. This type of grid is often known as a Wehnelt.

One of the most important requirements in valve design is to prevent the grid from becoming overheated. For this purpose the grid wire may be carbonized so as to enhance the heat radiation from it and cooling fins are often attached to the support rods. In some types these rods are welded directly to conducting pins in the base which permit the heat to be transferred outside the valve. Many modern high-performance valves employ gold or platinum-plated grids in order to avoid grid primary emission at the unavoidably high operating temperatures.

TYPES OF VALVES

Triodes

The insertion of a grid between the cathode and the anode enables the electron flow to be controlled by applying a relatively small voltage to this third electrode. To distinguish it from other grids serving different purposes in other types of valve, this grid is often referred to as the *control grid* (g_1). The power absorbed in the control-grid circuit is in many cases exceedingly small.

By varying the potential of this grid the space charge can be modified in the same way as in a diode. The grid, because it is an open helix or mesh, does not present any mechanical obstruction to the electron flow, and as long as the potential on it is negative with respect to the cathode (or the negative end of the filament) it will attract no free electrons, but if it becomes positive, electrons will flow to it exactly as if it were an anode: when this occurs *grid current* will flow. In most valves grid current begins to appear a little before the negative bias reaches zero: this is due partly to the random movement of electrons as they pass the grid on their way to the anode and partly to the contact-potential differences associated with the materials used in the cathode and grid construction.

When it is necessary to ensure that the grid is always negative, the grid circuit must include a source of adequate negative voltage. This voltage, known as the *grid bias*, can be obtained in four ways:

(a) An independent source of fixed voltage.
(b) A resistor in the cathode lead.
(c) Grid current in a grid resistor.
(d) Contact potential.

In (a) the source of voltage may be a battery (usually dry-cells), or in transmitting equipment or high-power audio amplifiers it may be a mains-driven power unit.

In (b) a resistor is connected in the cathode-to-earth lead of an indirectly heated valve or between the filament-supply centre-tap of a directly-heated valve and earth, and the total cathode current flowing through this resistor causes a voltage drop across it and hence the cathode becomes positive with respect to earth; if the grid return circuit is returned to earth the grid will be negative with respect to the cathode. The resistor is known as the *automatic-bias resistor* and the circuit arrangement as *cathode bias*.

In (c) if the valve is operated at such an amplitude that appreciable grid current flows and if the grid is connected to cathode and earth through a resistor, the flow of grid current will result in the grid becoming negative with respect to cathode. This form of grid bias is employed in oscillators, frequency multipliers and class C amplifiers and is known as *grid-leak bias*.

In (d) the grid is connected to the cathode through a high-value resistor (usually 1–10 Megohms) and the flow of a small amount of grid current will bias the grid to almost the contact potential. This form of grid bias is suitable only for small valves having very small signal input voltages.

If the grid potential is varied, the anode current will vary in a corresponding manner; further, if a resistance is connected in the anode circuit, the voltage drop across it will vary also in the same way. If this resistance has a suitable value the variation of voltage drop will be of greater amplitude than the variation of grid voltage; hence the ability of the triode valve to amplify.

If the relationship between the variation in grid voltage and

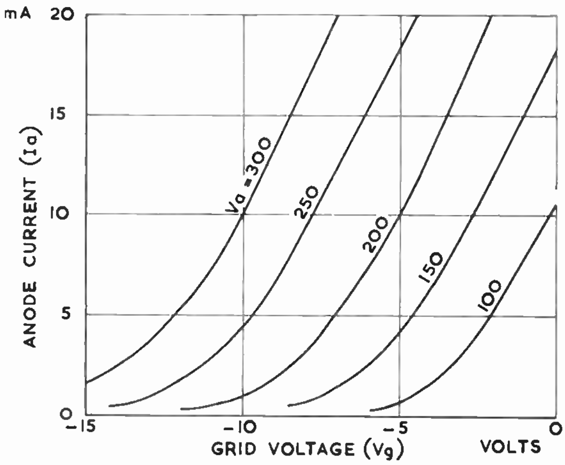


Fig. 2.2. Anode-current/grid-voltage characteristics of a triode.

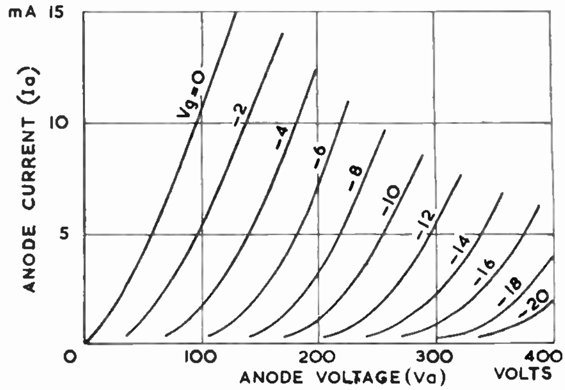


Fig. 2.3. Anode-current/anode-voltage characteristics of a triode

the change in anode current is plotted graphically, the resultant graph is known as a *characteristic curve*. A series of such curves may be plotted with different anode voltages: the set of curves is known as the anode-current/grid-voltage characteristics (I_a/V_g). A second series of curves may be plotted for anode current versus anode voltage for a series of different grid voltages. These are known as the anode-current/anode-voltage characteristics (I_a/V_a). Examples are shown in **Figs. 2.2 and 2.3.**

Tetrodes

A *tetrode* valve is a triode with an additional grid, known as the screen (g_2) interposed between the control grid and the anode. When this screen is maintained at a steady positive voltage, it is found that the amplification factor of the valve, as compared with the triode, is very much higher, the impedance is also greatly increased.

The reason for this increased amplification lies in the fact that the anode current in the tetrode valve is far less dependent on the anode voltage than it is in the triode. In any amplifier circuit, of course, the voltage on the anode must be expected to vary since the varying anode current produces a varying voltage-drop across the load in the anode circuit. A triode amplifier suffers from the disadvantage that when, for instance, the anode current begins to rise due to a positive half-cycle of grid voltage swing, the anode voltage falls (by an amount equal to the voltage developed across the load) and the effect of the reduction in anode voltage is to diminish the amount by which the anode current would otherwise increase. Conversely, when the grid voltage swings negatively the anode current falls and the anode voltage rises. Because of this increased anode voltage the anode current is not so low as it would have been if it were independent of anode voltage. This means that the full amplification of the triode cannot be achieved. The introduction of the screen grid, however, almost entirely eliminates the effect of the anode voltage on the anode current, and the amplification obtainable is thus much greater.

A screen is found to function best when its voltage is below the mean value of the anode voltage. Most of the electrons from the cathode are thereby accelerated towards the anode, but some of them are unavoidably caught by the screen. The resulting screen current serves no useful purpose, and if it becomes excessive it may cause overheating of the screen. The total cathode current is equal to the sum of the screen and anode currents.

Another important effect occurs when a screen grid is introduced into a triode: provided that the screen is kept at a constant voltage (not varying with the signal) it reduces the capacitive coupling between the control grid and the anode and therefore helps to eliminate unwanted feedback in amplifier circuits especially at radio frequency. To take full advantage of this feature the screen grid is made with a finer pitch or smaller mesh size than would be necessary merely to obtain greater amplification, and auxiliary electrostatic shields are built into the structure in an attempt to reduce the grid-to-anode capacitance to the lowest practicable value. If the size of the apertures in the screen is made too small the electron flow to the anode will be seriously impeded, but with a reasonable compromise the residual capacitance between control grid and anode can be made 1,000 times smaller than in a triode. The improved stability in a radio-frequency amplifier depends on the constancy of the screen voltage, and it is for this reason that thorough capacitive by-passing of the screen to earth (i.e. decoupling) is so important.

There is another type of tetrode, known as a *space-charge grid tetrode*, in which a positively charged grid is interposed between the control grid and the cathode. The purpose of this positive grid is to overcome the limiting effect of the negative space charge and thus enable the valve to operate efficiently with very low anode voltage (for example, a 12 volt supply as used for mobile equipment).

The tetrode suffers from the disadvantage that, when the anode voltage swing is so great that on downward peaks it falls below the screen voltage, there is a flow of secondary electrons from the anode to the screen. The effect of this secondary emission is to cause a drop in anode current and a rise in screen current, which in ordinary amplifier circuits results in serious distortion and a reduction in useful power output.

Pentodes

By introducing an additional grid between the screen and the anode and fixing its potential at some low value, usually cathode or earth potential, the flow of secondary electrons can be prevented. The valve can then be operated with much larger anode-voltage swings without distortion and without reduction in power output. This third grid is known as a *suppressor grid* (g_3) and a tetrode which has been modified in this way becomes a *pentode*.

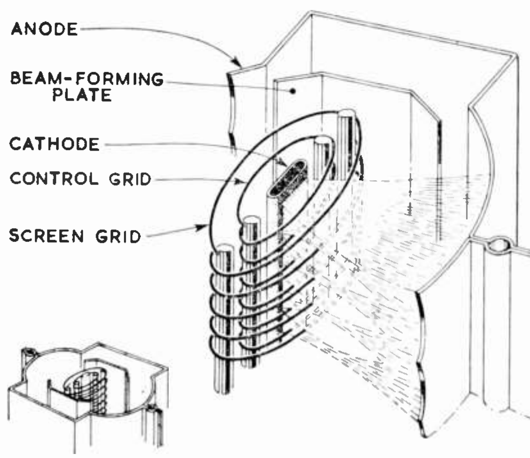


Fig. 2.4. Internal construction of a beam power tetrode valve. The drawing shows the focused streams of electrons between the aligned turns of the control grid and the screen grid, and also the beam-forming or confining plates which shield the anode from the electrons arriving from the regions of the support wires.

Other methods are also in use for suppressing secondary emission, for example, by increasing the separation between the screen and the anode until the secondary electrons have insufficient energy to reach the screen (as in the " Harries Critical Anode-Distance Valve ") or by introducing small fins projecting inwards from the anode. In another method a pair of deflecting plates is fitted near the anode and connected to cathode or earth. Valves having this form of construction are known as *beam tetrodes* or *kinkless tetrodes*. The term " kinkless " refers to the absence of the kink which appears in the characteristic curve of the simplest type of tetrode due to the flow of secondary emission electrons from anode to screen at low anode voltages.

Beam Tetrodes

A beam tetrode employs principles not found in other types of valves in that the electron stream from the cathode is focused towards the anode. The control grid and the screen grid are made with the same winding pitch and they are assembled in the valve so that the turns in each grid are in optical alignment: see **Fig. 2.4.** The effect of the grid and screen turns being in line is to reduce the screen current compared with a non-beam construction. For example, in a pentode of ordinary construction the screen current is about 20 per cent of the anode current, whereas in a beam valve the figure is 5–10 per cent.

The pair of plates for suppressing secondary emission referred to above are bent round so as to shield the anode from any electrons coming from the regions exposed to the influence of the grid support wires at points where the focusing of the electrons is imperfect. These plates are known as *beam-confining* or *beam-forming plates.*

Beam valves were originally developed for use as audio-frequency output valves, but the principle is now applied to many types of radio-frequency tetrodes both for receiving and transmitting. Their superiority over pentodes for a.f. output is due to the fact that the distortion is caused mainly by the second harmonic and only very slightly by the third harmonic, which is the converse of the result obtained with a pentode. Two such valves used in push-pull give a relatively large output with small harmonic distortion because the second harmonic tends to cancel out with push-pull connection. **Fig. 2.5** shows the characteristic curves of a beam valve and a pentode of equivalent size. The line *AB* is a load line drawn on the curve. It will be seen that the line extends farther to the left before it cuts the zero grid-volts curve than is the case with the pentode; this indicates a greater power output because the power output is proportional to the product of the change in anode voltage between points *A* and *B* on the horizontal axis and the change of anode current between points *A* and *B* on the vertical axis.

The widespread use of beam power valves as r.f. amplifiers and frequency doublers, etc. is referred to later in this chapter.

Gated Valves

It is convenient for many purposes to be able to control the output of a valve from two different sources. If sufficient negative voltage is applied to the suppressor grid of a pentode

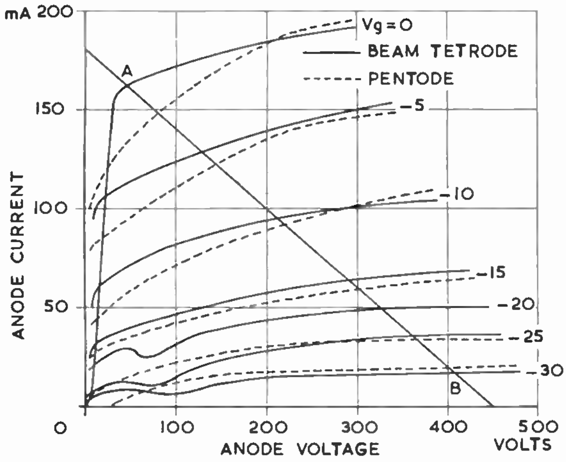


Fig. 2.5. Comparative characteristic curves of a pentode and a beam tetrode of equivalent sizes. AB is a load line.

valve the anode current can be cut off. The effect of cutting off the anode current is to cause a considerable increase in screen current; care must then be taken to avoid exceeding the safe screen dissipation. If the suppressor grid is biased negatively to cut off the anode current and if a positive pulse is then applied to it, an input fed to the control grid will affect the anode current only during the pulse, as if a gate were opened, the pulse being called a *gating pulse.* Because it is difficult to control accurately the magnitude of a pulse, it may happen that the suppressor grid is pulsed positively into the region where it will take current and " rob " the anode. To avoid such a possibility a diode is usually connected between this grid and cathode to act as a low-impedance shunt and thereby prevent a significant positive voltage. In certain valves specially intended for this purpose, a diode is built on the suppressor grid, which is then known as a *dioded suppressor grid*, and the pitch of the turns in the suppressor grid is made finer so that the anode current is cut off by a voltage of the same order as that of the control grid. Instead of a suppressor grid, specially designed beam plates may be employed (as in the beam tetrode) in conjunction with aligned control and screen grids, and the resultant valve is a *gated-beam valve*.

Examples of the use of these valves are television synchronizing circuits, radar, and volume compression and expansion.

VALVE CHARACTERISTICS

The performance characteristics of all valves as published by the makers are based on the assumption that the cathode is maintained at the proper operating temperature, which in most cases also has an important effect on the useful life of the valve. Under-running may be as detrimental as over-running, particularly in gas-discharge valves such as mercury-vapour rectifiers.

The characteristics are also related to the specified values of anode and grid voltages, and while it is possible and often convenient to operate a valve with anode and screen voltages below the rated values there will generally be a deterioration in performance but the life and reliability will be increased.

Mutual Conductance

The ratio of the change of anode current to the change of grid voltage while a constant anode voltage is applied is known as the *mutual conductance* or *slope* (g_m) of the valve and is expressed in milliamperes per volt. It is also sometimes expressed in micromhos, which is a unit of conductance (1 mA/V is equal to 1000 micromhos).

Amplification Factor and Impedance

If the grid is made more negative and the anode voltage raised, or if the grid is made less negative and the anode voltage lowered, it is possible to adjust the relative values so as to keep the anode current constant and at the same value in each case. The ratio of change of anode voltage to change of grid voltage for constant anode current is known as the *amplification factor* (μ). Valves are frequently classified as low-μ (less than 10), medium-μ (10–50) and high-μ (greater than 50).

If the anode voltage of a valve is altered and the grid voltage kept constant, the anode current will change. The ratio of change in anode voltage to change in anode current for constant grid voltage is known as the *impedance* (r_a), *a.c. resistance* or *slope resistance* of the valve. It can be considered as behaving in the same manner as the internal resistance of a battery or generator.

Impedance, mutual conductance and amplification factor of a valve are related by the equation—

$$\textit{Impedance (ohms)} = \frac{\textit{Amplification Factor}}{\textit{Mutual Conductance (mA/V)}} \times 1000$$

Mutual conductance and impedance are equal to the slopes of the I_a/V_g and of the V_a/I_a characteristics respectively.

Inner and Outer Amplification Factors

In a tetrode or pentode valve the screen grid (g_2) can be considered as an anode and in fact may be employed as part of the anode. For example, if the valve is used as a triode by connecting the anode and screen together, the amplification factor as defined above when measured between control grid (g_1) and the screen grid (g_2) is known as the *inner amplification factor* or *inner* μ, and this figure can be taken for practical purposes to be the μ of the valve when connected as a triode.

Similarly if the screen grid (g_2) is considered as a control grid in conjunction with the anode, the ratio of change of anode voltage to change of screen-grid voltage for a constant anode current is known as the *outer amplification factor* or *outer* μ.

Load Impedance

The resistance or load connected in the anode circuit of a valve is known as the *load impedance*. For triodes this has a value of 1–10 times the valve impedance (except in special cases), depending upon whether the valve is employed as a voltage amplifier or as a power amplifier. A voltage amplifier is one such as would be used as an intermediate stage in an audio amplifier and is usually resistance-coupled. A power amplifier would be used in an output stage feeding a loudspeaker or as an anode modulator in a transmitter.

Cut-off Tail: Variable-μ

When an increasing negative voltage is applied to the control grid of a valve the anode current falls, and the grid voltage at which the anode current falls to a specified low value is defined as the *cut-off* point; the measurement is called a *cut-off* or *tail* test. The reduction in anode current towards cut-off is accompanied by a corresponding reduction in mutual conductance. For some purposes such as radio frequency and intermediate frequency amplifiers having automatic gain control, the reduction in mutual conductance with increasing negative grid bias should be gradual rather than abrupt. This is achieved by employing a grid which either has some of its turns missing (i.e. in effect a hole) or is wound with a non-uniform pitch. Valves using such grids are known as *variable-μ* or remote cut-off valves. The cut-off in these valves is generally stated to be the grid voltage at which the mutual conductance falls to one hundredth of its normal operating value.

Variable-μ valves are classified in the following way in terms of the grid cut-off voltage, expressed as a percentage of the normal rated screen voltage:

Semi-remote cut-off .. .. 10–20 per cent
Remote cut-off .. .. Greater than 20 per cent

Sometimes valves are merely divided into two varieties—*short grid base* and *long grid base*—corresponding to the relative grid voltage required for cut-off, the name originating from the appearance of the I_a/V_g characteristics.

Voltage Gain

In a voltage amplifier, the gain per stage is proportional to the ratio of the external load impedance to the valve and load impedances because these impedances act as a potential divider. This can be expressed as:

$$\text{Voltage Gain} = \frac{\text{Amplification Factor} \times \text{External Load Impedance}}{\text{External Load Impedance} + \text{Valve Impedance}}$$

$$= \mu \times \frac{R}{R + r_a}$$

For example, a valve having an amplification factor of 50 and an impedance of 50,000 ohms working into an external load of 50,000 ohms will give a stage gain of:

$$50 \times \frac{50{,}000}{50{,}000 + 50{,}000} = 25$$

That is to say, the potential division between the equal impedances has halved the effective amplification factor. If a valve having a higher mutual conductance (i.e. for the same amplification factor a lower impedance) is chosen, the stage gain will be increased.

In the case of an amplifier where the load impedance is very low compared with the valve impedance (such as in a television receiver), the gain can be obtained approximately from the expression:

Voltage Gain = Mutual Conductance × Load Impedance

This is derived from the fact that in the earlier formula the term " amplification factor " can be replaced by " mutual conductance × valve impedance " and the denominator can be written as " valve impedance " since the load impedance is small in comparison.

Anode Dissipation

The power expended in the anode of a valve and liberated as heat is known as the *anode dissipation*. The term must not be confused with either power output or power input.

In any valve the anode dissipation is the remaining power in the valve after the useful power output has been deducted from the power input, i.e. the h.t. supply to the anode. In amplifiers operating in class A, the anode dissipation can be considered approximately as the input power, since the output is relatively small but for oscillators or class B or C amplifiers the dissipation may be only 40 per cent of the input power, depending upon the efficiency obtained.

Anode dissipation in excess of the rated maximum will raise the anode to a temperature above its safe limit. This results in the release of occluded gas which will poison the cathode and reduce its emission; it can also cause anode primary emission or even melt the anode.

Screen Dissipation

In a multi-grid valve such as a pentode an appreciable amount of power is dissipated in the screen grid and this is known as *screen dissipation*. It is of considerable importance in transmitting valves and line-scan valves for television.

Grid Dissipation

When power is dissipated in a grid other than the screen grid, it is referred to as *grid dissipation*. It requires special attention in class C amplifiers and space-charge grid tetrodes.

Optimum Load

The performance of a power amplifier is dependent on the value of the load in the anode circuit. For a given set of operating conditions it has an optimum value giving reasonable power output consistent with low harmonic distortion. This *optimum load* is usually quoted in valve data sheets in conjunction with the other circuit conditions. For push-pull operation, the optimum *anode-to-anode* load is quoted and has a value rather lower than twice that for a single valve. This is particularly true of pentode valves (as distinct from triodes and tetrodes) because the load in a single ended pentode stage is usually chosen for minimum *second* harmonic whereas that for a pair of valves in push-pull is chosen for minimum *third* harmonic.

If valves are operated in class B, the optimum load will differ from that for operation in class A.

Power Output

Power output is always a relative term and it is often ambiguous; it therefore needs to be carefully defined. In the case of an amplifier the harmonic content of the output must be stated, and in the case of oscillators the circuit conditions and frequency must be known. For fully driven class A amplifiers the approximate power output may be calculated by multiplying the r.m.s. input voltage (0·707 × grid bias) by the mutual conductance at the operating point (converted to amperes per volt), squaring the result and multiplying this by the output load impedance. For example, if a valve has a grid bias of −16 volts and a mutual conductance of 2·5 mA/V and is working into an output load of 6000 ohms, the maximum power output would be:

$$\left(0{\cdot}707 \times 16 \times \frac{2{\cdot}5}{1000}\right)^2 \times 6000 = 4{\cdot}8 \text{ watts}$$

It should be borne in mind that this formula takes no account of distortion. Since it is not possible in practice to drive the grid much nearer the positive region than −1 volt without causing grid current to flow, it is necessary to deduct 1 volt from the grid bias in order to obtain the maximum permissible grid-voltage swing. Furthermore, this formula assumes that the output load is low compared with the valve impedance.

Load Lines

Load lines are lines drawn on the characteristic curves of a valve from which can be calculated the performance under specified circuit conditions. A load line drawn through the operating point on the anode-current/anode voltage curves of an output valve having a slope equal to the output load can be used to determine the useful output power and the harmonic distortion. For example, in **Fig. 2.5** the line *AB* represents a load impedance given by:

$$\frac{450 \text{ volts}}{180 \text{ milliamperes}} = 2500 \text{ ohms}$$

Input Impedance and Input Admittance

When the grid and cathode of a valve are connected across a resonant circuit, the performance of this circuit is modified; this modification results from the input of a valve having resistive and reactive components effectively in parallel with the circuit. The resistive component is due to leakage, to dielectric losses in the valve base and grid insulators, and to electron transit-time phenomena, internal lead impedances and other factors. As the frequency is increased this resistive component decreases in value at a rate approximately proportional to the square of the frequency: it also varies inversely as the slope (g_m) and is known as *input impedance* sometimes quoted as *input admittance*.

The reactive component is due to the capacitance between the input electrode and the other electrodes and also includes the capacitances between the leads and the pins in the base. The values quoted by manufacturers are sometimes given as *input capacitance* and sometimes as an electrode capacitance to all other parts. They are generally measured with the valve cold, and the figure obtained when the valve is working normally will be higher, the value then depending upon the circuit and the operating conditions of the valve.

The input capacitance can be further increased by a factor depending on the nature of the output load. Since the signal voltage applied to the grid is opposite in phase to the resultant signal anode voltage, a circulatory current flows through the grid-to-anode capacitance. The input signal has to furnish this capacitative current, and therefore "sees" an effective capacitance additional to the reactive component mentioned above. The higher the anode load, the greater the voltage amplification and the higher this capacitance becomes. The effect is known as the *Miller effect* and can be avoided when essential by neutralization. The extra capacitance in a resistance-coupled amplifier is equal to $C_{(g\text{-}a)}\,A$ where A is the voltage amplification of the stage.

Equivalent Noise Resistance

If electrons were infinitesimally small and infinitely numerous, it could be imagined that they would reach the anode in a perfectly steady stream. However, because the electron emission consists of a flow of discrete electrical charges the rate at which the current arrives at the anode shows slight but appreciable fluctuation. This random fluctuation in anode current is known as *shot noise* and is so called because when amplified it resembles the noise produced by a hail of shot falling on a metal surface. Although the noise is developed (i.e. first appears) in the anode circuit it can be more conveniently regarded as the noise which would be generated if an equivalent noise voltage were applied to the grid. For the purposes of comparison with the noise generated by thermal agitation in a resistor (known as *Johnson noise*) it is convenient to express the shot noise generated by a valve also in terms of a resistor: this is an imaginary resistor which if connected in series with the grid would generate the same amount of noise as that produced by the valve and is termed *equivalent noise resistance* (R_{eq}). The value of R_{eq} may be calculated as follows:

For a triode used as an amplifier:

$$R_{eq} = \frac{2{\cdot}5}{g_m} \text{ kilohms}$$

For a pentode used as an amplifier—

$$R_{eq} = \frac{I_a}{I_a + I_{g2}}\left(\frac{2{\cdot}5}{g_m} + \frac{20\, I_{g2}}{g_m}\right) \text{ kilohms}$$

In these formulae the currents are expressed in milliamperes and the mutual conductance is expressed in milliamperes per volt, all being measured at the operating point.

The noise generated by a multi-grid valve (e.g. a pentode) is much higher than that generated by a triode owing to the partition effect of the anode and screen currents. Although not all of the noise originating at the cathode is developed in the anode circuit (since not all of the cathode current reaches the anode) there is a large partition-noise contribution caused by the element of chance as to whether any given electron goes to the anode and then through the anode circuit or to the screen and *not* through the anode circuit.

Since the value of R_{eq} is determined by the valve characteristics, which do not alter with frequency over its normal working range, it can usually be taken that the equivalent noise resistance is also independent of frequency.

Induced Grid Noise

The random variations in the electron stream as it passes the grid on its way to the anode induce a noise voltage in the grid circuit by reason of the electro-static charges carried by the electrons. This voltage depends on the nature of the impedance in the grid circuit and on the frequency at which the circuit is operating. At low and medium frequencies the effect is negligible because the voltage induced by the electrons approaching the grid is neutralized by that induced by the receding electrons, whereas at very high frequencies the time of transit of the electrons corresponds to an appreciable fraction of a cycle and this leads to a phase difference between the approaching and the receding electrons. The resultant induced grid current has a component which can be more or less taken into account by supposing the connection of an appropriate resistance across the grid circuit. This resistance acts as a generator of thermal noise at an equivalent temperature to that of the valve cathode, i.e. about five times the absolute room temperature; the noise so generated is known as *induced grid noise*. Because it appears as a shunt to the grid circuit this resistance also results in a reduction of the input impedance; it is usually expressed as a conductance in micromhos and is known as the *transit-time conductance*, the value increasing approximately as the square of the frequency.

Noise Factor

The effects described make necessary the evaluation of the performance of a valve by taking into account the frequency of operation. At frequencies above 20 Mc/s, the *noise factor* must be measured. This can be done with the aid of a thermionic noise diode up to 400 Mc/s or a gas noise source for higher frequencies.

The noise factor may be calculated from the formula:

$$\text{Noise Factor } F = \frac{e}{2KT} \quad I_d R_s.$$

where e = electron charge $1{\cdot}59 \times 10^{-19}$ coulomb
K = Boltzman's constant $1{\cdot}372 \times 10^{-23}$ joules per °K
T = Temperature of source °K
I_d = Noise diode current in amps required to double the receiver noise output
R_s = Source resistance in ohms

At normal room temperature the formula may be simplified to:

$$F(db) = 10_{log}(20\, I_a\, R_s).$$

Hiss

The noise generated by a valve when used as the first stage of a high-gain amplifier, usually at audio frequency, is known as *hiss*. This noise includes shot noise and that due to various inter-electrode leakages but it does not include hum. It is generally expressed as an equivalent noise voltage in microvolts applied to the control grid.

Hum

Valve hum is the component of the anode current resulting from modulation of the electron stream due to the use of an a.c. filament or heater supply. When a cathode is heated by a.c. the current generates a magnetic field which can modulate the electron stream, and a modulating voltage is injected into the control grid through the inter-electrode capacitance and leakages: additionally there can be emission from the heater in an indirectly-heated valve. The hum is usually expressed as an equivalent voltage (in microvolts) applied to the control grid. Valve hum should not be confused with hum generated in other circuit components.

Primary Emission

In an indirectly-heated valve it is possible for the tungsten heater to become contaminated with emissive material and so emit electrons. This is known as *heater emission*. The emitted electrons will be attracted to any electrode which is positive with respect to the heater, and such an electrode may be the cathode or the screen grid or anode; it can even be the control grid, which although generally negative to the cathode can become sufficiently positive to one end of the heater during at least part of the a.c. cycle of the heater supply. This explains why it is sometimes recommended that the centre-tap of the heater supply should be connected to a positive point in the circuit to reduce hum. The grid can thereby be maintained at a negative potential with respect to the heater.

Grid primary emission is a condition in which a grid commences to emit electrons itself and, figuratively speaking, competes with the cathode. The effect is produced by the heating of the grid which may be caused by an excessive flow of grid current or by the close proximity of the hot cathode or by radiated heat from the anode. The effects are accentuated if the grid becomes contaminated with active cathode material, which can happen if the valve is appreciably over-run even for a relatively short time.

The control grid, the screen and the suppressor grid are all subject to these effects. They are avoided by keeping the grid-cathode resistance low and by avoiding excessive heater, anode or bulb temperature (i.e. by not over-running the valve).

Two examples of the effects of control-grid primary emission can be given:

(a) In a small output valve the anode current rises steadily accompanied by distortion due to grid current flowing in the high-resistance grid leak in such a direction as to oppose the grid bias.
(b) In a power amplifier or frequency multiplier in a telegraphy transmitter the drive diminishes when the key is held down, accompanied by rising anode current.

A consequence of primary screen emission in an oscillator or amplifier is that when the screen voltage is removed, for example by keying, the output is maintained often for quite long periods if the valve is already hot.

Anode primary emission is similar to grid emission but occurs when the anode attains a sufficiently high temperature to emit electrons. This effect occurs mostly in rectifiers and causes breakdown between anode and cathode.

Secondary Emission

When an electron which has been accelerated to a high velocity hits an electrode such as a grid or anode, electrons are dislodged and these electrons can be attracted to any other electrode having a higher potential. This effect is termed *secondary emission*. Under controlled conditions one electron can dislodge several secondary electrons, and a series of secondary-emitting "cathodes" will give a considerable gain in electrons. This principle is used in the electron-multiplier type of valve.

Conversion Conductance

The term *conversion conductance* is used in regard to detectors or frequency changers to represent the ratio of the output current of one frequency to the input voltage of another frequency. As applied to the mixer of a superheterodyne receiver, for example, the conversion conductance is the current in the anode circuit at intermediate frequency (measured in microamperes) divided by the input voltage to the grid at signal frequency. The symbol commonly used is g_c and it is generally measured in microamperes per volt.

Conversion or Translation Gain

Conversion or translation gain is the ratio of intermediate-frequency output voltage to radio-frequency input voltage. It can be obtained from the conversion conductance if the dynamic resistance and other parameters of the i.f. transformer used in the mixer anode circuit are known. Both terms relate to the efficiency of a mixer but the *conversion gain* takes into account the output circuit while the *conversion conductance* refers only to the valve and its operating conditions.

Cathode-interface Impedance

When a valve is operated for long periods particularly with low cathode current or at complete cut-off, the mutual conductance steadily falls and so also does the available peak emission. This effect is due to the growth of a film between the metallic cathode and its emissive coating. This film possesses an impedance—the *cathode interface impedance*—which may be represented by a resistance with capacitive shunt connected in series with the cathode and acting as an automatic bias resistor. The falling performance is sometimes colloquially termed *sleepy sickness*. The rate of growth of interface resistance is considerably affected by the material of the cathode and is accelerated by high temperatures resulting from excessive heater voltage. Since the cathode-interface resistance is normally of the order of a few hundred ohms it has a most serious effect on valves having a high slope and a short grid base because the normal cathode resistor is likely to be comparable with this value. The effect of the parallel capacitance is to make the drop in performance less noticeable as the frequency is increased.

Contact Potential

A small potential difference exists between electrodes of dissimilar materials in a valve irrespective of any externally applied potentials. This is known as the *contact potential*. In a simple diode there is a potential difference between the anode and the cathode which causes a current to flow in any external circuit from anode to cathode. The magnitude of this contact potential depends on the cathode material, the type of emissive coating, the anode material and any contaminating film present upon its surface. Its value (anode to cathode) is between +1 volt and −0·5 volt, but is most frequently positive; it is affected by cathode temperature and varies throughout the life of the valve. In a triode or any other valve with a control grid a potential difference exists similarly between the control grid and cathode and between other electrodes. All electrodes except the control grid and possibly the suppressor grid can be ignored in practical applications, because the current due to it is small compared with other currents flowing in the circuit. The contact potential is effectively in series with the control grid, with the result that if no external grid bias is applied grid current will flow in the external circuit in the same manner as in a diode. If an increasing external grid bias is applied, a value of negative bias is reached when the grid current ceases and this is a measure of the contact potential for this particular valve. In order to operate the valve satisfactorily it is necessary in most cases to increase the bias still further to a point where the maximum positive signal input does not swing the grid more positive than the contact potential.

USES OF VALVES

The Valve as an Amplifier

When a valve has an impedance connected in series with its anode supply and the voltage on its grid is varied, the resultant change in anode current will cause a voltage change across the impedance. This impedance may be a resistance, an inductance, or in some cases it may behave like a capacitance, as for example in a circuit resonant at a frequency below the operating frequency.

Class A Operation

The curve in Fig. 2.6 shows graphically the operation of a valve working as an amplifier. This curve, which is known as

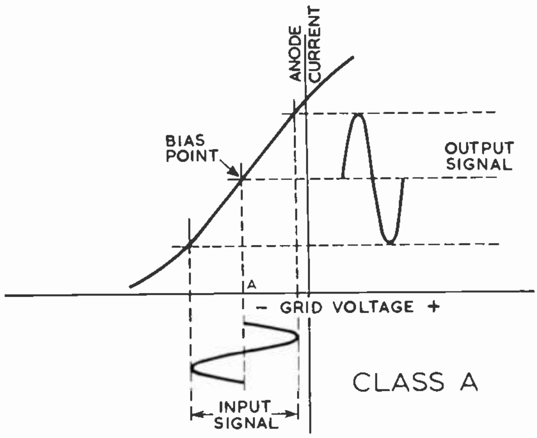


Fig. 2.6. A triode valve operating as a class A amplifier.

the *characteristic curve* of the valve, shows how the anode current varies with grid voltage. If the grid bias is fixed at the point A, which corresponds to the centre of the useful straight portion of the curve (i.e. where the grid potential remains negative), any complex alternating voltage applied to the grid will be reproduced as a similar complex wave of increased amplitude in the anode circuit. Since the travel of the operating condition is up and down the straight portion of the curve, the output wave shape will be exactly similar to that of the input and no distortion will result, but if the bias is adjusted so that the valve operates over a curved portion of the characteristic, there will be a flattening of the corresponding half-cycle and distortion will result. The method of operation where the grid bias is fixed at the centre of the straight portion of the curve is known as *class A operation*.

The amount of distortion produced by a non-linear amplifier is usually expressed in terms of the harmonics which are generated by it. If the input has a complex waveform, the analysis of the distortion is very difficult and it is much more convenient to consider the distortion of a simple sine-wave input. When a sinusoidal input is applied to an amplifier operating over a curved portion of its characteristic, the distorted output is found to contain not only the amplified fundamental but also a number of harmonics (all of which are of course sinusoidal). This can be deduced mathematically by what is known as Fourier Analysis. The respective amplitude of each harmonic expressed as a percentage of the fundamental is known as the *percentage harmonic content*. Taken collectively the total effect can be calculated by adding together all the individual harmonic components. Since the power represented by the oscillatory wave is a function of the square of the voltage (or the current) the effective *total harmonic content* of the distorted wave will be given by the square root of the sum of the squares of the voltage (or current) amplitudes of the respective harmonic components and not by simple addition.

Distortion of a similar kind occurs when the input signal has a complex waveform. In just the same way as the distorted output of the original sinusoidal signal referred to in the previous paragraph can be resolved into a number of components comprising the fundamental and a series of

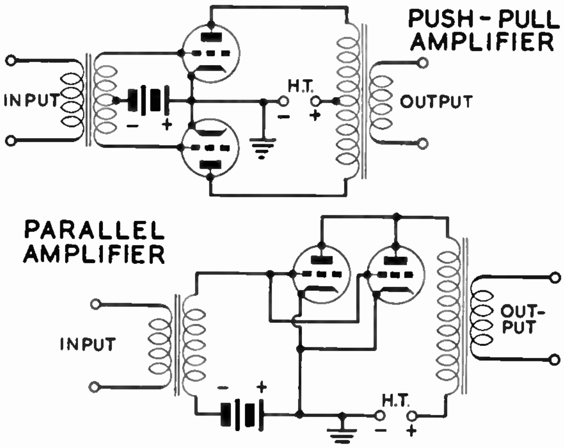


Fig. 2.7. Push-pull and parallel amplifier circuits.

harmonics, so can a complex input signal itself be resolved into a fundamental and a series of harmonics. If such a complex input signal is fed into a non-linear amplifier each of the sinusoidal components comprised in it will be subject to distortion by the curvature of the characteristic and will appear in the output as a series of harmonically related components. Cross-modulation will also occur and add considerably to the total distortion. The complete analysis of the distortion of a complex wave would need to take into account the relative phases of its harmonic components, but fortunately this is unnecessary in the design of communications equipment.

Push-pull and Parallel Amplifiers

When it is necessary to obtain more output than can be provided by one valve, two or more may be connected in push-pull or in parallel. Fig. 2.7 shows the connections for these two methods. The principle used in the push-pull arrangements is that whenever the grid of one valve is going more positive the other is going more negative, and in consequence the anode current of one valve is rising as the other is falling; hence the name *push-pull.*

In parallel operation, the grids are connected together and the anodes are connected together with the result that the output is increased in proportion to the number of valves used. In regard to distortion, the relative harmonic content is the same as for one valve alone, whereas in the push-pull system all even harmonics (second, fourth, sixth, etc.) are cancelled out, due to the method of operation. It follows therefore that two valves connected in push-pull will give a considerably greater output for the same degree of distortion than twice that obtained from "single-ended" operation, i.e. from a single valve.

Class AB1 Operation

As mentioned above, when valves are operated in push-pull some of the distortion produced by curvature of the characteristic is cancelled out. This is taken advantage of in *class AB1* operation by employing a negative grid bias slightly greater than that for class A operation and allowing the signal input voltage to swing over the entire curve up to the point where grid current starts. A large part of the distortion which is thereby introduced is automatically cancelled by the push-pull mode of operation.

Class AB2 Operation

If class AB1 operation is carried a stage further and the negative grid bias is increased to a value where the anode current is quite low and if the signal input is allowed to increase so that the grid voltage swings into the positive grid region, the power output is further increased. Class AB2 operation is intermediate between class AB1 and class B and is frequently employed for large a.f. power amplifiers. See Fig. 2.8.

Class B Operation

It has already been explained how a valve must be operated over the straight portion of the characteristic if distortion is to be avoided in single-ended operation. The straight portion of a valve curve is only a small part of the whole, and consequently if it were possible to use nearly all of the curve without introducing distortion, much more output would be available. This can be achieved in practice by a method known as *class B* operation, in which the valve is operated at the cut-off point on the curve. This is shown in Fig. 2.9, from which it will be evident that the negative half-cycle is almost completely suppressed. If the valve is used as an r.f. amplifier, this is unimportant since the oscillatory characteristic of the tuned circuit will restore the other half-cycle and remove the harmonic distortion, but for a.f. amplification the method if considered in this simple form is impossible. If, however, two such valves are used in push-pull, each valve will supply the missing half-wave of the other and a normal full-wave will result in the output. Because the mean anode current depends upon the signal voltage, the demand upon the h.t. supply for anode current will fluctuate with the signal amplitude. Hence a power

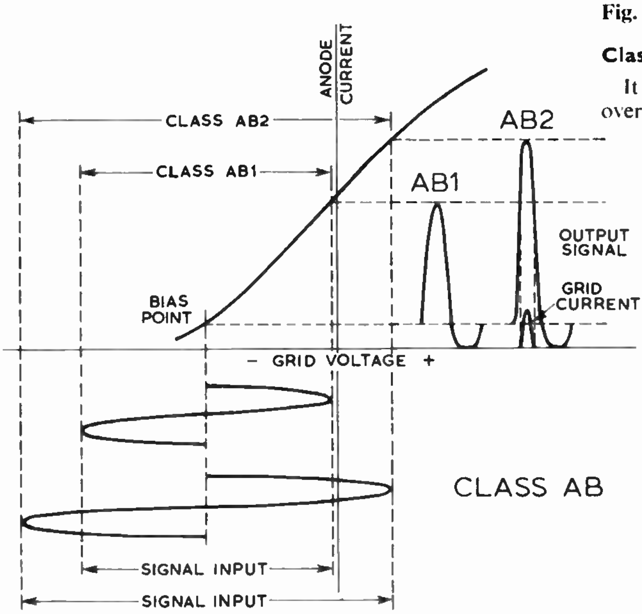


Fig. 2.8. Class AB operation showing how classes AB1 and AB2 differ.

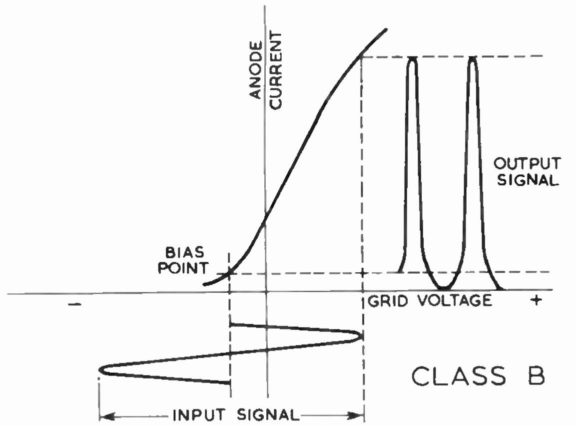


Fig. 2.9. In a class B amplifier the valve is operated at the cut-off point.

supply with good voltage regulation is essential. Moreover since the grid is driven positive, appreciable grid current will flow and good regulation in the grid-bias supply is also necessary. Since the grid takes current it consumes power, and this must come from the signal input. The preceding or driver stage must furnish this power adequately without distortion.

Class C Operation

The class B principle can be carried a step further to class *C* operation by using the entire valve characteristic. Here the grid is biased to approximately twice the cut-off voltage and is driven far into the positive region in order to reach saturation point, as is shown in **Fig. 2.10.** Since the grid is driven appreciably positive, considerable grid current flows and the anode is robbed of current at the peak of each cycle, thus causing the anode current pulse to be flattened at the top. As the waveform is poor and the distortion very considerable, class C operation is almost entirely restricted to r.f. amplification where high efficiency is very desirable and distortion relatively unimportant. Because of the high values

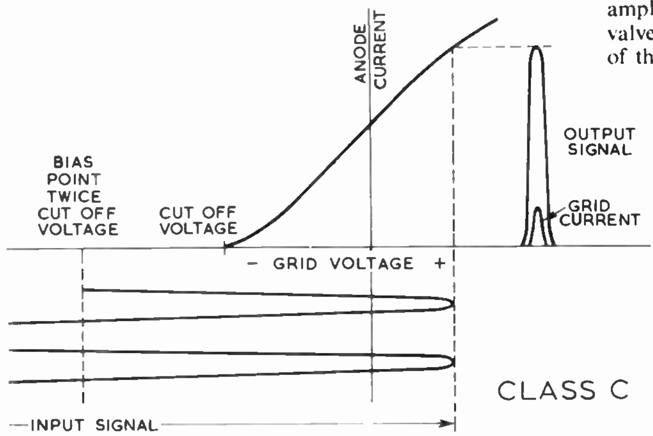


Fig. 2.10. In class C operation the valve is biased to approximately twice the cut-off voltage.

of grid current, considerable driving power is necessary.

Normally, the part of the positive half cycle during which current is drawn is about 140°. This angle of flow may be reduced to increase the output efficiency. Pulse operation is effectively super class C in which the angle of flow is very short.

Ultra-linear Operation

When a pentode or tetrode valve is operated normally the screen grid is decoupled with a capacitor of low reactance to cathode or chassis so that its potential may remain constant at all operating frequencies. If the screen grid were connected to the anode, the valve would perform as a triode: if, however, the screen grid is connected to a part of the output circuit intermediate between anode and h.t., such as a tap on the output transformer, a form of operation intermediate between that of a triode and that of a pentode results. In effect a proportionate amount of negative feedback is applied to the screen grid which lowers the gain but reduces the distortion. When the screen grids of two pentodes

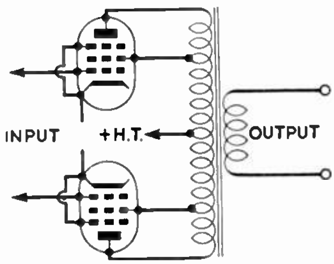


Fig. 2.11. Output circuit connections in a push-pull ultra-linear amplifier.

or tetrodes employed in push-pull class AB1 are connected to taps on the output transformer, the mode of operation is called *ultra-linear;* see **Fig. 2.11.** The tap positions are usually specified as a percentage of the total number of turns, the figure for optimum performance varying for different valve types.

Neutralized Amplifiers

Earlier mention has been made of the instability of amplifiers resulting from the feedback from the anode of a valve to the grid through the grid-to-anode capacitance and of the possibility of reducing this feedback by using an r.f. type of tetrode or pentode. At high frequencies, particularly if the grid and/or anode circuit has high dynamic resistance, this capacitance may still be too large for complete stability. A solution is to employ a circuit in which there is feedback in opposite phase from the anode circuit to the grid so that the effect of this capacitance is balanced out. The circuit is then said to be *neutralized.*

A typical arrangement is shown in **Fig. 2.12.** Here the anode coil is centre-tapped in order to produce a voltage at the " free " end which is equal and opposite in phase to that at the anode end. If the free end is connected to the grid by a capacitor (C_n) having a value equal to that of the valve grid-to-anode capacitance ($C_{g\text{-}a}$) shown dotted, any current flowing through $C_{g\text{-}a}$ will be exactly balanced by that through C_n. This is an idealized

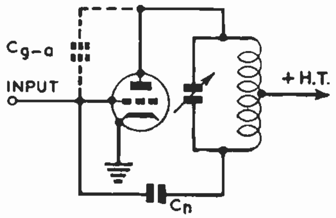


Fig. 2.12. Neutralizing a grounded-cathode triode amplifier. The circuit is equally suitable for a tetrode or a pentode.

case because the anode tuned circuit is loaded with the valve anode impedance at one end but not at the other; also the power factor of C_n will not necessarily be equal to that of $C_{g\text{-}a}$. The importance of accurate neutralization in transmitter power-amplifier circuits cannot be overstressed.

Grid Driving Power

An important consideration in the design of class B or class C r.f. power amplifiers is the provision of adequate driving power. The driving power dissipated in the grid-cathode circuit and in the resistance of the bias circuit is normally quoted in valve manufacturers' data. These figures frequently do not include the power lost in the valveholder and in components and wiring or the valve losses due to electron transit-time phenomena, internal lead impedances and other factors. Where an overall figure is quoted, it is given as *driver power output*. If this overall figure is not quoted, it can be taken that at frequencies up to about 30 Mc/s the figure given should be multiplied by two, but at higher frequencies electron transit-time losses increase so rapidly that it is often necessary to use a driver stage capable of supplying 3–10 times the driving power shown in the published data. The driving power available for a class C amplifier or frequency multiplier should be sufficient to permit saturation of the driven valve; i.e. a substantial increase or decrease in driving power should produce no appreciable change in the output of the driven stage. This is particularly important when the driven stage is anode-modulated. Care must be exercised, however, to ensure that the maximum grid current or other ratings of the valve are not exceeded.

Grounded Cathode

Most valves are used with the cathode connected to chassis or earth or where a cathode-bias resistor is employed it is shunted with a capacitor of low reactance at the lowest signal frequency used so that the cathode is effectively earthed.

Grounded Grid

Although a triode must be neutralized to avoid instability when it is used as an r.f. amplifier this is not always essential if an r.f. type of tetrode or pentode is employed. However, at very high frequencies (above about 100 Mc/s) a triode gives better performance than a tetrode or pentode, providing that the inherent instability can be overcome. One way of achieving this is to earth the grid instead of the cathode so that the grid acts as an r.f. screen between cathode and anode, the input being applied to the cathode. The capacitance tending to make the circuit unstable is then that between cathode and anode, which is much smaller than the grid-to-anode capacitance.

The input impedance of a grounded grid stage is normally of the order of 100 ohms and therefore appreciable grid input power is required. Since the input circuit is common to the anode-cathode circuit, much of this power is, however, transferred directly to the output circuit, i.e., not all of the driving power is lost.

Grounded Anode

For some purposes it is desirable to apply the input to the grid and to connect the load in the cathode circuit, the anode being decoupled to chassis or earth through a low-reactance capacitor. Such circuits are often referred to as grounded-anode circuits and are employed in cathode followers and infinite-impedance detectors.

Voltage Amplifiers

When it is desired to use a valve primarily for the purpose of increasing the signal voltage (as in an a.f. or i.f. amplifier) as distinct from increasing the power (as in an a.f. or r.f. output stage), there are four possible methods. These are known as:

(a) *resistance-capacitance coupling*
(b) *choke-capacitance coupling*
(c) *tuned-anode* (or *tuned-grid*) *coupling*
(d) *transformer coupling*

In general, the first method is suitable only for audio frequencies unless the gain per stage is very low and the output voltage required fairly small. The second is suitable for audio frequencies and for moderately high radio frequencies, provided that the choke is suitably chosen. The last two may be used up to very high frequencies, the limitation being the shunting capacitances in the tuned circuit or the transformer and in the valve itself.

Resistance-capacitance Coupled Amplifier

The basic circuit for a resistance-capacitance coupled amplifier is shown in **Fig. 2.13 (a)** for a triode and in **Fig. 2.13 (b)** for a tetrode or pentode. Although the essential features are the same, the values of the circuit components will depend on which type of valve is used.

In designing an amplifier using this form of coupling the choice of valve should be determined by:

(a) the h.t. line voltage available,
(b) the required stage gain,
(c) the required output voltage of the stage in question,
(d) the grid-leak value of the succeeding valve,
(e) the highest frequency to be amplified.

The h.t. line voltage will usually be fixed by the design of the remainder of the equipment, but for a.f. amplifiers as a general guide this voltage must not be less than about four times the required r.m.s. output voltage and preferably not less than six times. The actual anode voltage will of course be less than the h.t. line voltage by an amount equal to the voltage-drop in the anode resistor R, and provided that the anode voltage at zero anode current does not exceed the normal maximum rating, all will be in order. The values of the resistor and the mean anode current should be so chosen that the actual anode voltage is about 40 per cent of the h.t. line voltage. The grid bias should be such that with this anode current the valve will operate on the straight portion of its characteristic and will never be driven into grid current; hence the bias must always be greater than about -1 volt.

When a tetrode or pentode is employed the screen voltage will be such as to provide the correct anode current at a suitable grid bias, the voltage being generally 10-30 per cent of the h.t. line voltage. A triode provides less stage gain than a pentode but is capable of handling larger input voltages, and the succeeding grid leak R_L into which it delivers its output can be of lower value than in the case of a tetrode or pentode. This may be important because the grid leak is effectively in parallel with the anode resistor as regards the signal and hence it lowers the stage gain. The optimum value of grid leak is dependent on the size and type of valve employed. Output stages should not as a rule have grid leaks higher than 100,000 ohms (unless they are used with automatic bias in which case up to 500,000 ohms may be used), whereas small valves may use a Megohm or higher. The coupling capacitor C to the following valve should have a reactance of not more than one-half to one-quarter of the resistance of the grid leak at the lowest frequency to be used. Similarly the decoupling capacitors for the cathode and screen grid must have a low reactance compared with the associated resistors.

The high-frequency performance depends upon the output capacitance of the valve, the input capacitance of the succeeding valve, the valveholders and the wiring capacitance. All of these appear as shunt reactances across the anode resistor (which in effect is itself in parallel with the anode impedance of the valve) and will lower the resultant output load impedance and thereby reduce the gain. For audio frequencies the value of the anode resistor should not exceed 1 Megohm and should preferably be less. The effect of the shunt capacitance is more detrimental when the valve is a pentode than when it is a triode because the anode impedance of a pentode is much higher. In addition the higher output impedance of a pentode makes it unsuitable for driving an output stage that is operated in any condition other than strictly class A. These factors indicate that pentodes (or tetrodes) are more suitable for the early stages of a multi-stage amplifier, but it is well to bear in mind that they are inherently more noisy than triodes.

When a good h.f. response is essential, as in wide-band video amplifiers, it is usual to employ high-slope pentodes with very low values of anode resistor and grid leak, so that the shunting effect of the capacitance is reduced. With low

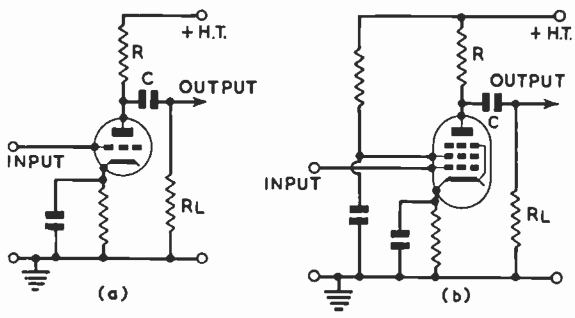


Fig. 2.13. Basic resistance-capacitance coupled amplifier: (a) triode, (b) tetrode or pentode.

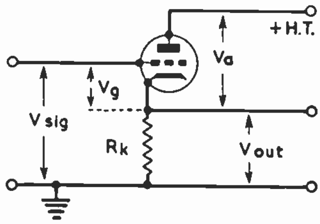


Fig. 2.14. Cathode follower.

values of anode resistor (for example, 4700 ohms) the anode voltage is nearly equal to the h.t. line voltage and the valve is then operated nearly at the makers' class A ratings. The stage gain and output voltage are small, however, due to the low value of the resistors employed.

Cathode Followers

If a valve is operated with the output load connected to the cathode instead of to the anode, the load is common to both grid and anode circuits and the arrangement is called a *cathode follower*: see Fig. 2.14. Here the input voltage V_{sig} is applied between grid and earth while the output load R_k, across which the output voltage V_{out} is obtained, is connected between cathode and earth. If at any instant V_{sig} is made more positive, the grid potential will become more positive and the anode current will increase: hence the cathode current will increase and the increased voltage-drop across R_k will cause V_{out} to increase in a positive direction also. Therefore the output voltage is in phase with the applied grid voltage: i.e. it " follows " the input voltage.

The amplification of a valve is concerned with the applied voltage between grid and cathode, and in the cathode follower this voltage V_g is not V_{sig} but $V_{sig} - V_{out}$. The higher the amplification of the valve the nearer V_{out} approaches V_{sig}. The voltage gain of the stage is—

$$\frac{V_{out}}{V_{sig}} = \frac{\mu\, R_k}{r_a + R_k\,(1 + \mu)}$$

where μ is the amplification factor of the valve and r_a is the anode impedance. It is evident that the stage gain must always be less than unity, although when μ is large and R_k is large compared with the anode impedance r_a the gain is nearly unity. Because there is no actual gain in voltage amplitude, the cathode follower can hardly be regarded as an amplifier. However, the circuit has certain advantageous properties, including a low output impedance (from about 50 to several hundred ohms), a high input impedance, low distortion (owing to the high degree of negative feedback through R_k) and negligible phase shift. One of its most important applications is as an impedance transformer where a very wide frequency range is required.

The load in the cathode lead may be a resistor, a tuned circuit or a loaded transformer, but if a tuned circuit or transformer is employed and if it is operated at frequencies off resonance the impedance will fall and the circuit will cease to act as a cathode follower.

For a given valve the optimum load for minimum distortion is the same irrespective of whether the load is connected in the anode or the cathode lead, but in the latter case, i.e. in the cathode follower, the distortion and output impedance are lower than in the former. It is also convenient at times to divide the load so that part of the load is in the anode and part in the cathode circuit. This arrangement gives a greater gain than the true cathode follower and results in less distortion and lower output impedance than the conventional amplifier arrangement.

If a tetrode or a pentode were used as a cathode follower it would behave as a triode because its anode, like the screen, would be at a constant potential. The relevant values of μ and r_a (anode impedance) are therefore those quoted for the valves connected as triodes.

Anode Followers

When a voltage amplifier with a uniform gain over a wide frequency range is required and when the gain is to be independent of changes in supply voltages, valve replacements, etc., it is convenient to use the *anode follower* principle. The anode follower circuit is essentially a single-stage amplifier in which a high degree of negative feedback from the anode to the grid is provided. A typical arrangement is shown in Fig. 2.15. Here a calculated fraction of the output

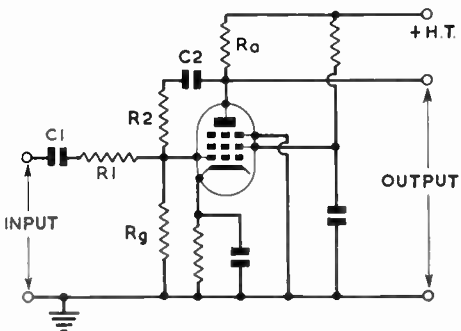


Fig. 2.15. Anode follower.

voltage is injected into the grid circuit by a potential divider consisting of the two resistors R1 and R2. The effect of this feedback is to reduce the stage gain to a value at which it is almost independent of any variations in the amplification actually produced by the valve. Thus, if the stage gain *without* feedback is A, the effective gain *with* feedback due to R1 and R2 is:

$$n = \frac{R_2}{R_1 + \left(\frac{R_1 + R_2}{A}\right)}$$

where the resistances are expressed in ohms. From this formula it is apparent that especially when A is large the gain depends mainly on the ratio R2/R1.

In a practical example, if the frequency characteristic of an amplifier *without* feedback were such that the gain fell by, say, 20 db at either end of the frequency range, this diminution in gain could be reduced to a much smaller amount, e.g. 3·5 db by the use of the anode follower principle. At the same time, however, the maximum gain would be lowered from perhaps 200 to something less than 10.

While it is the ratio of R2 to R1 that mainly determines the amount of feedback, they should both be reasonably high compared with R_a, the anode load impedance. Ordinarily R2 may be made equal to R_a.

The two capacitors C1 and C2 serve as d.c. blocking capacitors, but since they are associated with the feedback-determining resistors R1 and R2 it is important that the time constants R1C1 and R2C2 should be made equal so as to maintain a flat frequency response.

Provided that the grid leak R_g is not low compared with R2 it will not affect the performance. In practice it can be made approximately equal to R2.

The input impedance of an anode follower is determined by the value of the series grid impedance represented by R1. Because in general the input impedance is low, say 50,000 ohms, it is essential that the impedance of the source from which an anode follower is driven is relatively low. This would of course be so if a previous amplifying stage was used because this stage would have to supply an adequate output voltage of low distortion into a load of value R1 : a valve with an impedance of the order of 10 per cent of R1 would be required.

The output impedance is $1000\ (n + 1)/g_m$, where n is the stage gain *with* feedback and g_m is the mutual conductance in milliamperes per volt.

The output load need not be resistive as shown. If desired it could be in the form of a suitable choke or transformer.

Phase Splitters

Where a push-pull type of amplifier is required the simplest arrangement is usually one based on the use of a centre-tapped driver transformer. If it is preferred, however, to use a resistance-capacitance coupled driver stage, some ingenuity is needed to produce the two oppositely phased driving voltages from a single valve. Several different circuit arrangements have been devised for this purpose, but only four of them are described here. They are all known as *phase-splitter* circuits.

Paraphase Amplifier. In this arrangement two similar triodes are used, the grid of the second valve being driven from the anode of the first valve to produce the oppositely phased output. The basic circuit is shown in Fig. 2.16. Here V1 is a normal resistance-capacitance coupled voltage amplifier, its output appearing across R1 and R2 in series; this supplies one phase of the output. The part of this which appears across R2 drives the second valve V2 which also acts as an r.c. coupled voltage amplifier and supplies the opposite phase. The ratio of R1 to R2 is adjusted so that (R1 + R2)/R2 is equal to the voltage gain of V2, and R1 + R2 is made equal

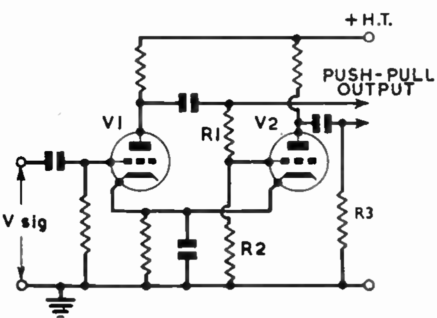


Fig. 2.16. Paraphase Amplifier.

to R3. Then the output voltages of both valves will be equal and opposite. The operating conditions for each valve are calculated in the same way as for voltage amplifiers but the values of R1 and R2 are critical and are chosen to suit the voltage gain. A common cathode resistor can be used and its value should be half that for each valve alone. The bypass capacitor may be omitted with some loss of balance between the two phases at the higher frequencies.

Split-load Phase Inverter. This type uses a single valve having half the load in the anode circuit and half in the cathode circuit as shown in **Fig. 2.17**. The two load resistors are R1 and R2. With respect to the load in the anode circuit the valve acts as a normal voltage amplifier, but with respect to the load in the cathode lead it behaves as a type of cathode follower. Since the current in both loads is the same, the voltages are equal if R1 = R2 and R3 = R4, but because of the large negative feedback through R2 the gain is reduced nearly to unity. Since the ratio of the feedback voltage to

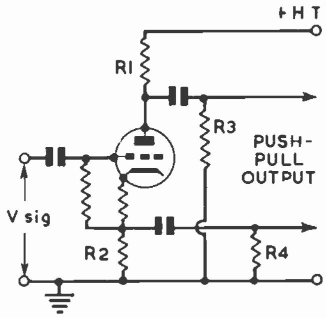


Fig. 2.17. Split-load phase inverter.

the output voltage is R2/(R1 + R2), and since R1 = R2 the feedback ratio is 0·5. In other words, an input signal V_{sig} of 1 volt will produce an output of 1 volt across both anode and cathode loads. Because of the feedback developed across R2, the output impedance in regard to R1 will be increased while the output impedance in regard to R2 will be decreased, and the source impedance from which the subsequent push-pull valves are to operate will therefore be unbalanced. Hence this type of phase splitter should not be used if the push-pull stage is driven into grid current at any time or is operated under any condition other than strictly class A. The operating conditions should be determined as if the valve were working as an RC coupled amplifier with R2 omitted and R1 twice the value actually used. The effective output voltage will be half the value so obtained.

The anode voltage is measured with respect to the cathode, not the earth connection.

Cathode-coupled Phase Splitter. In this arrangement two similar triodes are used, the input being applied to the grid of one of the triodes while the second valve receives its drive from a coupling resistor R_b in the common cathode load: see **Fig. 2.18**. When the grid potential of the first valve V1 swings positively the current through V1 increases and the voltage at the point *G* becomes more positive with respect to earth; i.e. the earth line becomes more negative with respect to the point *G* and the grid of V2 which is coupled to earth through its capacitor C1 consequently becomes more negative with respect to its cathode. The two grids thus operate in opposite phases. The outputs are taken from the two anodes and are balanced in amplitude and impedance, and the voltage

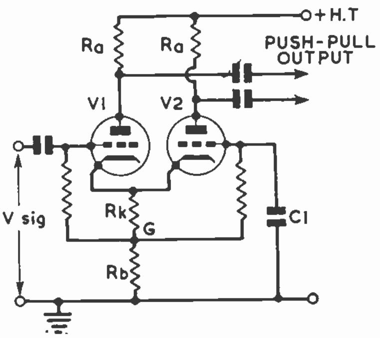


Fig. 2.18. Cathode-coupled phase splitter.

gain is about 25 per cent of that obtainable for each triode when used as a normal resistance-capacitance coupled amplifier. The operating conditions can be determined for each triode by assuming an anode load resistance of $R_a + 2R_b$ and a cathode-bias resistance of $2R_k$. The value of R_b should be approximately $\frac{1}{2}R_a$.

Anode-follower Paraphase Amplifier. If an anode follower is designed to have a high degree of feedback, the gain will be unity and the output will be equal in voltage but opposite in phase to the input. A suitable circuit is shown in **Fig. 2.19.** As in the anode follower circuit (Fig. 2.15), if R1 = R2 and C1 = C2 and if the voltage gain without feedback is very high, the gain is unity. The outputs have impedances differing to some extent because one output has a value equal to the source impedance of V_{sig} and the other is equal to $2/g_m$. The circuit constants are derived as for an anode follower.

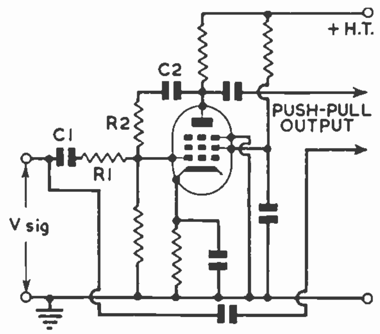


Fig. 2.19. Anode-follower paraphase amplifier.

Rectifiers

If a diode is connected in series with a resistor across an a.c. supply the diode will conduct current only in one direction and a pulsating d.c. voltage across the resistor will result from each positive half-cycle of the supply to the anode. If a large capacitor is connected across the resistor it will be charged during a d.c. pulse and maintain a voltage (falling somewhat due to discharge through the resistor) until the next pulse occurs a cycle later. Small diodes are used for signal rectification and grid-bias supplies, while larger diodes are used for h.t. supplies.

In a practical power supply the resistor is replaced by the output load, and to obtain the maximum efficiency the forward resistance of the diode must be small compared with the resistance of the load. A low forward resistance is also essential for good regulation of the output voltage, regulation in this context being the degree to which the output voltage is unaffected by the current load. One way of designing a rectifier valve with a low forward resistance is to fill the diode bulb with mercury vapour or xenon. The voltage-drop across a mercury-vapour rectifier is almost independent of current variations and is only about 15 volts. This is considerably lower than that of any high-vacuum rectifier.

The cathode of a gas-filled rectifier must be allowed time to attain its proper operating temperature before the anode voltage is applied. This is absolutely essential. Failure to observe this requirement will result in the disintegration of the cathode by ionic bombardment. In some types of equipment, automatically delayed h.t. switching is provided as a safeguard against carelessness of the operator.

The use of diodes as power rectifiers is described in Chapter 17 (*Power Supplies*).

Noise Diodes

If a diode is operated in a condition where the emission is temperature-limited, the anode current will contain a component of noise which is readily calculable. The frequency spectrum of the noise would be infinite unless it was limited at the higher frequencies by the shunt capacitance of the valve and its holder and by the electron transit-time between the cathode and the anode.

The noise output is controlled by adjustment of the anode current. This is achieved, while maintaining the temperature limited condition, by variation of cathode temperature. A directly heated pure tungsten filament is used and the filament current adjusted to give the required anode current. An indirectly-heated cathode or a coated or thoriated filament cannot be used since the emissive properties would be destroyed by operation in a temperature-limited condition.

Noise diodes are specially designed for the purpose of noise generation and if required for u.h.f. circuits they are frequently of coaxial construction.

Detectors

In a receiver the detector serves to rectify or demodulate the incoming carrier. If the carrier is modulated with speech or vision signals it must be demodulated in order to extract the modulation signals from it. If the carrier is unmodulated, as in telegraphy, it is necessary to mix with it another carrier of slightly different frequency, usually generated by a local oscillator (known as the *beat frequency oscillator*) and to rectify the mixture and thus produce an audible beat note or heterodyne in the loudspeaker. A variety of valve circuit arrangements are commonly used for this purpose and some of these are described below. Diodes are also used as signal rectifiers in receivers and in monitoring and measuring equipment.

Diode Detectors. The circuit of a conventional diode detector is shown in **Fig. 2.20**. An incoming unmodulated carrier V_{sig} produces a steady d.c. voltage across the load R_L the r.f. component of the rectified carrier being by-passed by the capacitor C_L. When the carrier is modulated the instantaneous voltage across R_L varies in accordance with the amplitude variation of the carrier and at the same frequency as the modulation. Thus, a carrier which is modulated by a 1000 c/s tone will produce a 1000 c/s voltage across R_L. The shunt capacitor C_L is sufficiently small to have no perceptible effect at audio frequencies. The coupling capacitor C_0 is included to provide d.c. isolation between R_L and any circuit connected to the detector output, such as an a.f. amplifier. To obtain a high degree of efficiency R_L should be high compared with the impedance of the diode, which in the case of a normal diode varies from about 200 ohms for low-impedance television types to about 10,000 ohms for those incorporated in multiple valves such as double-diode triodes.

A diode detector will cause distortion if the signal voltage applied to it is too low—say less than 10 volts peak—or if the impedance of the circuit following it is not high compared with the diode load. In Fig. 2.20, the combined impedance of the a.f. output control and the a.f. amplifier should be at least 10 times that of the d.c. load, R_L.

Leaky-grid Detectors. The operation of a leaky-grid detector is somewhat complex but can best be considered as analogous to that of a diode followed by a direct-coupled amplifier, i.e. one in which the grid of the amplifier is connected directly to the anode of the diode detector. The action can conveniently be illustrated by reference to **Fig. 2.21**. Here three equivalent arrangements of an ordinary diode detector circuit are shown at *A*, *B* and *C*, the first being similar to that shown in Fig. 2.20; in any of them the rectified output can be taken from the opposite ends of the load resistance R_L. If the diode is replaced by the grid and cathode of an ordinary triode, as in *D*, the arrangement becomes a *leaky-grid detector*, the grid now being in effect the diode anode. In this circuit, the load resistance R_L is known as the *grid leak*: if preferred it may be connected across the reservoir capacitor C_L instead of between grid and cathode without affecting

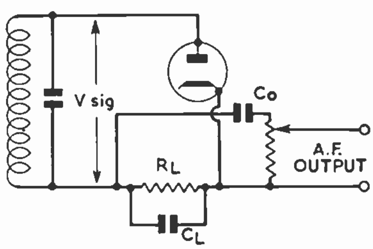


Fig. 2.20. A conventional diode detector.

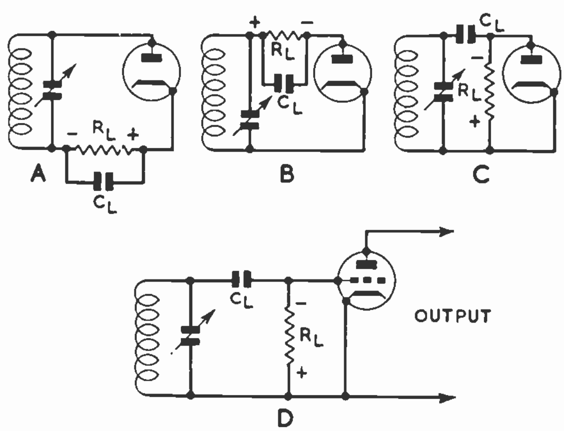


Fig. 2.21. Derivation of the leaky-grid detector.

the mode of operation. Any d.c. voltage developed across the grid leak by the rectification of a modulated or an unmodulated signal will thus constitute a negative bias for the grid and the anode current in the triode will fall. If the carrier is modulated, this bias will vary in accordance with the modulation signal and so likewise will the anode current vary. An excessively strong signal will tend to bias the valve beyond the cut-off point, and therefore a leaky-grid detector ceases to function satisfactorily when the input voltage is too great. To overcome this limitation, the *power grid detector* was introduced. This is substantially the same arrangement except that it employs a triode having a lower amplification factor and that it operates with a greater h.t. voltage; the valve thus has a longer grid base and consequently it can accept a larger input while still operating as a direct-coupled amplifier on the linear part of its characteristic. Hence there is less distortion but generally less sensitivity because of the lower gain.

The r.f. carrier is of course amplified by the normal triode action and will appear in the anode circuit. By applying some of this energy in the form of positive feedback (i.e. *reaction*) to the resonant circuit connected between the grid and the cathode, a leaky-grid detector can be made to oscillate. In this way it can be used for the reception of c.w. telegraphy. To prevent overloading and other harmful effects in subsequent a.f. amplifier circuits, the r.f. component in the detector anode circuit must be bypassed to earth by a suitable capacitance in combination with a resistor or an r.f. choke.

Tetrodes or pentodes can be substituted for triodes in leaky-grid detector circuits, although some difficulties may be encountered in avoiding *threshold howl* when the self-oscillating condition is required for c.w. reception.

Anode-bend Detectors. This type of detector depends for its action on the curvature in the anode-current/grid-voltage (I_a/V_g) characteristic. If a valve is biased nearly to cut-off and an a.c. input is applied to the grid, the negative half-cycle will drive the grid more negative and only a small change in anode current will occur, whereas the positive half-cycle will drive the grid more positive and cause a considerable increase in anode current. A valve having a short grid base and the sharpest possible bottom bend in its characteristic should be chosen. The valve may be coupled to the next stage by any method applicable to voltage amplifiers.

As in the case of the leaky-grid detector, the output contains an amplified r.f. voltage at the carrier frequency and the circuit can be made to oscillate by feeding back some of this r.f. energy from the anode to the grid circuit, and c.w. reception is thereby made possible without the use of a separate oscillator.

Infinite-impedance Detectors. If an anode-bend detector is rearranged in such a manner that the output load is in the cathode lead of the valve instead of the anode lead, the result is a type of cathode follower. The input impedance is extremely high owing to the 100 per cent negative feedback, and because of this the arrangement is referred to as an infinite-impedance detector. Fig. 2.22 shows the essential circuit. The modulated carrier is applied between grid and earth, the output load R_k being in the cathode lead. The load is bypassed by a capacitor C_k which should have a reactance which is low at the carrier frequency but high at the modulation frequency. The anode is also bypassed with a capacitor C_a large enough to have a low reactance at both frequencies. The rectified output is available across the resistance R_k and is supplied to the a.f. amplifier through a filter circuit comprising C_oR_o and C_gR_g. Such a filter is generally necessary because there may be quite an appreciable carrier voltage remaining across R_k despite the effect of C_k. The resistor R_o may be replaced advantageously by an r.f. choke if there is a diminished response at the higher audio frequencies.

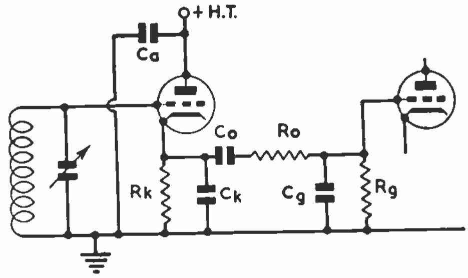


Fig. 2.22. Essential circuit of an infinite-impedance detector.

The resistor R_k should have a high value compared with the reciprocal of the mutual conductance of the valve. Since for most valves this is of the order of 100–1000 ohms, suitable values for R_k will be found in the range 10,000–100,000 ohms.

The gain of such a detector, like that of cathode follower circuits, will be less than unity and of the same order as that of a diode, but owing to the very high input impedance and the much smaller damping effect on the input circuit the actual gain will be appreciably greater than that of a diode. In the simplest case—a detector operating directly from an aerial—a step-up of signal voltage from the aerial into the grid of the same order as that achieved into an r.f. amplifying stage will be obtained.

In practice a detector circuit in which the cathode load R_k is 25,000 ohms and $R_o + R_g$ 250,000 ohms with an input of the order of 10 volts r.m.s. will produce a total harmonic distortion not exceeding 3 per cent at 100 per cent modulation. Inputs of less than about 5 volts to such a detector are not advised since the distortion increases at low input levels.

The value of C_k must not be made too small since the valve would then appear as a negative resistance to the tuned circuit, thereby causing instability or at least such an increase in the Q of the coil that the sidebands would be restricted owing to the increased selectivity. A suitable value for C_k, assuming that R_k is about 25,000 ohms, is 250–500 pF for a carrier frequency of 1 Mc/s.

Oscillators

An oscillator is usually a class C amplifier which obtains its grid excitation from its own output circuit and employs either an inductance-capacitance tuned circuit or a quartz crystal as the frequency-determining element.

If the value of the neutralizing capacitor C_n in Fig. 2.12 is increased to a high value compared with the grid-to-anode capacitance, this circuit will readily oscillate, and in fact it becomes the basic Hartley oscillator circuit. If the amount of feedback through C_n is too small (i.e. if there is insufficient excitation), a small increase in load may tend to throw the circuit out of oscillation, whereas if it is too large the grid current will be excessively high and there will be a wastage of power in the grid circuit.

An important consideration in oscillator design is frequency stability. The principal factors that cause changes in frequency are (*a*) temperature, (*b*) anode voltage, (*c*) loading, and (*d*) mechanical variations of the tuned-circuit components. Temperature variation will change the valve capacitances by causing expansion or contraction of the electrode assembly and will also alter the inductance and capacitance of the components in the tuned circuit; the resultant change in frequency is called *drift*. A change in anode voltage will change the power level in the circuit, resulting in a change in temperature and also a change in the effective valve capacitance. Many of these effects can be reduced by using a tuned circuit of high efficiency (high Q) and a low L/C ratio (high C), and the circuit should be only lightly loaded.

There are probably more circuit arrangements for oscillators than for any other valve application and these are more appropriately dealt with in the various chapters describing transmitters and receivers.

Frequency Multipliers

Frequency multipliers are designed in the same way as r.f amplifiers for class C operation but the anode circuit is tuned to a harmonic (i.e. a simple multiple) of the grid-circuit frequency. Because of this difference in resonant frequency neutralization is not normally required. The efficiency attainable is hardly ever likely to exceed 50 per cent and it diminishes as the order of harmonic multiplication increases, as shown in the typical figures given here:

Harmonic	2	3	4	5
Efficiency (%) ..	50	40	30	20

If the order of harmonic multiplication required is an even number and if the output obtainable from a single valve is barely adequate, a convenient arrangement is to use two valves connected in push-push. In this method two identical valves have their grids driven in opposite phases (as in the push-pull circuit) but their anodes are joined together and are treated as if they were a single anode: see Fig. 2.23. The anode circuit receives twice as many pulses in a given time as the pulses produced in *either* of the grids, and if it is tuned to twice the frequency of the grid circuit it will develop a voltage of twice the input frequency. A useful output can be obtained at higher even-order harmonics (4, 6, 8 . . .), although of diminishing amplitude. The push-push arrangement is inherently incapable of producing odd harmonics of the input frequency, but of course since its output waveform is not sinusoidal it will generate odd as well as even harmonics of its own nominal output frequency. Thus, if the input to a push-push multiplier were 3·5 Mc/s and the anode circuit were tuned to 7 Mc/s, the output would consist mostly of a 7 Mc/s component but it would also contain relatively weak harmonics of 14, 21, 28, 35 . . . Mc/s. There would be no 10·5 or 17·5 Mc/s components.

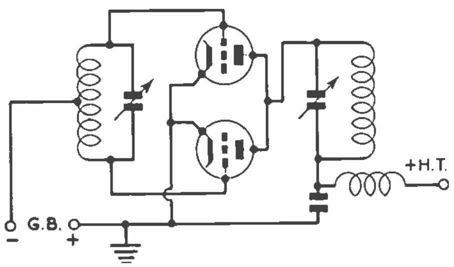


Fig. 2.23. The push-push frequency multiplier. The circuit produces only even-multiple harmonics: odd-multiple frequencies tend to be automatically cancelled out.

Frequency Changers

For a valve to act as a frequency changer it generally operates over a non-linear part of its characteristic. This implies that the process is similar to signal detection, and in fact any type of valve that will operate as a detector will operate as a frequency changer. It does not follow from this that any such valves will be efficient. The essential requirement is that an input signal of frequency f_1 is applied together with a heterodyning voltage of frequency f_2 and of such amplitude that it swings over a non-linear part of the characteristic and produces sum and difference frequencies $f_1 + f_2$ and $f_1 - f_2$ in the output circuit. The overall voltage gain of a frequency changer (i.e. the ratio of i.f. voltage output to r.f. signal-voltage input) is known as the ***conversion gain***, and the ratio of the beat-frequency component of frequency $f_1 - f_2$ or $f_1 + f_2$ in the output current to the input voltage of frequency f_1 is known as the ***conversion conductance*** (g_c) and is usually stated in microamperes per volt.

Since the primary purpose of frequency conversion in a communications receiver is to convert the high frequency of the incoming signal carrier to a more convenient lower frequency, it is only the *difference* frequency that is of any interest. To produce a difference frequency, the heterodyne oscillator frequency f_2 can, in principle, be chosen either higher or lower than the signal frequency f_1 and still yield the same result, but for various reasons connected with frequency stability and tuning range it is customary to make f_2 higher than f_1 in h.f. receivers and lower than f_1 in v.h.f. receivers.

Generally the signal voltage should be small compared with the heterodyning voltage: this is usually the case because the frequency changer is necessarily one of the early stages in the receiver circuit.

The voltage gain of a frequency changer is approximately equal to the conversion conductance multiplied by the dynamic resistance of the i.f. transformer primary. Proper allowance must be made in calculating the Q and the selectivity of the i.f. transformer for the anode-cathode impedance of the valve in shunt with the primary. If it is not possible to measure the conversion conductance, the gain of the stage may be calculated as if it were an i.f. amplifier operating at the same point, and as a rough guide the gain as a frequency changer will be about half the i.f. gain.

The basic types of frequency-changer circuit are described below.

Diode Frequency Changers. A diode may be used as a frequency changer, a typical circuit being shown in Fig. 2.24. The three circuits operating at the signal frequency (f_1), the local heterodyne oscillator frequency (f_2) and the output or intermediate frequency ($f_1 - f_2$) respectively are all connected in series with the diode. A resistive load R_k is also connected in series to prevent the diode from acting as a short-circuit to any of the tuned circuits and also to produce a suitable bias for determining the operating point on the characteristic. The impedance presented to any of the three

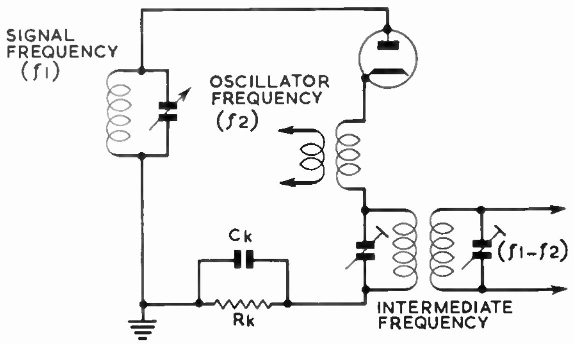


Fig. 2.24. Typical frequency changer circuit using a diode.

circuits is approximately equal to half the value of R_k. The value of C_k is chosen so that its reactance is low at the three frequencies involved.

Triode Frequency Changers. These are usually of the anode-bend type. The signal is applied to the grid while the oscillator voltage is injected in series with the cathode, and the i.f. transformer is connected in the anode circuit. The anode should be bypassed to earth with a capacitor of low reactance at the signal and oscillator frequencies. The capacity used for this purpose will form part of the i.f. tuning circuit.

A typical triode frequency changer is shown in **Fig. 2.25.** The resistor R_k and capacitor C_k now provide the automatic bias, the reactance of C_k being low at all three frequencies.

The amplitude of the heterodyne voltage used should be such that the grid voltage swings from beyond cut-off to nearly the point where grid current begins to flow. Triodes are used in this manner where a very high signal-to-noise ratio is essential and where the intermediate frequency is greatly different from the signal frequency. Care should be taken to match the i.f. transformer to the anode impedance.

The value of the conversion conductance g_c will be approximately one quarter of the value of the mutual conductance g_m when the valve is considered as an amplifier.

Pentode Frequency Changers. A pentode valve may be employed as a frequency changer in several different ways. Generally the signal is applied to the control grid and the

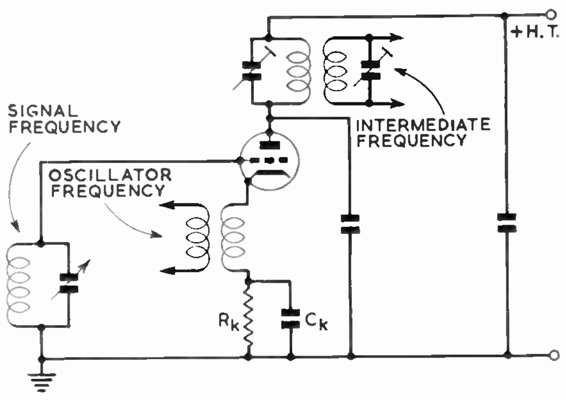


Fig. 2.25. A triode used as a frequency changer.

i.f. output is taken from the anode circuit, while in some arrangements the heterodyne voltage is injected in the control-grid or cathode circuit. In such cases the valve functions in the same way as an anode-bend detector. The same considerations apply as for the triode frequency changer, except that due to the much smaller capacitance between the control grid and the anode any interaction between the signal and i.f. circuits when they are not greatly different in frequency is less likely to occur.

Alternatively, the heterodyne voltage may be injected on the screen grid or the suppressor grid, but both of these methods suffer from the disadvantage of requiring a large heterodyne voltage to achieve a reasonably good conversion efficiency.

When suppressor-grid injection is employed, either of the two major bends in the characteristic may be used; if the cut-off bend is chosen, there is a possibility that the maximum safe screen dissipation will be exceeded unless steps are taken to limit the screen current. Because of the large heterodyne voltage that is necessary, it may happen that some of it appears in the control-grid circuit, for instance through the inter-electrode capacitances or as a result of poor layout, and this unwanted heterodyne injection will be amplified in the valve (the screen acting as an anode) and it can become comparable in effect to the heterodyne voltage intentionally applied to the suppressor. If it is of opposite phase, demodulation will take place and the conversion efficiency will be lowered. This effect is particularly troublesome at the higher frequencies.

Multi-grid Frequency Changers. These valves, which are specifically intended for use as frequency changers, are designed to have a reasonably high conversion conductance with as low a value of heterodyne voltage as practicable and a high anode impedance to prevent damping the i.f. transformer unduly. In the design of the receiver the signal frequency can be allowed to approach closely to the intermediate frequency without causing any interaction, and a.g.c. may be used if required. The only disadvantage of multi-grid frequency changers is their limited performance at the higher frequencies.

Some types are designed for use with a separate oscillator to provide the heterodyne voltage and others have a self-contained triode to generate this voltage. In certain types of heptode mixer, a part of the valve is arranged to oscillate as an electron-coupled oscillator, a section of the oscillator coil being included in the cathode circuit. The essential feature of the oscillator is that the voltage which it generates must as far as practicable remain constant over the required frequency range.

A *heptode* (sometimes referred to as a *pentagrid*) has seven electrodes, five of them being grids. Grid No. 1 is the oscillator grid, Nos. 2 and 4 are screen grids, No. 3 the control grid and No. 5 the suppressor grid. In some heptodes, however, the functions of No. 1 and No. 3 grids are interchanged, No. 1 being the control grid. This applies also to *hexode* mixers, which are basically similar to heptodes except that the suppressor is omitted and as a result the anode impedance is somewhat lower.

Some obsolete types of heptode employed grid No. 2 as an oscillator anode, grid No. 3 being the screen grid. This form of construction was frequently called a *pentagrid*, particularly in the USA, and when provided with an additional grid used as a suppressor it was referred to as an *octode*.

It should be borne in mind that multi-grid frequency changers are inherently noisy, having a high equivalent noise resistance. The equivalent noise resistance of a multi-grid mixer is given by—

$$R_{eq} = 20 \times \frac{I_a\,(I_k - I_a)}{I_k \times g_c^2} \text{ kilohms}$$

where I_a and I_k are the anode and cathode currents respectively (in milliamperes) and g_c is the conversion conductance (in milliamperes per volt).

A recently developed heptode is used as a locked oscillator f.m. detector in dual standard television receivers. It is also referred to as a " co-incidence detector " because if two sets of oscillations are applied to grids 1 and 3 anode current will only flow, assuming the amplitudes of the oscillations are sufficient, when the positive half cycles of the input waveforms coincide on the two grids. If both are negative or only one is positive and the other negative no anode current will flow. Coincidence of the positive half cycles will depend upon the frequency and phase relationship of the two oscillations. One oscillation is obtained from the TV inter-carrier i.f., usually 6 Mc/s, the other by using the valve grid 3 as a self-oscillator tuned to 6 Mc/s and frequency locked to the mean i.f. As the i.f. is frequency modulated the frequency applied to grid 1 will vary in relation to that of grid 3 with a result that pulses at the modulation frequency will appear in the anode circuit. The arrangement described has the advantages that no limiter is required and it is relatively easy to align compared with conventional ratio detectors but great care is essential in the design to ensure that the oscillator will lock to the i.f. under all conditions of signal level, mains voltage and temperature.

Multiple Valves

The trend is to make radio equipment as compact as possible and it is therefore convenient to take advantage of the special multiple valves which have more than one unit contained in a single envelope but only one multi-pin base. Many such valves are available and they generally contain two or more units which are associated with each other in a conventional circuit application.

Double-diodes are used for signal rectification and for providing a.g.c. voltages; they are often combined with a triode or pentode for a.f. or i.f. amplification. Treble-diode-triodes are used for frequency-modulation discriminators. Double-triodes are used for push-pull amplification, for multivibrators and many other purposes; the two sections may be similar or dissimilar in characteristics. Triodes are combined with r.f. pentodes for use as frequency changers and with a.f. pentodes for audio equipment or as generators for use with cathode-ray tubes. Double-tetrodes and double-pentodes may have certain electrodes commoned to avoid external circulating currents or may have a separate connection to each electrode.

Metal Valves

These valves, which at one time were very popular in the USA, are ordinary valve types assembled in a metal bulb, the lead-out wires being passed through eyelets mounted in the bulb. They are not now fitted in newly manufactured equipment.

Magic-eye Tuning Indicators

There are two basic types of magic-eye indicator. One type contains a cone-shaped piece of metal, known as the *target*, coated with fluorescent material. Electrons emitted from a cathode impinge on the target and produce a green glow over a part or the whole of its surface. Two or more small rods act as a grid and when the voltage applied to them is altered a shadow of controllable width appears on the target. A small triode may also be mounted within the bulb to amplify the voltage applied to the grid rods and thus improve the sensitivity. The grid of the triode is normally connected to

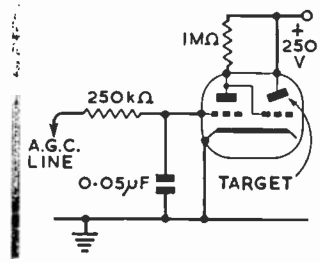


Fig. 2.26. A " magic eye " used as a tuning indicator.

the a.g.c. line in the receiver, the result being that the " eye " opens and closes in accordance with the signal voltage in the receiver.

The other and more recent type has a fluorescent coating on the inside surface of the bulb as in a cathode-ray tube, but the operation is essentially similar.

Certain types give a double indication, one for small voltage variations and one for large variations, and others give a differential indication suitable for the output of f.m. discriminators. **Fig. 2.26** shows a typical circuit arrangement for a magic-eye indicator in a superheterodyne receiver.

Its applications are by no means limited to tuning adjustments. In fact magic-eye tubes are often used as voltage indicators in a.f. and r.f. measuring equipment and also widely as indicators of signal level for tape recorders.

Voltage Stabilizers

Two methods are employed for voltage stabilization. The choice depends on the degree of output-voltage control required and the magnitude of the output current. Where all that is needed is a limited range of current and a control within only a few volts, a simple stabilizer of the cold-cathode type is adequate. This comprises an anode and a cathode (not heated) mounted in a bulb filled with an inert gas: when a sufficiently high voltage is applied to the anode the gas ionizes. The voltage at which this occurs is called the *striking voltage* or *starting voltage*. If the current through the tube is varied it is found that the voltage-drop across the tube is almost constant within a certain range of current variation. The voltage-drop in the centre of the current range is called the *maintaining voltage* and is significantly lower than the striking voltage. The maintaining voltage depends on the nature and pressure of the gas employed and to a limited extent on its temperature.

Some types have an additional electrode like a grid, known as a *trigger*. This electrode is situated close to the cathode, and if a sufficiently high positive voltage is applied to it the

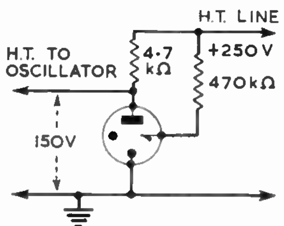


Fig. 2.27. A cold-cathode stabilizer used to provide a stable voltage of 150 volts from a supply line of 250 volts for the oscillator in a superheterodyne receiver. This eliminates frequency drift caused by h.t. voltage variations.

gas will ionize in the vicinity of these electrodes so that when the anode voltage rises it strikes at a lower value than it would without the trigger. For this reason the trigger is sometimes known as a *priming electrode*. More elaborate types are provided with a series of anodes and cold cathodes, each pair giving the same voltage-drop and thereby a series of discrete steps of stabilized voltage is made available.

A typical circuit for a trigger type of cold-cathode tube used to stabilize the voltage of the h.t. supply to the oscillator in a superheterodyne receiver is shown in Fig. 2.27.

Where a larger range of current than can be accommodated by a cold-cathode tube or where a closer regulation of voltage is necessary, a triode, tetrode or pentode valve may be used in series with the h.t. supply. A fraction of the output voltage is amplified and applied as grid bias to the stabilizer or regulator valve: this modifies the voltage-drop across the valve and so stabilizes the output voltage at a value depending on the circuit adjustments. Frequently normal amplifier-type valves are used for this purpose but specially designed valves having a lower voltage-drop permit a more economical design, details of practical circuits are given in Chapter 17 (*Power Supplies*).

Voltage-reference Tubes

A voltage-reference tube is a cold-cathode stabilizer designed to pass a very small current but maintain a voltage-drop within exceptionally close tolerances over long periods. Such tubes are used to provide a reference voltage for calibration purposes and they should not be used for heavy current in the manner of a normal stabilizer.

Electron Multipliers

The production of secondary emission by bombardment of a cold electrode is used in an electron multiplier valve to increase the density of the electron current, i.e. to increase the slope of the valve. Since a considerable velocity is essential in the bombarding stream of electrons and since each stage of electron multiplication therefore requires a certain minimum voltage, the number of stages that can be operated from a normal h.t. supply is limited

To prevent the secondary electrons from being attracted back towards an earlier electrode by reason of the relative electrode potentials it is necessary to focus them towards the anode. This is generally achieved by means of beam-forming plates or a construction similar to a cathode-ray tube gun.

Electron multiplier tubes are sometimes used in television receivers and in amateur television transmitters.

TRANSMITTING VALVES

In a transmitter most of the valves employed in the early stages are normal receiving-type valves, but in the final stages valves of higher power-handling capacity are necessary. Such valves fall into two classes:

(a) triodes and
(b) tetrodes or pentodes.

Those designed for operation on the higher frequencies are usually provided with special connections on the top or side of the bulb for the anode or grid, or sometimes both, and have special bases or in some types no base at all. This avoids the bunching together of the leads at one end of the valve and thereby lowers the inter-electrode capacitance and reduces the possibility of leakage or breakdown at high voltage.

Triodes. Triode transmitting valves are generally used with grounded-cathode or grounded-grid connections, the former being preferred at frequencies below 150 Mc/s and the latter at higher frequencies. They require more drive than tetrodes or pentodes but are more readily matched into the associated circuits and are often more convenient to modulate: they are seldom satisfactory as frequency multipliers.

Tetrodes and Pentodes. In amateur equipment, tetrodes and pentodes are used because of their high efficiency. Many of the popular tetrodes are in fact beam valves and have pentode-type characteristics. They require little or no neutralization because their grid-to-anode capacitances are sufficiently low, but if they are to be anode-modulated it is necessary to modulate the screen voltage as well as the anode voltage and the respective amplitudes must be correctly proportioned. Some pentodes have a separate connection to the suppressor grid and can be suppressor-grid modulated, though this system is not widely used on account of its low anode efficiency.

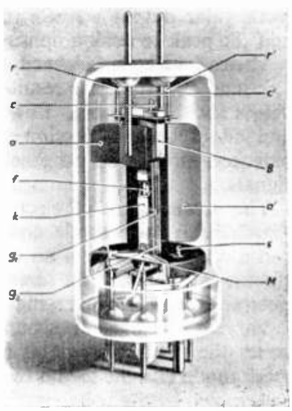
Double-tetrodes and Double-pentodes. A range of double valves having anode dissipations in the range 6–50 watts is available and is well suited to amateur requirements. Some have two separate units within one envelope Fig. 2.28, while others have a common flat cathode, one face being used for one section and the opposite face for the other section. The control grids are "half grids," one on each side of the cathode, and the screen is a common grid surrounding the two control grids and the cathode, while the anodes are flat plates, one on each side. This form of construction avoids

Fig. 2.28

Cut-away view of a v.h.f. double tetrode, the QQV06-40A.

r, r′—electrode support rods; c, c′—neutralizing capacitors; a, a′—anodes; B—beam plate; M—mica electrode supporting plate; k—cathode; g_1—control grid; g_2—screen grid; S—internal screen.

(Photo by courtesy of Mullard Ltd.)



circulating currents in a common cathode lead or in the screen lead when the two units are used in push-pull and enables a greater efficiency to be obtained at high frequencies. Some types also embody internal neutralization. Satisfactory operation up to about 600 Mc/s can be achieved with some of these double valves, both as class C amplifiers and as frequency multipliers using the third or fifth harmonic.

SPECIAL VALVES

Disc-seal Valves. The cathode is in the form of a thimble coated on the end, while the grid is composed of parallel wires tightly stretched across a hole in a metallic disc and the anode is a plain flat disc, the various parts being spaced by means of glass or ceramic seals. The outer rims of the discs extend beyond the seals for connection to the external circuit which can advantageously be of cylindrical co-axial form to suit the construction of the valve. Such valves are intended for use in grounded-grid circuits and will operate up to frequencies of several thousand megacycles with quite good efficiencies: they are available both as triodes and as tetrodes. The anode is generally cooled by an air blast through radiating fins or by conduction to the external co-axial circuit elements. Valves of this form are also known as *planar*, *lighthouse*, *rocket* or *pencil* valves.

A 2C39A disc seal transmitting valve.

(Photo by courtesy of S.T. & C. Ltd).

In the ceramic stacked triode the electrode discs are sealed directly to accurately machined ceramic washers. The valve can give 15db gain at 1000 Mc/s with a noise factor of 8·5db.

Quick Heating Valves. A range of single and double transmitting tetrodes has recently been introduced for use in mobile equipment. These valves are of the directly heated type but have a very low filament voltage of the order of 1·1 volts and cannot be operated directly from a 6 or 12 volt car battery. They have characteristics similar to the normal indirectly heated types. The filament is energized from additional centre tapped windings on the transformer of a transistor d.c. to d.c. converter. These windings have a rectangular output waveform, the r.m.s. value of which is half the peak to peak amplitude.

Valves for Single Sideband. The recent extension of single sideband operation has resulted in the production of special valves for this purpose. Low power types such as the 7360 and 6AR8 beam deflection valve are intended for use as balanced modulators to generate single or double sideband signals.

The arrangement of electrodes in the 7360 is shown in **Fig. 2.29**. A flat cathode, coated on one side, a control grid and a screen grid, form an electron gun which produces a ribbon-like beam. The screen grid and the two deflecting electrodes act as a converging electron lens to focus the beam.

Varying the bias or signal applied to the control grid varies the anode current as in a conventional valve. The total anode current to the two anodes at a given voltage is determined by the voltages applied to the control and screen

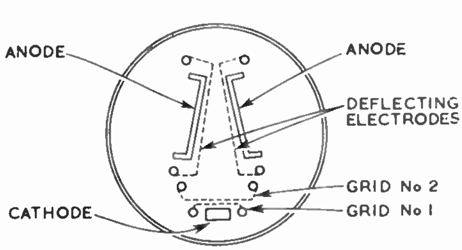


Fig. 2.29. Arrangement of electrodes in the 7360 beam deflection valve.

grids (accelerator). The division of the total anode current between the two anodes is determined by the relative voltage difference of the two deflector electrodes.

Linear amplifiers for s.s.b. operation are required to amplify the input signal without introducing distortion above an acceptable level. This means the valve must be capable of delivering high peak cathode currents without approaching emission saturation which would cause " flat-topping."

Two classes of operation are normally employed: class AB1 with tetrode or pentode valves and class B zero bias with high μ triodes. There is no doubt that class AB1 gives the better performance. Inter-modulation distortion of less than -25db for both third and fifth harmonics can be readily obtained. Valves such as the TT21, 6DQ5 and 6HF5 will give better than -30 and -40db for third and fifth harmonic distortion. Triodes operated in zero bias class B required considerably greater driving power than tetrodes or pentodes in class AB1 and make more severe demands on the driver stage. In both types, linearity can be improved by applying 6–10db negative feedback to both the linear amplifier and driver stages. The most satisfactory results are obtained by the use of large valves drawing relatively high zero signal inputs, which can be as high as 70–80 per cent of the total anode input power capability.

The design of linear amplifiers is considered in detail in Chapter 10—*Single Sideband Transmission*.

Nuvistors. This type of valve is intermediate between valves of normal construction and disc-seal or stacked ceramic. The electrodes are of cylindrical shape but smaller than those generally employed and are rigidly supported from one end by a tripod-like structure on a ceramic base. The supports are continuations of the pins which are sealed through the base—no use is made of mica or glass supports or spacers. This construction produces rugged and efficient valves having low leakage and low microphony. Because the structure is so small the capacitance between electrodes and the lead inductance is low, resulting in a good performance and low noise at v.h.f. A current type in wide use is the 6CW4 and 6DS4. These valves are no longer manufactured in Europe but triodes are available from the USA in both single and double ended types. An example of the latter is shown in **Fig. 2.30**.

Magnetrons. A magnetron is a diode with a cylindrical and concentric cathode and anode. In its simplest form the anode is split into two parts but more generally it comprises multiple cavities which resonate at the operating frequency. The whole assembly is placed in a magnetic field parallel to the axis of the electrodes. The magnet may be an integral part of the valve and it is then known as a *package mag-*

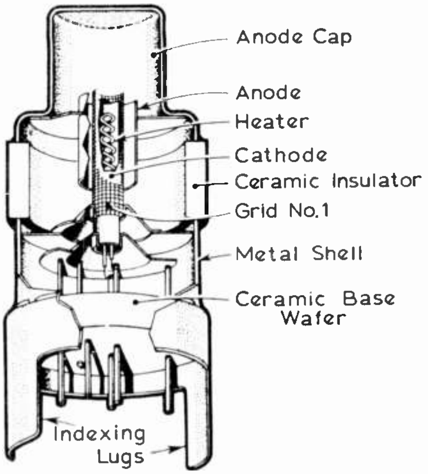


Fig. 2.30. Cut-away view of a double ended Nuvistor triode showing cyclindrical electrodes and tripod-like supports.

netron. Owing to the axial magnetic field the path of the electrons leaving the cathode is curved, and for a certain field intensity the electrons stop short of the anode and return to the cathode, travelling in circular orbits. The time taken for an electron to complete its orbital journey determines the frequency of oscillation. The energy associated with the moving electrons, is given up in the space around the cathode, whence it is transferred to the tuned cavities and picked up by a coupling loop. During operation the valve tends to become overheated due to back bombardment of the cathode unless the heater power is suitably reduced.

Klystrons. A *klystron* is a valve containing an electron gun similar to that used in a cathode-ray tube from which a narrow beam of electrons is projected along the axis of the tube and focused through small apertures across which one or more oscillatory circuits are connected; these circuits are in the form of hollow toroidal chambers known as *rhumbatrons*. The beam of electrons is velocity-modulated by the application of an r.f. field, for example by passing the beam through small apertures having an r.f. voltage between them. If this velocity-modulated beam is now passed through a field-free space, known as a *drift space*, the faster-moving electrons will begin to overtake the slower ones so that at some point along the beam alternate regions of high and low electron density will exist. The rhumbatrons are located at points of maximum density. If two rhumbatrons are coupled together by means of a feedback loop, or if a reflector electrode maintained at a slightly negative potential with respect to the cathode is mounted at the end of the tube the energy will be reflected back into a single rhumbatron. The electrons become bunched due to being velocity-modulated, and since the energy is retarded by reason of the distance of travel to the reflector and back again sustained oscillation will take place. Klystrons are employed as oscillators, the reflector type being preferred for low-power work, e.g. as a local oscillator in a superheterodyne receiver.

Travelling-wave Tubes. A *travelling-wave tube* consists of a wire helix supported in a long glass envelope which fits through two wave-guide stubs used as input and output elements. At one end of the tube is a conventional electron-gun assembly which directs an electron beam to the other end: see **Fig. 2.31.** The r.f. signal input is coupled into the helix and will run round the turns of the helix at roughly its normal velocity but its axial velocity may be only about a tenth of that value, depending on the pitch of the helix. If the electron beam is sent along the centre of the helix and focused by a magnet it will be charge-density modulated by the voltage pulse (i.e. amplitude-modulated) owing to the interaction between the magnetic field of the current in the helix and the electrons. This modulation grows in amplitude along the length of the helix roughly according to the square of the axial distance from the beginning of the helix. By suitable design the output wave-guide coupling will extract more energy than was put in by the input coupling and a considerable power gain can be achieved. For example, travelling-wave tube amplifiers with a bandwidth of 500 Mc/s centred on frequencies in the range 1700–8500 Mc/s are available. Outputs of 5–10 watts and power gains of 30–40db are possible.

To prevent self-oscillation, the central part of the helix is normally covered by a resistive material (usually on the outside of the tube) known as an attenuator.

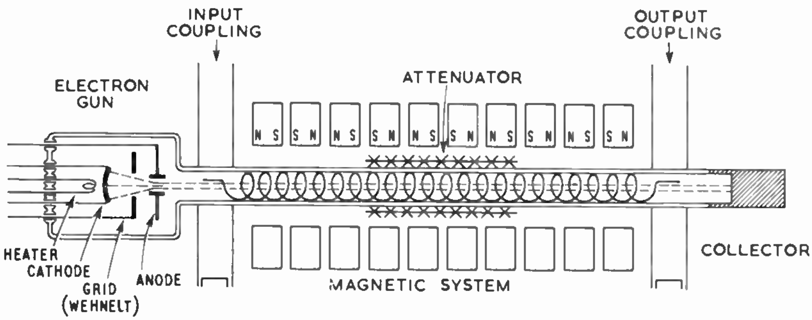


Fig. 2.31. Travelling-wave tube.

Beam Parametric Amplifiers (Adler Tubes). These tubes utilize a normal electron gun generating a focused electron beam which passes through electrodes to which the input signal is connected (generally coaxially), then through a quadrupole of four curved electrodes, which may be cross connected or coupled by inductances to which the pump signal is applied. The quadrupole lies in a transverse magnetic field provided by a permanent or electromagnet. The next electrodes are the output electrodes followed by one or more collectors or anodes. The pump frequency, of power only a few milliwatts, is usually at approximately twice the signal frequency. They have so far been made for frequencies between 200 and 4000 Mc/s and achieve a noise factor of about 2db, a gain of over 20db and a wide bandwidth. Unlike most parametric devices they are inherently stable because the input, pump and output connections are quite separate.

CATHODE RAY TUBES

In a cathode-ray tube, the electrons emitted from a hot cathode are first accelerated to give them a high velocity and are then formed into a beam and finally allowed to strike a fluorescent screen, thereby causing a spot of light to appear at the point where the beam strikes. The part of the tube which generates the beam of electrons is known as the *gun*. A beam of moving electrons can be deflected laterally by electric or magnetic fields, and since its weight and inertia are negligibly small it can be made to follow almost instantaneously the variations of periodically changing fields even up to very high radio frequencies, producing a visible image of these variations.

In a simple type of tube the gun consists of a heater, cathode, grid and two anodes: see Fig. 2.32. The intensity of the electron beam and therefore the brightness of the trace on the screen is regulated by the grid potential as in a valve. Anode No. 1 has a positive potential with respect to the cathode and accelerates the electrons which have passed through the grid: it is provided with small apertures through which the electron stream passes. The stream, on emerging from the apertures, is formed of electrons travelling in parallel straight-line paths. The electrostatic fields set up by the potentials on anodes Nos. 1 and 2 form an *electron lens* which causes the electron paths to converge or focus to a

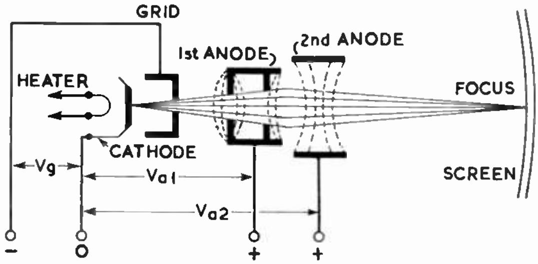


Fig. 2.32. The electron gun used in one type of cathode-ray tube for focusing the electron beam into a narrow pencil of rays. The electric fields of force act in the manner of lenses for deflecting the paths of the electrons.

point at the fluorescent screen. The focal length of the lens depends on the spacing of these anodes and their respective potentials, and this type of focusing is known as *electrostatic focusing*. The beam can, however, also be focused by a current passing through a coil mounted axially on the neck of the tube or by a permanent magnet. These methods are known as *magnetic focusing*. An electron gun, in its simplest form, can be a triode in which case it must be magnetically focused, but more complicated guns in the form of tetrodes, pentodes and hexodes are commonly used. In general, electrostatic focusing is much more convenient for oscilloscope tubes, but either method is applicable to television or radar tubes.

The focused beam can be deflected electrostatically or electromagnetically. For *electrostatic deflection* two pairs of plates are mounted at right angles to each other, as in Fig. 2.33. The beam passes between the plates and is bent upwards and downwards by potentials between the horizontal

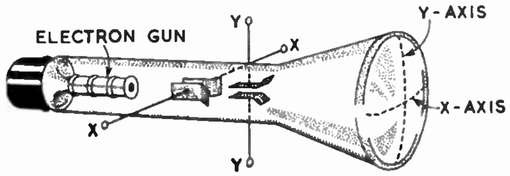


Fig. 2.33. Electrostatic deflection of the electron beam in a cathode-ray tube.

pair of plates (the *Y*-plates) and from side to side by the vertical pair (the *X*-plates). For *magnetic deflection* two pairs of coils are mounted radially on the neck of the tube, as in Fig. 2.34, one pair causing vertical deflection (*Y*-axis) and the other pair horizontal deflection (*X*-axis).

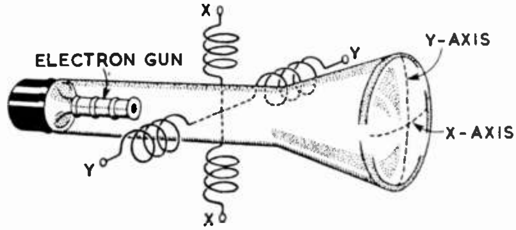


Fig. 2.34. Electromagnetic deflection of the electron beam in a cathode-ray tube. The coils are sometimes provided with magnetic cores.

In the electrostatic-deflection type of tube, the beam can be split into two by a separator plate thus enabling the screen to show two independent traces simultaneously. Such a tube is known as a *double beam* tube. Two or more beams can also be produced by tubes having multiple gun assemblies mounted within a single neck.

The brightness of the trace on the screen depends on the final anode voltage and on the beam current and the speed at which the spot moves. Care must be taken to avoid too bright a spot since an intense concentration of high-speed electrons will burn the screen, and too great a beam current will shorten the life of the cathode.

Many tubes are provided with a metallized screen which reduces the risk of burning and also considerably increases the light output. After the fluorescent coating has been deposited on the glass face during manufacture an extremely thin film of aluminium is formed by vaporization over the back of the screen. The layer is so thin that the flow of electrons to the screen is not seriously reduced, but it is sufficiently smooth and mirror-like to reflect the light produced, much of which would otherwise be lost.

Apart from giving a much brighter image, metallized screens enable tubes to be operated at much higher voltages and also reduce the burning caused by bombardment by heavy negative ions generated at the cathode. The latter effect normally only becomes a serious problem in tubes which are operated at high voltages such as in television sets and radar equipment.

CHAPTER 3

SEMICONDUCTORS

SEMICONDUCTOR devices perform similar functions to those performed by thermionic valves in radio equipment and fall broadly into two categories: *crystal diodes* and *transistors*. Unlike their thermionic counterparts semiconductors require no heater voltage and are current operated, as opposed to voltage operated, devices.

Although the crystal diode in its modern form was accepted easily by radio amateurs as a variant of a device with which they are already familiar, the transistor presents a more difficult problem, and demands an understanding of its characteristics before it can be successfully used. Most texts on the subject are directed towards the specialist in the field and it is the purpose of this chapter therefore to introduce the reader to the fundamentals in particular of semiconductor circuitry without resorting to advanced physics or complex electron flow theory and with particular reference to the requirements of the Amateur Radio enthusiast.

It was not until 1948 that the first practical transistors were made but development has been so intensive since then that a very wide range of types is now available including those suitable for v.h.f. and u.h.f. The practical advantages include low power consumption, small physical size and ruggedness.

A possible disadvantage lies in the fact that, unlike the thermionic valve, many of the transistor characteristics are temperature-dependent. In most applications, however, the effect can be very largely offset by careful attention to the circuit design and by skilful development of the bias circuits. The requirements are so different from those of thermionic valve circuits that it is advisable to consider the basic electron-flow processes before attempting a full understanding of transistor circuit technique.

THEORY OF SEMICONDUCTORS

The ability of a substance to conduct electricity depends on the readiness with which electrons can become detached from their parent atoms. In the case of pure germanium, which is a crystalline substance, there are very few free electrons at room temperatures, and the material is classed as a *semiconductor*. Each atom has four valence electrons, which are the electrons on which conductive properties depend, and atoms " knit " with their neighbours in a fashion typical of crystalline substances as indicated in Fig. 3.1. The few electrons that may be available as current carriers are those which have been " shaken " free by thermal agitation. The effect of thermal agitation is to cause electrons to break away from their bonds and drift into the orbits of neighbouring atoms. Normally the process is quite random, but when a voltage is applied between two points on the substance the movement of electrons becomes a general drift towards the positive pole of the applied voltage.

Each time an electron breaks away from its atom it leaves a vacancy, or " hole," which remains until it is filled by another electron from a neighbouring atom. Thus the " hole " effectively moves in a direction opposite to that of the electron, and a general flow of electrons in one direction can be looked upon as a flow of " holes " in the reverse direction. Although this may appear to introduce an unnecessary or purely hypothetical concept it undoubtedly simplifies the problem of understanding the way in which crystal diodes and transistors operate.

The conductivity of a germanium crystal may be greatly increased by the addition of a small amount of impurity into the crystal, and two different types of impurity are purposely added in this way—those with five valency electrons and those with only three. Such an impurity is arsenic which is a penta-valent substance. When added to germanium it interrupts the crystal formation because each arsenic atom has one excess electron which prevents it from fitting into the pattern. The surplus electrons are very mobile and are readily available as current carriers when a voltage is applied. Germanium which has been treated in this way is known as *n* type germanium because the current carriers are electrons and therefore negatively charged. If indium is added as an impurity, however, a similar effect occurs but this time each impurity atom introduces a " hole " because indium is a tri-valent element. As explained above,

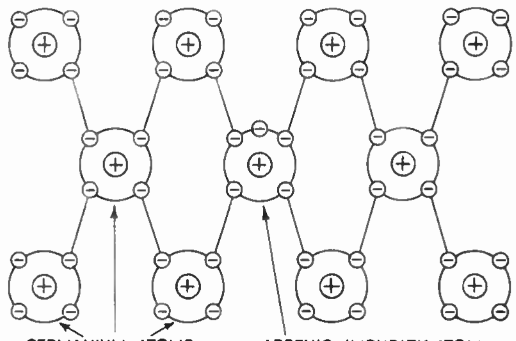


Fig. 3.1. In pure germanium, which is a crystalline substance, the four electrons in the outer orbit of each atom form tight bonds with those in neighbouring atoms and leave few electrons available for carrying electric current. The addition of a small amount of impurity such as arsenic, which has five electrons in the outer orbit, introduces free electrons to the structure and increases the electrical conductivity. Germanium " doped " in this way is known as *n* type.

a " hole " may be mobile and thus increase the conductivity of the crystal. Germanium doped with indium is known as *p* type because conduction is due to positively charged carriers (i.e. " holes ").

The P-N Junction

Fig. 3.2 shows a germanium crystal, half of which has been doped with arsenic and the other half with indium. One half is therefore *p* type and the other *n* type. If such a crystal is prepared, then at the junction between the two regions, free

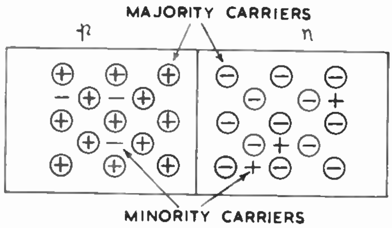


Fig. 3.2. If an impurity with only three electrons in the outer orbit is added to a germanium crystal, electron deficiencies, or " holes " are introduced and the germanium is known as *p* type. A combination of *p* region and *n* region in a single crystal, as shown in this diagram, is the basis of a junction diode.

electrons will move across from the *n* region and begin to fill up the " holes " in the *p* region. Because the recombination process entails a flow of charge carriers through the junction, a contact potential is developed across the junction of such a polarity as to oppose the flow of carriers and a state of equilibrium is quickly reached. There remains a thin region in the vicinity of the junction which has been cleared of mobile carriers and is known as the *depletion layer*.

In Fig. 3.3 the *p-n* junction is shown with an external voltage applied such that the *p* region is made positive and

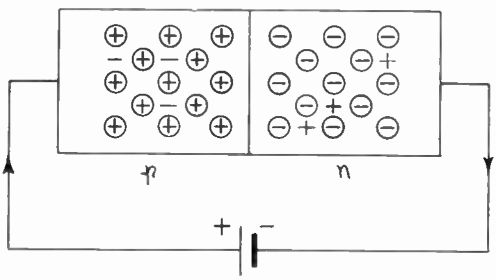


Fig. 3.3. A *p-n* junction with an external voltage applied as above is said to be forward biased. The relationship between voltage and current is not linear but follows an exponential law.

the *n* region negative. Under these conditions the device behaves as a good conductor because the holes in the *p* region and the electrons in the *n* region are of the correct polarity to support current flow. The magnitude of the current flowing will depend on (among other things) the applied potential, but the relationship between current and voltage is affected by the presence of the depletion layer, and the resultant characteristic takes the form shown in Fig. 3.4. This is the forward characteristic of the *p-n* junction.

If the polarity of the external voltage is now reversed, the available current carriers in the two halves are then of the wrong polarity for conduction and the device behaves rather

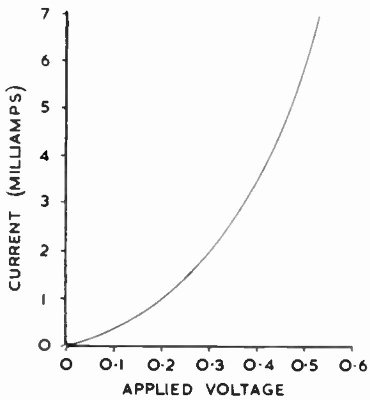


Fig. 3.4. Forward characteristics of a *p-n* junction diode.

like a very high resistance. The relatively small current that does flow is due to the presence of thermally-generated holes and free electrons. Thus, at room temperature there will always be a few *n* type carriers in the *p* region and, likewise, some *p* type carriers in the *n* region. Such carriers are present in only small proportions and are termed *minority carriers*. Because they are so few in number, the reverse current reaches saturation at a very small applied voltage. Fig. 3.5 shows the complete forward and reverse characteristics of a typical *p-n* junction; its suitability as a rectifier can be clearly seen and the device is known as a *junction diode*.

It is important to note that the reverse conduction is entirely dependent on thermally-generated carriers, and it therefore varies markedly with the temperature of the junction. In fact, the relationship in the case of germanium is such that the current approximately doubles for each 8°C. rise in the temperature of the junction. This feature is perhaps the most significant difference between thermionic and crystal diodes.

Many crystal diodes in current use are of the *point-contact* type. These differ from the junction diode in that they

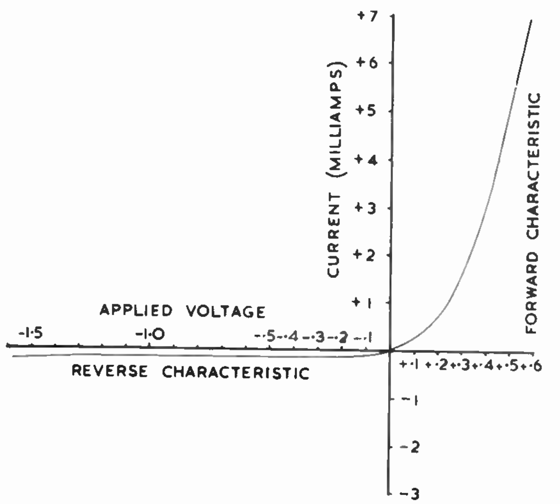


Fig. 3.5. Complete characteristics of a *p-n* junction diode showing the way in which reverse current saturates at a very low value.

employ a single piece of *n* type germanium with a "cat's whisker" in contact with its surface. Although the underlying principle of operation is based on that of the junction diode, the exact mechanism of the process is not yet completely understood; it is believed, however, that the effect of the "cat's whisker" is to create a small region of *p* type substance in the vicinity of the contact. The particular characteristics of the point-contact diode are high forward resistance and good high frequency performance.

The Transistor

In Fig. 3.6 an elaboration of the basic *p-n* junction is shown in which the *n* region is made very thin and a second region of *p* type germanium is added to form a *p-n-p* sandwich. This device can be considered as two *p-n* junctions mounted back to back: each junction is capable of being biased separately and of reproducing the characteristics described previously. If, however, both junctions are biased simultaneously rather special conditions apply. The left-hand junction is biased in the forward direction and therefore presents a low resistance to the flow of current, while the right-hand junction is biased in the reverse, i.e. high resistance, direction. Because the *n*

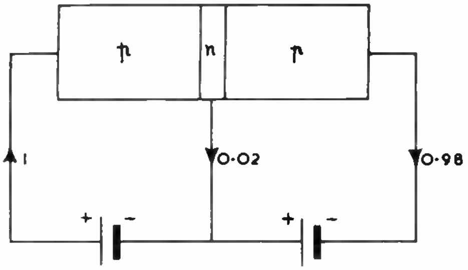


Fig. 3.6. A *p-n-p* sandwich forms two *p-n* junctions mounted back to back and is the basis of the transistor. The left-hand junction is forward biased but a large proportion of its input current becomes transferred to the right-hand junction, which is reverse biased. Power gain arises because an input current into a low resistance appears in a circuit of much higher resistance.

region is very thin, many of the current carriers flowing into the *n* zone from the forward biased circuit become captured by the *p* region in the reverse biased circuit. For example, if a current of 1 mA flows into the left-hand *p* region, instead of the whole of this current appearing in the connection to the *n* zone, about 98 per cent of it may flow into the right-hand *p* region and thus appear in the normally very high resistance circuit. Only about 2 per cent of the "input" current will then be extracted from the *n* region. This is the basic transistor action, power gain being achieved by virtue of the fact that an input current fed into a low resistance circuit reappears in a circuit of much higher resistance. If the "sandwich" is made *n-p-n* instead of *p-n-p* the same process will occur provided that the polarities of the bias circuits are reversed. Many transistors available at the present time are of the *p-n-p* type and of a construction similar to that shown in Fig. 3.7. Germanium and silicon are the most commonly used substances; silicon has the advantage of being less temperature dependent.

In all transistors the outer electrode of the forward biased junction is termed the *emitter*, the centre electrode the *base*, and the reverse biased electrode the *collector*.

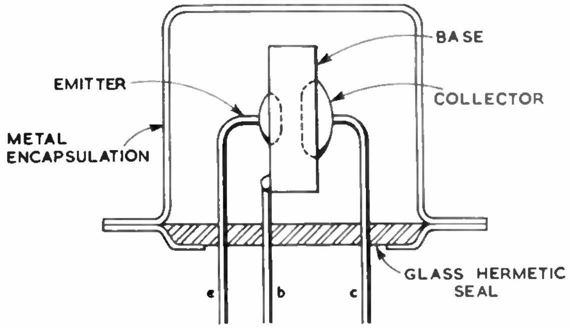


Fig. 3.7. Construction of a typical *p-n-p* audio frequency or r.f. transistor. This is known as the alloyed junction type in which the junctions are formed by alloying *p* type impurities into opposite faces of a wafer of *n* type germanium. The centre region is known as the base electrode, the forward biased electrode is termed the emitter, and the reverse biased electrode the collector.

Early transistors were not junction types but were known as *point-contact transistors*. These employed a method of construction similar to that of the point-contact diode except that two "cat's whiskers" were involved. They tended to be unreliable and inconsistent and have been almost completely superseded by the junction type.

BASIC CIRCUIT CONFIGURATIONS

So far in this description the transistor has been considered as operating in such a way that the base connection is common to both the emitter and the collector circuits. This is known as the *common base* configuration and is characterized by a current "gain" of a little less than unity, and by a low input resistance and a very high output resistance. A recognized symbol for the current gain in this arrangement is α (alpha). From Fig. 3.8, which indicates the current distribution when one unit of current is fed into the emitter or input electrode, it can be seen that the ratio of collector current to base current is $\alpha/(1-\alpha)$: this ratio is given the symbol β (beta). In a typical transistor, β is about 50. Thus, if a common emitter arrangement is adopted, the device becomes capable of providing a worthwhile current gain. The available power gain is very high and, for this reason, the *common emitter* arrangement is by far the most widely used.

Nevertheless when using a transistor at frequencies near

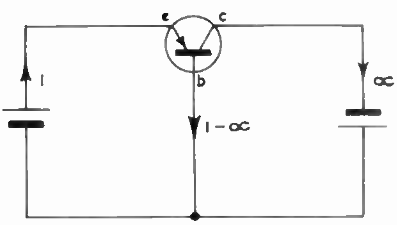


Fig. 3.8. In this diagram the *p-n-p* transistor is represented by its circuit symbol. The symbol α is used to express the ratio of collector current to emitter current and is typically 0·98. Thus the ratio of collector current to base current is $\frac{\alpha}{1-\alpha}$ and is known as β.

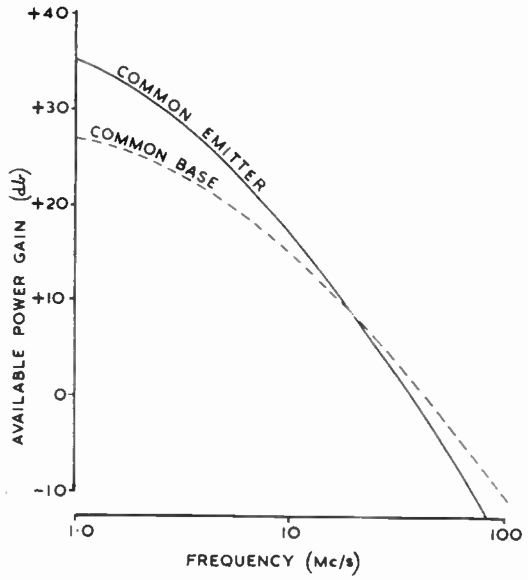


Fig. 3.9. Because of the finite transit time of the current carriers, the current gain of a transistor falls off as frequency is increased. The frequency at which the common base current gain has fallen by 3 db is known as the alpha cut-off frequency and is given the symbol $f\alpha$. The available power gain falls off at a rather faster rate, as shown in this graph which is plotted for a transistor with an $f\alpha$ of about 10 Mc/s.

the practical limit for its type there may be an advantage in using the common-base circuit. This arises because, although the common-emitter arrangement is capable of greater low frequency power gain, the gain falls off at a faster rate with increasing frequency than it does in the common-base arrangement. Typical gain frequency characteristics of a transistor in each of these two configurations are shown in **Fig. 3.9.**

Configuration	Input Resistance	Output Resistance	Current Gain	Power Gain
COMMON BASE (e, c, in, out, b)	40 ohms	1 Mohms	0·98	1,000 (30 db)
COMMON EMITTER (b, c, e, in, out)	2 K ohms	30 K ohms	50	10,000 (40 db)
COMMON COLLECTOR (b, e, c, in, out)	100 Kohms	1 K ohms	50	40 (16 db)

Fig. 3.10. Typical small-signal low frequency characteristics of a junction transistor in each of the three basic configurations are shown in this table.

The third possible configuration is the *common collector* arrangement but this has only specialized application. It is analogous to the cathode follower valve circuit and is characterized by a high input resistance and a fairly low output resistance. The current gain is $\frac{1}{1-\alpha}$, which is equal to $\beta + 1$, and therefore approximately the same as that of the common-emitter circuit. In spite of the high current gain, the power gain is not very great because of the adverse impedance transformation. This arrangement is however well suited for use as an input stage for a crystal microphone

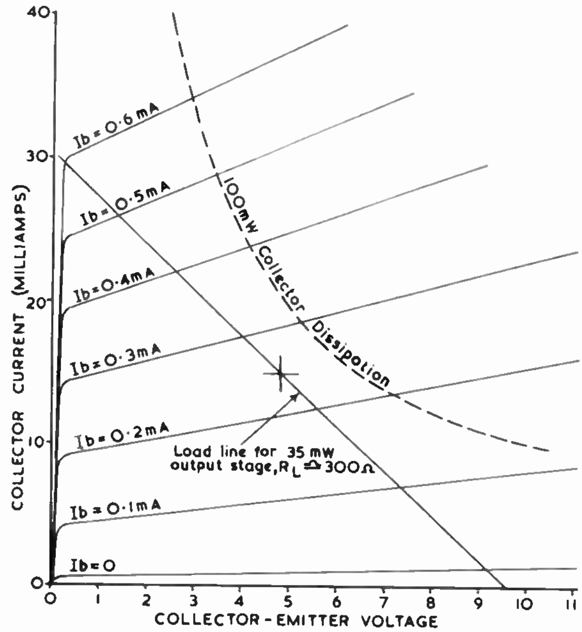


Fig. 3.11. A typical family of collector characteristics for a transistor in the common emitter configuration. Their general shape is very similar to the anode characteristics of a pentode valve and they can be used in a similar way. To demonstrate this, a load line has been superimposed to represent the working conditions of the class A output stage of Fig. 3.15. Note the low "knee" voltage which enables very high efficiencies to be obtained in a transistor power stage.

or pick-up owing to its high input impedance and low output impedance.

Typical characteristics of a transistor operating in each of the three configurations are listed in **Fig. 3.10.**

IMPORTANT TRANSISTOR CHARACTERISTICS

In spite of the fact that the basic operation of the transistor is so different from that of the thermionic valve, there are distinct resemblances between some of their characteristics.

Fig. 3.11 shows a family of collector characteristics for a typical transistor in the common emitter configuration. The general appearance is very similar to that of the anode characteristics of a pentode valve, but each curve in the family is plotted for a particular value of input *current* (I_b) to the base electrode in contrast to the input *voltage* applied to the grid of the thermionic valve. This clearly illustrates

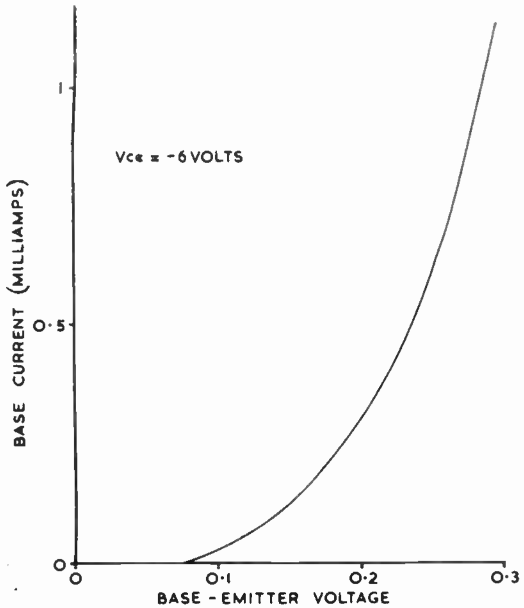


Fig. 3.12. The input characteristics of a transistor resemble the forward characteristics of a junction diode. The slope of the curve gives the small-signal low frequency input resistance of the transistor and varies with bias level.

the importance of regarding the transistor as a current operated device. The lowest curve of the family represents the collector current that will flow even without any input current and is termed the *collector leakage current*. This is the basic temperature dependent parameter and causes the whole family of collector characteristics to move upward with increasing temperature.

When using the collector characteristics for determining the working point and the load impedance for a particular application the procedure is similar to that for a thermionic valve: see Chapter 2 (*Valves*). As an example, a load line has been drawn across the characteristics to represent a class A power stage operating from a 6 volt supply and delivering a power output of about 35 mW into a transformer coupled 300 ohm load. (This load line represents the conditions in a stage operating from a 6 volt supply but with a voltage drop of 1·2 volts across an emitter stabilising resistor.) Because of the very small " knee " voltage it is possible to achieve very high output circuit efficiency in a transistor stage.

The small signal output resistance of the transistor at any point is given by the slope of the characteristic at that point, while the current gain (β) can be readily deduced by relating the base and collector currents at the same point.

Fig. 3.12 represents the input characteristics of a typical common emitter connected transistor and is obviously similar to the normal forward characteristic of a junction diode. This curve may be used for determining the input bias conditions when the required collector current has been established. The small signal input resistance is indicated by the slope of the curve at the working point and, as can be seen, varies quite considerably with the base current. Only one curve is usually shown because the input characteristics are fairly independent of the collector voltage.

The frequency limitation of a transistor is normally expressed by quoting the frequency at which the current gain has fallen to 70 per cent of its low frequency value in the common base arrangement. This is usually known as the frequency of *alpha cut-off* and is denoted by the symbol $f\alpha$. Because of various capacitive effects, the power gain begins to decrease at a much lower frequency and it is not ordinarily practicable to operate a transistor at frequencies greater than $f\alpha/2$, except perhaps as an oscillator.

Alternative terms defining the limiting frequency of operation include f_{hfb}; f_1 (the frequency at which $\beta = 1$) and f_T (the gain bandwidth product β *f* where β* is measured at f* above the β cut off frequency). All these terms are approximately the same as $f\alpha$ but, if the β cut off frequency $f\beta$ or f_{hfe} is encountered, the value is approximately $\frac{f\alpha}{\beta}$.

A very important rating for a transistor is the *maximum collector dissipation*. Most manufacturers quote this for a specified ambient temperature and give a figure which indicates the derating factor to be applied at higher temperatures.

STABILIZATION OF THE WORKING POINT

To give some idea of the importance of good stabilization of the working point, **Fig. 3.13** shows a curve of collector leakage current plotted against the junction temperature for a typical low power transistor in the common emitter circuit. At room temperature the leakage current is quite insignificant, but at 40° C. (104° F.) it rises to almost 1 mA, which may be considerably greater than the desired working current in a small signal stage. A good bias circuit will reduce the maximum leakage current to an acceptable level over the required working temperature range and also reduce the

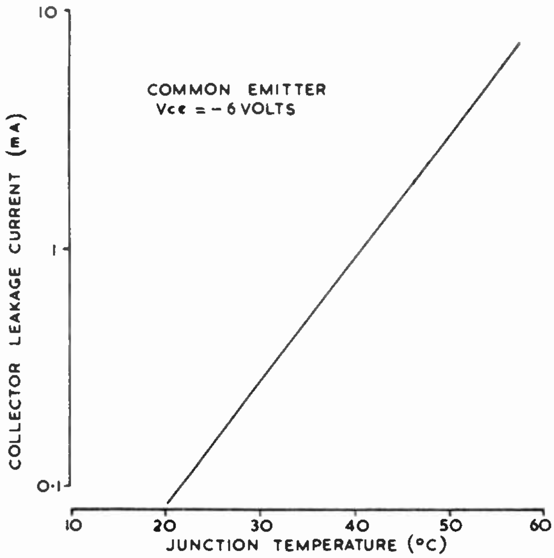
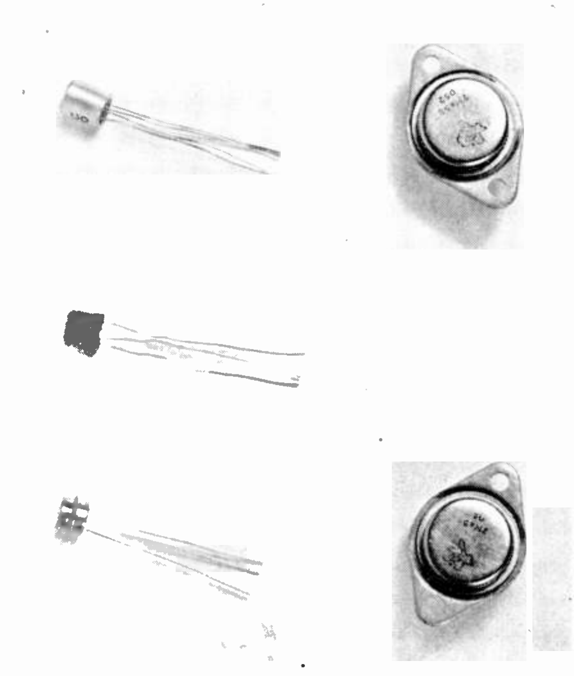


Fig. 3.13. Variation of collector leakage current with junction temperature for a typical l.f. transistor. Leakage current increases so rapidly with temperature that it is possible for a condition known as " thermal runaway " to occur. This arises when an initial increase of temperature leads to sufficient rise of leakage current to cause a further significant rise of temperature. The process then becomes rapidly cumulative and the transistor is destroyed. A carefully designed bias circuit and a knowledge of the transistor's limitations are the best safeguards against this.



Typical modern transistors. Those on the right are power types.

effects of the variation of characteristics between transistor samples.

In Fig. 3.14 a transistor is shown with a skeleton bias circuit of a form which is applicable to most types of transistor stages used in communications equipment and which uses a single battery to supply the input bias and collector voltage. The base of the transistor is fed from a potentiometer formed by R1 and R2, while a third resistor R3 is inserted in the emitter lead. The amount of bias current I_b fed into the base electrode, and therefore the amount of collector current I_c that will flow, is decided by the difference between the voltages V_b and V_e. If, for any reason, the collector current tries to increase then the voltage drop across R3 (i.e. V_e) will also increase. This will leave less voltage available for providing bias, and therefore the bias current will decrease and try to oppose the change in the collector current. The effectiveness depends on (*a*) the bleed current in the base potentiometer being sufficient to ensure that V_b remains constant, and (*b*) the value of R3 being large enough to produce a significant voltage change across it when the collector current changes. A good rule of thumb for small signal stages is to arrange that the bleed current is about one fifth of the desired collector current and to drop about one-fifth of the total battery voltage across the emitter resistor R3. This is best demonstrated by a worked example.

The Practical Design of a Bias System

Suppose that the transistor has the characteristics shown in Figs. 3.11 and 3.12 and that it is to be used in the 35 mW output stage represented by the load line in Fig. 3.11. Suppose also that the battery voltage is 6v and the required collector current is 15 mA. Following the recommendation above, the bleed current in the base potentiometer is therefore $\frac{1}{5} \times 15 = 3$ mA, and the voltage drop across R3 is $\frac{1}{5} \times 6 = 1{\cdot}2$ volts. Ignoring the base current taken by the transistor (which is much smaller than the bleed current) the required total resistance of the potentiometer is $\frac{6\text{V}}{3\text{ mA}} \times 10^3 = 2{,}000$ ohms. As already explained, the emitter current is approximately equal to the collector current (see Fig. 3.6) and it is therefore convenient to find the value of R3 by dividing the voltage drop across it by the collector current, thus—$\frac{1{\cdot}2\text{V}}{15\text{ mA}} \times 10^3 = 80$ ohms. From the output characteristics shown in Fig. 3.11, the required base current I_b is 0·25 mA, and reference to the input characteristics shown in Fig. 3.12 indicates that the base emitter voltage required to produce this current is 0·19 volt. Adding this voltage to V_e gives $V_b = 1{\cdot}2 + 0{\cdot}19 = 1{\cdot}39$ volts and therefore R2 $= \frac{1{\cdot}39\text{V}}{3\text{ mA}} \times 10^3 = 463$ ohms. Since the total potentiometer resistance required has been calculated to be 2,000 ohms, R1 must be $2{,}000 - 464 = 1{,}537$ ohms.

This output stage is used in the 35 mW amplifier shown in Fig. 3.15 and it will be noticed that resistors of the nearest preferred values to those calculated are specified. Decoupling capacitors are used to isolate the bias circuit at signal frequencies.

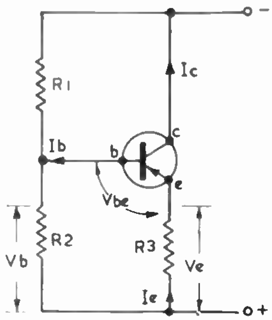


Fig. 3.14. The basic elements of a single battery bias circuit. Good stability of working point can be obtained with a bias system of this type at the expense of a relatively small extra battery drain.

COUPLING CIRCUITS FOR TRANSISTOR STAGES

Audio Frequency Stages

Because the transistor is a *current* operated device and has a relatively low impedance input circuit compared with the usual thermionic valve arrangement, it is more important to ensure proper impedance matching of the coupling circuits linking small signal transistor stages than in the corresponding valve circuits. The procedure is complicated by the fact that transistor impedances vary with operating conditions and it is necessary to establish the working point before attempting to calculate the coupling requirements. This will normally be chosen on the basis of allowing a sufficiently large quiescent collector current to handle the signal without any clipping of the current waveform.

For audio frequencies, the input and output impedances can then be determined by measuring the slope of the static characteristics at the working point discussed previously or by referring to spot values which the manufacturer may include in the published data. The turns ratio of the coupling transformer is then given by $N = \sqrt{\frac{Z_{out}}{Z_{in}}}$, where Z_{in} is the input resistance of the driven stage and Z_{out} is the output resistance of the preceding stage. Such low power coupling transformers can be made quite small, perhaps within a half-inch cube, but care must be taken to ensure that the d.c. resistance of the windings is not too high. Winding resistance is much more important than in intervalve transformers: for instance a voltage drop of five volts across the primary of an interstage transformer in a valve amplifier would be of little consequence but a similar drop in the collector circuit of a transistor stage could be disastrous when operating from a 6 volt supply. While on this point, it is worth noting that a similar situation may occur when high resistance headphones are used in transistor circuits.

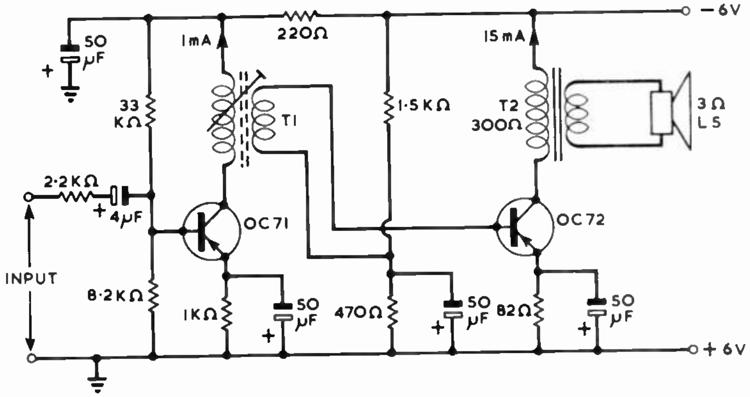


Fig. 3.15. A class A 35 mW audio output stage and driver. Many "personal" transistorized receivers use an output stage of this type.

The design of transformers for power output stages demands a rather different approach. Here the requirement is not to achieve maximum power *gain* from the stage, but maximum power *output*, and this entails a lower load impedance than for the matched condition. The class A output stage in Fig. 3.15 requires a collector load of 300 ohms and this is provided by a 10 : 1 step-down transformer feeding a 3 ohm loudspeaker. A typical specification for such a transformer is given in Table 3.1. Because the 35 mW output stage requires very little drive, the driver transformer can be designed on the basis of that for a small signal stage; but this may not always apply. In some cases where larger power outputs are required it will be necessary to compute the input resistance of the power stage, not on the basis of the slope resistance at the operating point but by considering the peak-to-peak input voltage and current requirements at full drive. By dividing the input voltage excursion by the current excursion a more realistic figure for input resistance is then obtained.

Resistance-capacitance coupling may also be used in transistor amplifiers but the stage gain is then not so high as with transformer coupling because the effective collector load can never be higher than the input resistance of the following stage. The amplifier thus relies solely on the current gain of the transistor, and the common base stage is therefore unsuitable for this type of coupling. However, common emitter and common collector stages using *RC* coupling are preferred for certain applications because of their simplicity.

TABLE 3.1

Transformer details for the circuit of Fig. 3.15

	T1	T2
Turns Ratio	10 : 1	10 : 1
Resistance of primary	≯ 250 ohms	≯ 10 ohms
Resistance of secondary	≯ 30 ohms	≯ 0.3 ohm
Primary inductance	≮ 20H	≮ 0·3H

Radio Frequency Stages

In the case of tuned radio frequency transistor amplifiers, such as those in the i.f. stages of a communications receiver, complications are added to the general problems of power matching. These arise because of the need to preserve adequate frequency selectivity and because of the presence of internal feedback within the transistor which become significant at higher frequencies and can result in instability.

To maintain selectivity it is necessary to tap the transistor electrodes into the tuned circuits in such a way that a reasonable working Q value is maintained. Fig. 3.16 illustrates the way in which this is usually achieved in a single-tuned i.f. transformer for use with an OC45 type transistor. Such a coil would be enclosed in a ferrite pot and the whole assembly, including the tuning capacitor, mounted in a small aluminium can about ¾ in. diameter. If, in an effort to preserve a high working Q value, the tapping points are brought nearer to the "earthy" end of the tuned winding, the insertion losses increase sharply and overall

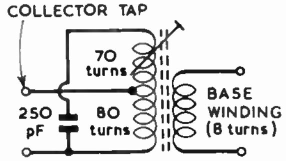


Fig. 3.16. Typical i.f. transformer for use with transistors. To maintain a satisfactory working Q value, the transistor electrodes are tapped into the tuned winding. The unloaded Q value is about 100 and the tappings are chosen such that the working Q value is about 50.

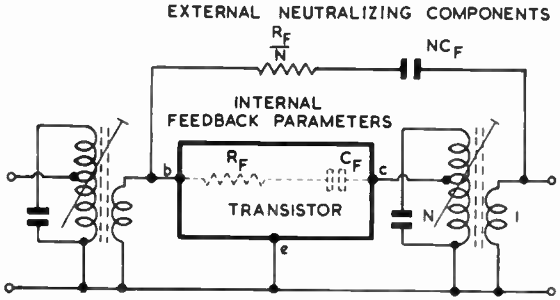


Fig. 3.17. Representation of a neutralized transistor i.f. stage. The internal feedback parameters of the transistor are neutralized by feeding back an antiphase signal to the input circuit via a series RC network. For perfect neutralizing, the external feedback components are related to the internal parameters by a factor equal to the ratio of the collector turns to base turns on the output i.f. transformer.

gain will decrease. However, the compromise between selectivity and gain is not as straightforward as it may seem because the maximum stage gain which can be safely used without risk of instability is closely tied up with the internal feedback capacitance of the transistor, i.e. the higher the capacitance, the lower is the maximum usable gain. This limitation can be considerably relieved if the internal capacitance is neutralized by feeding back an antiphase signal from an external circuit. The neutralizing feedback may be conveniently taken from the secondary winding of the following transformer but connections to the winding must be in the proper sense to provide the opposite polarity required.

To obtain complete neutralization, a capacitor in series with a resistor is required but the resistor may often be omitted in practice: see **Fig. 3.17**. Correct values for the neutralizing components can be determined by taking the manufacturer's published data on internal feedback parameters and applying a factor equal to the turns ratio of the transformer which is supplying the feedback voltage. For example, if the transistor feedback parameters are quoted as a capacitor of 12pF in series with a resistor of 3000 ohms, and if the primary-to-secondary turns ratio of the i.f. transformer is denoted by N, the neutralizing components should consist of a capacitor equal to $N \times 12$pF in series with a resistor of $\frac{3000}{N}$ ohms.

Perfect neutralizing will rarely be achieved and it is impracticable to design on that basis. A good approach is to assume that neutralizing error may be as high as 20 per cent of the nominal collector-base capacitance which, in the case quoted above, would reduce the net feedback capacitance from 12pF to about 2.5pF.

It is interesting to note that many v.h.f. transistor types on the market have actual internal feedback capacities of about 2pF, which is less than the neutralizing error which must be catered for on less sophisticated types. Obviously if such a low initial capacitance is neutralized the likely error is extremely low and a very carefully designed stage could achieve high power gain, but a more significant observation is that the low-capacitance transistor is capable of giving as much gain without a neutralizing circuit as the high-capacitance sample gives when carefully neutralized. For this reason, and because of their higher input and output resistances, v.h.f. transistors are becoming increasingly popular in i.f. amplifiers, even at the relatively low frequencies commonly used in broadcast receivers.

As a guide to i.f. and r.f. transformer design, **Table 3.2** lists suggested base circuit resistance and collector circuit resistance values at various frequencies for a common emitter stage operating at 1 mA and with a net feedback capacity of about 2pF. For most tuned transformers, the resistance of a winding at a tapping point may be calculated from $R_{TAP} = \frac{R_D}{N^2}$, where R_D is the dynamic resistance of the whole tuned winding at the frequency concerned $\left(R_D = \frac{Q}{2\pi f C_{TUNE}}\right)$, and N is the ratio of total turns to tapped turns.

TABLE 3.2

Suggested base circuit and collector circuit resistances at various frequencies for a common emitter stage operating at collector current of 1 mA and with a net feedback capacitance of 2 pF.

Frequency	Base circuit resistance at stated frequency	Collector circuit resistance at stated frequency
110 kc/s	200 ohms	50 K ohms
470 kc/s	100 ohms	20 K ohms
1·6 Mc/s	60 ohms	10 K ohms
10·7 Mc/s	30 ohms	3 K ohms

Complementary Symmetry Amplifiers

To supply the drive requirements of a push-pull pair, some form of phase-splitting arrangement is commonly involved. For many conventional amplifiers this takes the form of a transformer, but transistors now permit this requirement to be avoided by using devices of opposing polarity (i.e. *n-p-n* paired with *p-n-p*) in the driven stage so that the input electrodes can be common and driven in phase with each other. **Fig. 3.18** shows an amplifier of this type which is designed to deliver about 300 mW into a speaker. It can be seen that apart from the speaker there are no bulky components, the elimination of transformers greatly reducing the weight and size of the unit. There are, however, far more significant rewards than this because iron-cored transformers have many undesirable electrical characteristics and their removal opens the way to new standards of sound quality.

Apart from the direct effects of transformer distortion, difficulties often arise due to phase shift within the component at the higher audio frequencies. This sets a limit to the amount of negative feedback which can be safely employed and may mean that distortion introduced by other components cannot be reduced to the desired level.

In Fig. 3.18, the driver transistor is an AC113 in class A common-emitter configuration. The collector is d.c. coupled to the output transistors which have a d.c. and a.c. negative feedback path to the driver stage input via a preset resistor which is used to adjust initial working conditions. The preferred method of adjustment is to examine the output waveform on an oscilloscope and set the resistor to the position where the waveform clips symmetrically under overload conditions, but if this facility is not available it is sufficient to adjust to give a quiescent voltage of 5·0 volts at

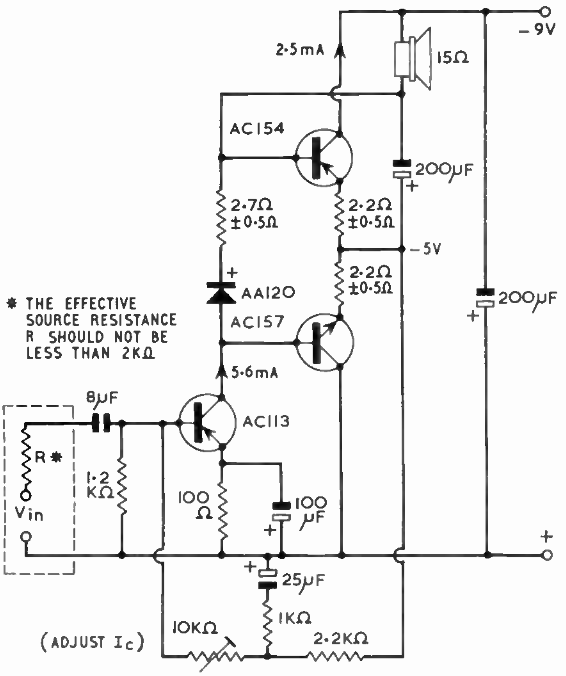


Fig. 3.18. A 300 mW class B push-pull amplifier using *n-p-n* and *p-n-p* transistors in the complementary symmetry configuration. This arrangement avoids the need for a phase splitting driver transformer and leads to a completely transformerless a.f. amplifier.

the junction of the two 2·2 ohm resistors in the emitter circuits of the output transistors. The AA120 germanium junction diode in the driver collector circuit is operating in the forward direction and helps to stabilize working conditions against the effects of temperature change and variations of supply voltage.

An AC154 *p-n-p* transistor and an AC157 *n-p-n* transistor form a matched output pair, it being essential that their characteristics are complementary over the entire working range of current and voltage. This is very important and it is of little use to pair *n-p-n* with *p-n-p* devices at random in this application. The 15 ohm loudspeaker is fed via a 200μF electrolytic capacitor which is sufficiently large to ensure good bass response. The 9 volt supply is shunted by a similar capacitor to provide a low-impedance path for the high peak currents involved.

The amplifier of Fig. 3.19 is similar to the circuit just discussed but uses a coupling arrangement which partly avoids the problem of obtaining matching between *n-p-n* and *p-n-p* transistors at high peak currents and opens the way to much higher power output levels. Each half of the push-pull output stage consists of a pair of transistors connected in the so-called super β or Darlington configuration. Such a pair behaves rather like a single transistor of very high current gain and furthermore, by using a low power *n-p-n* transistor coupled with a high power *p-n-p* transistor, the combination behaves as though it were a high power *n-p-n* device. It is thus possible in a complementary push-pull stage to obtain relatively high output power levels without the difficult requirement of matching the characteristics of an *n-p-n* power transistor to those of a *p-n-p* power transistor. Amplifiers of this type, delivering output powers of many tens of watts, are replacing thermionic valve amplifiers in high quality sound apparatus.

In the amplifier shown, it is very important that the AF167 output transistors are mounted in an effective heat sink to ensure satisfactory operation up to an ambient temperature of 45° C and also that the AA120 compensating diodes are mounted on the same heat sink so that their temperature is closely related to that of the transistors whose conditions they are intended to stabilize. The preset resistor should ideally be adjusted to give symmetrical overload conditions as described for the previous amplifier but alternatively may be set to give the values of quiescent collector currents shown on the diagram.

Because the output pairs are in series connection across the supply line, each pair operates from only half of the available supply voltage. This explains why the unusually high level of 26 volts is used on this and many similar amplifiers. For operation from a mains supply the power unit need consist only of a transformer-fed bridge rectifier circuit with a single reservoir capacitor of 2000μF. No other smoothing is required. If the amplifier is being operated from batteries it is advisable to use a capacitor of similar

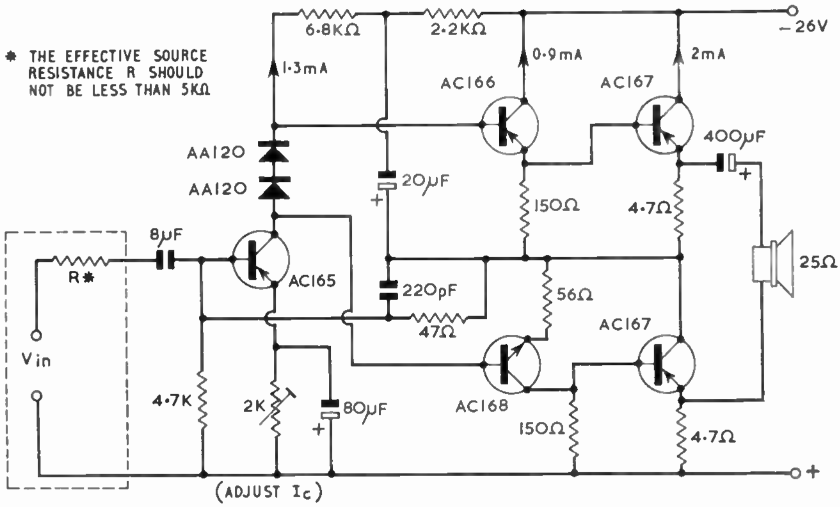
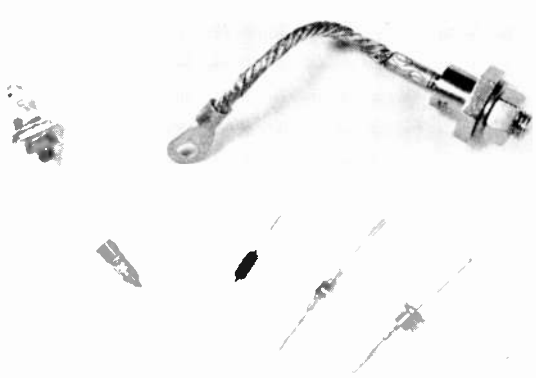


Fig. 3.19. A 2 watt transformerless a.f. amplifier using a modified complementray symmetry arrangement which permits the use of matched *p-n-p* transistors in the final power stage. This is advantageous because of the difficulty of matching an *n-p-n* device to a *p-n-p* transistor at high power levels.



Semiconductor diodes. Top row, power rectifiers; below, signal-diodes.

value to shunt the supply to avoid premature overload.

The frequency response of the amplifier is substantially level between 25 c/s and 10 kc/s and the total harmonic distortion at maximum rated output is less than 1 per cent.

SEMICONDUCTOR DIODES

Semiconductor Power Rectifiers

Rectification always involves a certain loss of power in the form of heat. In the case of power supplies for radio equipment, the amount of heat to be disposed of may be significant enough to influence markedly the mechanical design if thermionic diodes or selenium diodes are used as the rectifying elements.

By using germanium or silicon junction diodes, however, rectification efficiency can be so greatly improved that it becomes possible for a diode no larger than a flashlamp bulb to rectify the h.t. power for a 100 watt transmitter. Examples of such diodes are shown in the photographs and it can be seen that they are small wire-ended components which can be hung in the wiring in much the same way as one would mount a resistor or capacitor of comparable size. Diodes for larger power requirements may be fitted with means for clamping to a metal cooling plate.

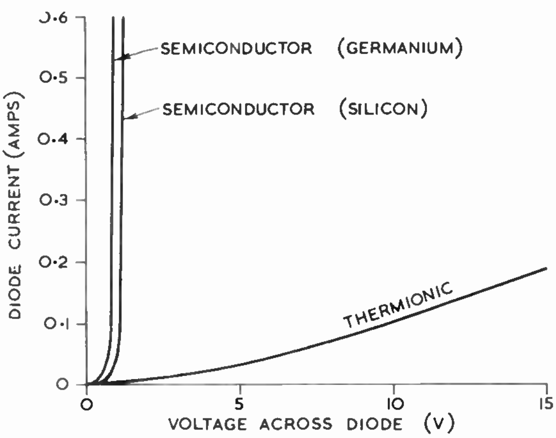


Fig. 3.20. Comparison of the voltage drops across semiconductor and thermionic power rectifier diodes.

Most of the losses in a power rectifying diode are incurred during the conductive part of the cycle by the passage of current through the forward resistance of the device. Fig. 3.20 compares the forward characteristics of a thermionic diode with those of similarly rated silicon and germanium devices and clearly shows the advantages to be obtained in this respect. The apparent advantage of germanium over silicon is largely offset by its lower maximum working temperature (90° C against about 150° C for silicon) and also by inferior reverse characteristics.

Apart from reducing device consumption, lower forward resistance results in higher d.c. output voltage for a given a.c. input level, but also involves higher peak forward currents in the diode circuit, particularly if a capacitor input filter is used for smoothing. In the case of most transformer-fed rectifying circuits, the resistance of the transformer windings will limit the charging current to a safe value, but this feature should be considered carefully at the design stage and reference made to the manufacturer's data to establish minimum permissible circuit resistance for the value of reservoir capacitor envisaged. Any extra resistance required should then be connected directly in series with the diode.

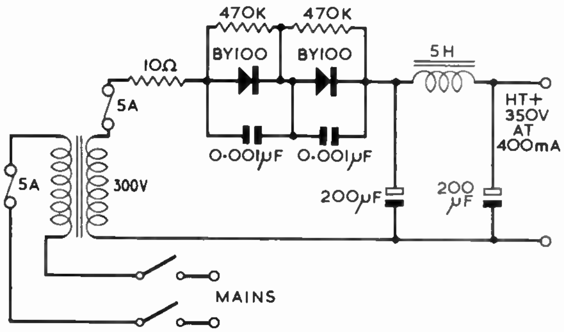


Fig. 3.21. Circuit of a typical power supply using two silicon rectifiers type BY100 in series to provide 350 volts at 400 mA. The supply embodies features discussed in the text.

One of the difficulties encountered in the development of semiconductor power diodes was the high reverse voltage rating required in a single junction if the need for several units to be connected in series was to be avoided. To provide a suitable safety factor, a reverse voltage rating of at least three times the r.m.s. value of sine wave input is required. Junctions having voltage ratings of 800 to 1,000 volts are being produced cheaply and are in common use in domestic television receivers but, if it is necessary to operate diodes in series, rather special precautions must be taken to ensure equal distribution of voltage between samples. Inequalities arise during the non-conducting period due to differences in reverse resistance and should be reduced by shunting each diode with identical fixed resistors of sufficiently low value to "take-charge" of reverse voltage distribution yet not low enough to cause excessive heat dissipation within the resistors themselves. For precise adjustment, knowledge of the reverse leakage tolerances on the diodes in question will be required but for most amateur radio purposes, matched resistors in the order of 500 K ohms will be adequate.

Further precautions may be required in order to protect the diodes against permanent damage due to short-duration pulses such as interference spikes which may " ride " on the peaks of the incoming mains waveform and cause excessive reverse voltage. Unlike thermionic diodes, semiconductor devices may be very susceptible to this form of overload, but protection may be provided by shunting each diode with a capacitor of about 0·001μF, bearing in mind the working voltage requirements.

A typical power supply incorporating these safety precautions is shown in **Fig. 3.21**.

A recent development in the semiconductor field is the device known as the controlled avalanche rectifier which is similar to the diodes discussed above but has a specially shaped reverse characteristic which allows the rectifier to absorb more transient energy. The so-called avalanche characteristic is particularly useful for series operation of rectifiers, especially when a choke-input filter is used or when interference spikes are plentiful.

If a single diode will not deliver the *current* requirements, it will be necessary to operate two or more in parallel. The danger then is of unequal distribution of forward current which may have to be corrected by the insertion of a low-value resistor in series with each diode. Any doubts about the well-being of diodes in this respect can be met by including a suitable fuse in series with each device.

It is assumed that most power rectifying equipment will be operating from the domestic mains supply, but occasionally it may be required to rectify waveforms at frequencies several times higher than 50–60 c/s. The use of semiconductor power diodes may present some difficulty here because of a phenomen known as " hole storage " which causes a delay in switching off the diode after each period of conduction. When the delay becomes significant, compared with the duration of the applied waveform, loss of efficiency arises and, perhaps more important, voltage inequalities between diodes in series may be exaggerated. The latter effect may be relieved somewhat by adding shunt capacitors of values similar to those discussed in connection with protection against interference spikes.

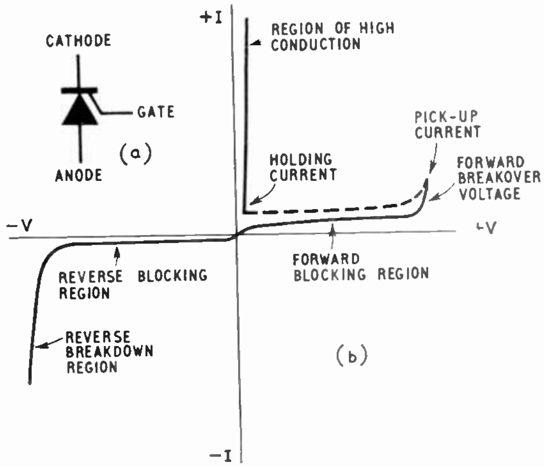


Fig. 3.22. (a) Circuit symbol for a silicon controlled rectifier (b) Electrical characteristics of an SCR. Application of a relatively small signal to the gate electrode can switch the device from the forward blocking region to the high conduction region. The blocking condition can only be restored by reducing the forward current below the holding level. The action can be likened to that of a thyratron valve.

The significant difference between a switching transistor and an SCR is that a transistor does not exhibit the latching action of a SCR and requires a much higher control current. Whereas a base current of perhaps 1 amp would be needed to control a collector current of 20 amps, the required gate current in an SCR would be about 50 mA. Because the SCR, by virtue of its construction, is able to make more effective use of the junction area for current conduction, it is somewhat smaller in physical size than a transistor of comparable rating but will need to be fixed to a heat sink and is usually fitted with a stud for this purpose.

Although it seems unlikely to find extensive use in amateur radio apparatus, the SCR is proving to be a very versatile device, with application to such items as controlled power supplies, power regulators, d.c. to a.c. inverters and time bases for cathode ray tube scanning.

Silicon Controlled Rectifiers

The silicon controlled rectifier is much the same as an ordinary silicon rectifier except that it is " blocked " in the forward direction until switched on by a small signal applied to a third electrode termed the *gate*. Once switched on it will conduct even after the gate signal is removed and will switch off only when the forward current has been reduced below a critical level known as the holding current.

As can be seen from the static characteristics shown in **Fig. 3.22**, the SCR can also be switched on by increasing the voltage across the main electrodes to the level where avalanche breakdown begins to occur, but in typical operation the applied voltage is kept below this point and triggering is accomplished by injecting current into the gate lead. The action is very similar to that of a gas thyratron valve but whereas the thyratron is essentially voltage triggered, the SCR is current triggered. This must be borne in mind when designing firing circuits because the SCR requires a low impedance source.

Zener Diodes

Zener diodes are semiconductor diodes with specially shaped reverse characteristics which enable them to be used as stable voltage references and are named after the discoverer of the effect upon which their operation depends.

Typical reverse characteristics of such a device are shown in **Fig. 3.23**. As reverse voltage is increased from zero, the diode behaves in much the same way as other semiconductor junctions until a critical voltage known as the Zener voltage is reached when a phenomenon known as Zener breakdown occurs and current begins to rise very steeply as voltage is further increased. In this region, provided the power dissipation limits are not exceeded, the diode may be used in a similar way to a neon stabilizer tube, with the advantage that a range of types with working voltages between three and 200 is available. The most common types operate at about 4–8 volts and it is worth noting that diodes in the range 5–6 volts can be made to have a temperature co-efficient of near zero at room temperature. This means that the Zener

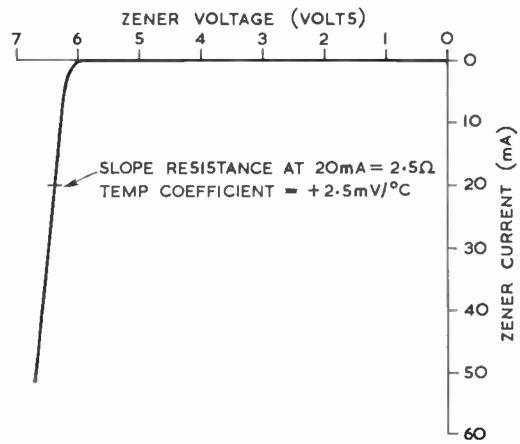


Fig. 3.23. Static characteristics of a typical low voltage Zener diode at 25°C. The effect of increasing the temperature is to displace the curve along the voltage axis in a direction depending on the polarity of the temperature coefficient. Reverse current at voltage below the Zener level is about 30μA.

voltage will be substantially independent of temperature over a useful range; 5–6 volts may therefore be the best stabilizing level to select if a choice is available when designing equipment.

Fig. 3.24 shows the basic manner in which the diode is used. The value of *R* should be chosen to pass the required nominal diode current which, for many purposes, will be about half of the maximum diode rating, typical values being about 50 mA for moderate sized devices. The order of voltage regulation obtainable is about 0·01 per cent change of output voltage for an input level change of 1 per cent.

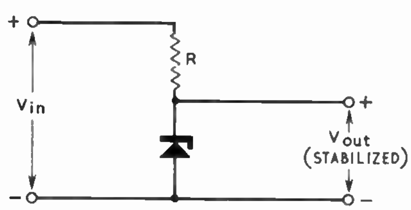


Fig. 3.24. Basic method of connecting a Zener diode to produce a stabilized voltage. The value of R should be chosen, after reference to the published characteristics, to give a diode current which will accommodate expected load and supply variations without exceeding the dissipation limits of the diode.

Possible uses include stabilizing working conditions of transistor amplifier stages, providing constant bias supplies for thermionic valves and providing voltage references in stabilized power supplies.

Variable-capacitance Diodes

Yet another application for the *p-n* junction diode is as a voltage-dependent capacitor in radio or electronic circuitry. As was explained earlier in this chapter, when a junction diode is biased in the reverse direction a region at the junction becomes swept clear of current carriers and is known as the *depletion layer*. The thickness of the region increases as the applied voltage is increased and can be considered as an insulator between two electrodes. The capacitor thus formed is therefore voltage-dependent and, in the case of a particular commercial sample, has a capacitance which is approximately proportional to $\frac{1}{\sqrt{V}}$, where *V* is the reverse bias voltage. Although any *p-n* junction will exhibit this effect, devices with carefully controlled parameters are produced specifically for this purpose and find application for processes such as automatic frequency control, or for parametric amplification (see Chapter 5).

The characteristics of a variable capacitance diode are given in Fig. 3.25.

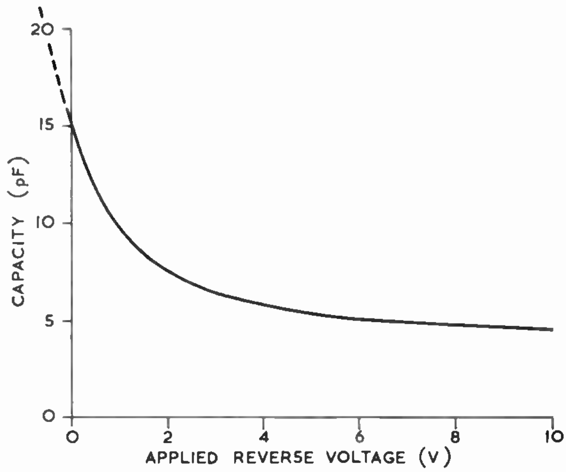


Fig. 3.25. Variation of capacitance with voltage in a variable capacitance diode. Although it is more usual to operate in the reverse bias mode, the characteristic does extend into the forward bias region where care must be taken not to exceed the maximum dissipation limit.

The Tunnel Diode

A device which may prove superior to the transistor in certain v.h.f. or u.h.f. applications, and which is undergoing intensive development, is the tunnel diode. This derives its name from the quantum-mechanical tunnelling effect on which its operation depends and promises to be useful at frequencies about 100 times greater than current transistors, yet requiring only a fraction of the power.

Briefly, it consists of a *p-n* diode made from heavily-doped

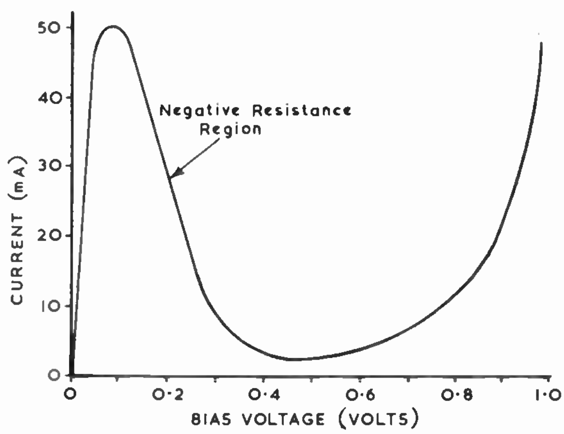


Fig. 3.26. Static characteristics of typical gallium-arsenide tunnel diode.

materials and with an extremely thin junction region. Unlike the transistor it operates largely on the passage of *majority* carriers across the junction, and a signal is thus transmitted through the diode at much the same speed as it would be through a metallic conductor, hence the excellent high-frequency performance.

When biased to a particular part of the characteristic, the device presents an incremental negative resistance, see **Fig. 3.26**, and thus may be treated as a two-terminal amplifier. At present, a major disadvantage lies in the difficulty of designing coupling circuits for cascaded stages. With a negative resistance characteristic perhaps extending up to a frequency of thousands of megacycles the risk of spurious oscillations occurring can be readily appreciated. The problem is largely tied up with the method of making connection to the diode junction, and the type of encapsulation becomes extremely important.

Transistors at U.H.F.

The general pattern of events has been for transistors to compete most effectively with valves at the lower frequencies, expansion of use being broadly related to improvement of high frequency performance. More recently the trend has been accelerated in a perhaps unexpected fashion by the introduction of germanium transistors which will give better noise performance than competitive valves at frequencies in the range 400–900 Mc/s.

The stimulus for this development has undoubtably been the spreading of domestic television transmissions into the u.h.f. bands where atmospheric and man-made interference

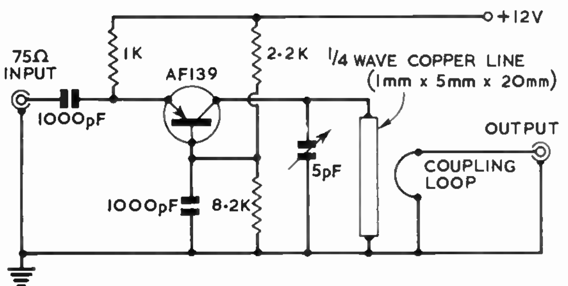


Fig. 3.27. Transistor u.h.f. amplifier tuneable from 500 to 800 Mc/s. The size and position of the output coupling loop should be adjusted to suit the load. The power gain is approximately 10db at 800 Mc/s and the noise factor about 8db.

effects are so much less troublesome than on the v.h.f. bands that receiver noise performance sets the limit to usable sensitivity. The order of improvement currently being obtained by using the AF139 transistor instead of the PC88 valve for example is about 2–3db.

To achieve this the AF139 is operated in the common-base mode at a collector current of about 2 mA. A 75 ohm unbalanced aerial circuit may be fed directly into the emitter, while the collector circuit is usually tuned by a quarter wave line. Ajustment of tuning may be by variable capacitor across the end of the line or by a sliding short-circuit to vary the line length. Power gain is typically 10db at 800 Mc/s. Noise factor varies between about 6 to 8db but, if the stage is followed by a self-oscillating mixer using the same type of transistor, the overall noise performance will deteriorate by about 1–2db due to the intrusion of mixer noise.

TABLE 3.3

Commonly-used symbols associated with transistor parameters.

Symbol	Parameter
α or h_{fb}	small-signal current gain, common base
β or h_{fe}	small-signal current gain, common emitter
h_{FE}	d.c. current gain, common emitter
I_{ceo}	collector leakage current, common emitter
I_{cbo}	collector leakage current, common base
V_{ce}	collector-emitter voltage
V_{cb}	collector-base voltage
V_{be}	base-emitter voltage

The circuit of a simple unneutralized tuneable 500-800 Mc/s r.f. stage is shown in **Fig. 3.27.**

PRACTICAL APPLICATION OF TRANSISTORS

Transistors, although mechanically quite robust, can easily be damaged or destroyed by careless use and before attempting to build or service transistorized apparatus it is therefore wise to bear in mind the following points:

(a) Never apply prolonged heat when soldering the connecting leads of a transistor without using a thermal shunt between the transistor and the soldering iron. A pair of pliers will do, but it is more satisfactory to fabricate a shunt by soldering a pair of copper plates to the jaws of a crocodile clip.

(b) Do not cut the transistor leads shorter than absolutely necessary. Remember that the thin wire used for these leads can break off close to the transistor.

(c) Do not use an ohmeter on transistor circuitry without first removing the transistors. Apart from the possibility of obtaining false ohmeter readings, there is danger of damaging the transistors.

(d) Do not apply excessive signal amplitude from test oscillators or signal generators. Some test apparatus may be capable of delivering sufficient power to damage the transistors. Start with a small signal and increase it until the desired indication is obtained.

(e) Always check connections carefully before switching on power to transistor equipment for the first time. Pay particular attention to correct battery polarity.

A simple transistor tester is described in Chapter 19 (*Measurements*). On no account should a multimeter of less than 20,000 ohms per volt sensitivity be used to measure the internal resistances of a transistor.

Development Trends

The factors which chiefly determine high frequency performance are, first, the finite time taken for current carriers to diffuse across the base region of the transistor, and secondly, the internal feedback path between output and input circuits.

In order to minimise the distance through which current carriers must travel, several involved manufacturing techniques have been developed for reducing the width of the base region, and there are a number of types in which improved high-frequency performance has been obtained solely in this way. As an extension of this process, the drift transistor has a base region which is not only very narrow but has a graded impurity concentration across its width. This has the effect of providing an accelerating force to the carriers, which then traverse the base region more rapidly than if they were merely diffusing through the crystal lattice.

Among various interpretations of this technique, the alloy-diffused transistor has a graded base region with base widths of only a few microns. Another line of development has been to control the area over which diffusion takes place in order to reduce the capacitance of a junction.

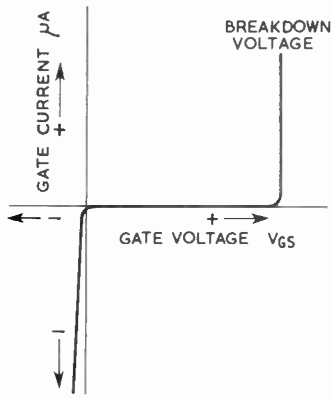


Fig. 3.30. Gate diode input characteristic.

Field Effect Transistors

In this type of device the connections are known as gate, source and drain. For normal circuits the source can be considered to be the terminal common to the input and output is comparable to the emitter of a bipolar transistor. The drain is the output terminal equivalent to the collector and the control electrode, the gate is similar in operation to the base except that the FET is a voltage controlled device instead of current controlled as is the case in a bipolar transistor.

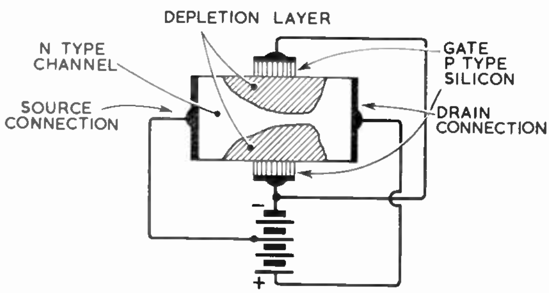


Fig. 3.28. Construction of n-type channel.

The general form of a junction FET is shown in **Fig. 3.28.** It consists of a piece of P or N type material, usually silicon, to which two ohmic connections are made for the source and drain electrodes and having *p-n* or *n-p* junctions sited so that the depletion layers existing around the junctions can extend into the path between the source and drain connections. **Fig. 3.29** shows the corresponding drawing symbols.

When the gate is sufficiently reverse-biased the depletion layers will extend into the channel and meet, cutting off the source-to-drain current. This is the "pinched-off" state, and the gate source voltage V_p at which it occurs is known as the "pinch-off" voltage.

If, however, the gate-to-source voltage is reduced to zero, the maximum possible drain current flows, denoted by I_{dss}. At this point, the width of the depletion layer in the channel has been reduced to zero. Conversely, if the gate-to-source voltage is increased beyond V_p, no greater drain current than I_{dss} can flow since the gate source diode conducts, clipping off the excess gate-to-source voltage **(Fig. 3.30)**. The useful portion of the junction FET mutual characteristic therefore occurs between the limits of zero and V_p for the gate-to-source voltage. This limitation can be removed by employing one of the forms of INSULATED-GATE construction, in which a very thin layer of glass is interposed between the gate and the channel to form an insulator, as in IGFETs and MOSFETs. It is then possible to arrange that no drain current flows until the gate-to-channel is forward biased.

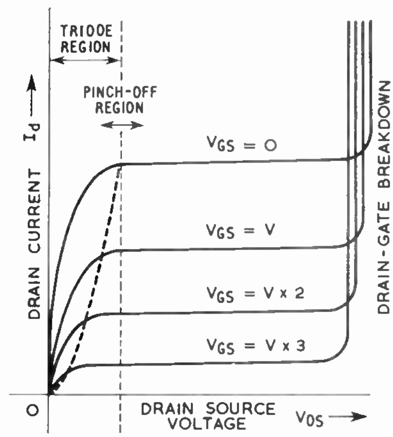


Fig. 3.31. Drain characteristic curves.

Fig. 3.29. Standard FET symbols.

These are called ENHANCEMENT-MODE devices: those which operate like normal junction FETs in which reverse bias may be used to reduce the drain current are called DEPLETION-MODE types. Finally, there are DUAL-GATE FETs, which may be operated as tetrodes. These are of value for mixing and for the application of gain control voltages to an amplifier stage handling large signals.

The resistance of an *n* or *p* type channel, even when no bias is applied to the gate, is very much higher than the equivalent emitter-to-collector resistance of a bipolar

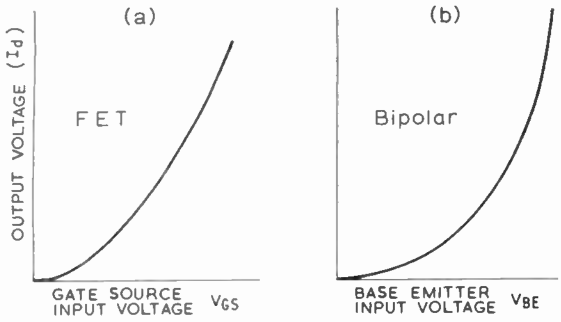


Fig. 3.32. Comparison of FET and bipolar transistor characteristics.

transistor when "bottomed". The channel resistance at "pinch-off" is many times greater than that of a bipolar transistor at cut off and the leakage current is very much lower.

An *n*-type channel is shown in Fig. 3.28. With a *p*-type channel the supply polarities are reversed, the drain being connected to a potential negative with respect to the source.

FET Characteristics

Fig. 3.31 shows a typical family of drain voltage—drain current curves. As can be seen there is a great similarity to those of the pentode valve.

With the gate tied to the source, the drain current increases with the drain voltage until a maximum value of drain current is reached. This occurs at a drain voltage approximately equal to V_p. When the gate bias voltage is increased the maximum value of drain current is reduced as shown.

The relationship between the value of drain current (Id) and the gate-to-source voltage (Vgs) is square law having a mutual characteristic as shown in **Fig. 3.32(a)**. This should be compared with the same characteristic of a bipolar transistor shown in **Fig. 3.32(b)**.

An important feature of an FET in which it is similar to the normal thermionic valve, is that the ratio of a small change in gate voltage to the resulting change in drain current for a constant drain voltage is the mutual conductance gm. Typical values are of the order of 3mA/V.

The amplification factor μ is defined as follows. If the gate voltage is increased by a small amount and the drain voltage is changed to restore the drain current to the original value, the amplification factor is the ratio of the drain voltage change to the gate voltage change. As with the valve the amplification factor is the product of the drain incremental resistance and the mutual conductance.

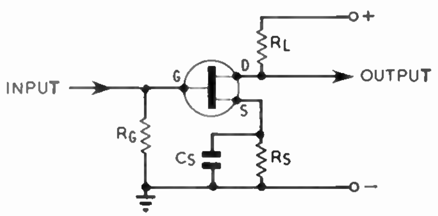


Fig. 3.33. Typical FET amplifier circuit.

Basic Circuits

One of the obvious applications of FETs is for high impedance input stage amplifiers, either at audio or i.f. frequencies, the latter allowing valve type transformers to be used.

The gain of an amplifier such as shown in **Fig. 3.33** is approximately equal to the product of the mutual conductance and the load (R_l). The source resistor (R_s) and the capacitor (C_s) function in much the same way as the cathode resistor and capacitor of a valve amplifier.

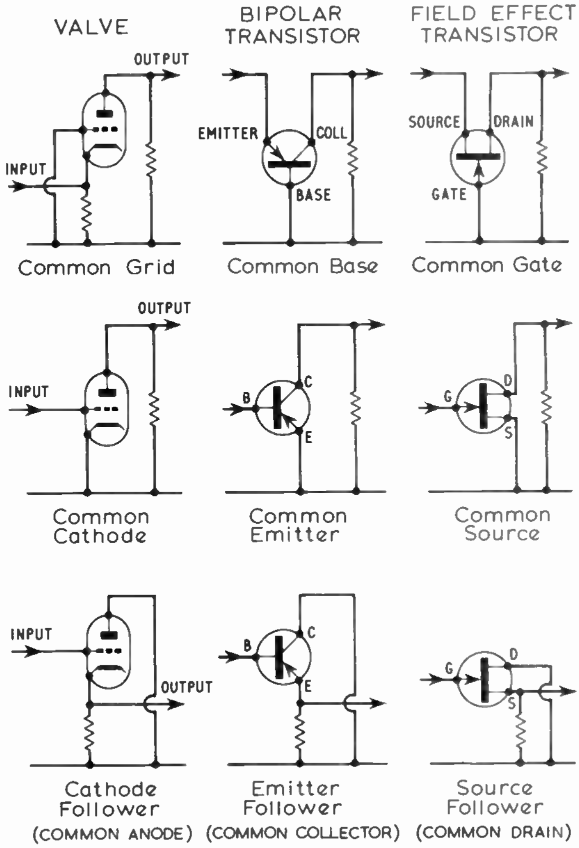


Fig. 3.34. Equivalent configurations.

The FET functions as an almost ideal mixer since the square law characteristic is superior to that of either the bipolar transistor or a valve. This is because the mutual characteristic of an FET consists almost entirely of linear and square law terms, higher order terms being almost entirely absent, in contrast to valves and even more to bipolar transistors where higher order terms greater than the second, lead to severe distortion in the mixing process with consequent generation of spurious signals.

Fig. 3.34 shows the general relationship between valves, bipolar and field effect transistors.

CHAPTER 4

H.F. RECEIVERS

RECEIVERS for Amateur Radio purposes vary widely in complexity, from simple two valve designs to sets with up to about 20 valves or even more transistors, but whether simple or complex, home-constructed or factory-built, are unlikely to prove satisfactory unless careful attention has been paid to the many problems presented by modern two-way h.f. operation.

The radio amateur needs a *communications receiver* which will provide maximum intelligibility from signals having extremely wide variation in strength (incoming signals may easily differ in voltage by 10,000 times and occasionally by up to about 500,000 times) in restricted frequency bands which are often extremely congested. Ideally, each main mode of transmission—telegraphy (c.w.), and telephony by amplitude modulation (a.m.), single sideband (s.s.b.), double sideband (d.s.b.), narrow band frequency modulation (n.b.f.m.) or RTTY—requires a receiver having different response characteristics. In practice, however, the amateur usually wishes to use only a single receiver for all purposes, on all amateur bands between 1·8 and 30 Mc/s. All but the most specialized designs are therefore likely to include some measure of compromise in order to provide a reasonable degree of flexibility.

The main requirements for a satisfactory receiver for the h.f. bands are:

(a) High *sensitivity* and *signal-to-noise ratio*, to permit the reception of weak signals.

(b) Good *selectivity*, to permit the selection of the required signal from among others—possibly much stronger—on closely adjacent frequencies. The degree of selectivity should preferably be variable to suit the differing requirements of the various modes of transmission.

(c) Maximum freedom from *spurious responses*.

(d) A high order of mechanical and electrical *stability*.

(e) Adequate, preferably calibrated, *bandspread* tuning of the amateur bands, coupled with the ability to reset the receiver accurately and quickly to a desired frequency. The *tuning rate* should be sufficiently low to facilitate the exact tuning of s.s.b., d.s.b. and c.w. signals.

(f) Internal *beat frequency oscillator* (b.f.o.) for the reception of c.w. and suppressed carrier signals.

(g) Sturdy and reliable construction and components providing ease of repair, alignment and general maintenance.

These requirements can be met using either valves or semiconductors or a mixture of the two. Later in this chapter, the particular problems posed by transistors in high performance receivers will be discussed. Since, in the majority of cases, amateur station receivers still tend to use valves, most of the circuits described employ valves; however, it should be appreciated that for almost all applications roughly equivalent circuits exist for transistors, although they will differ in detail.

At the present "state-of-the-art" it can be summarized that simple receivers using transistors are probably easier and more economical to construct than with valves; but that for highest performance, transistor designs are likely to prove rather more complex, mainly owing to the appreciably greater difficulty in achieving a high *dynamic range*—that is, the ability to cope with both very weak and very strong signals—due to the tendency of transistors to overload more easily.

BASIC TYPES OF RECEIVERS

Regenerative Detector (" Straight ") Receivers. The simplest practical receiver for amateur communication is a regenerative detector followed by one or more stages of a.f. amplification (0-V-1 etc.): see **Fig. 4.1(a)**. Because of the high gain which can be achieved in a correctly adjusted regenerative detector when set to a point just beyond that at which oscillation begins, this type of receiver can be used on surprisingly weak c.w. signals. Since the gain is much reduced as the regeneration (" reaction ") control is backed off, it will receive only fairly strong telephony stations. Because of regenerative feedback, the effective Q of the single tuned circuit and hence its selectivity, is reasonably good for c.w. reception, provided that the tuned circuit is not heavily damped by the effect of the aerial. The selectivity, however, is much reduced in the presence of strong signals on adjacent frequencies, and even in the most favourable circumstances cannot be made as good as in a modern superhet receiver. A further disadvantage is that the aerial is directly coupled to the oscillatory circuits and so interference may be caused to other receivers. This form of receiver was popular in the early days of short wave work and even now can provide a useful introduction to receiver construction.

Tuned Radio Frequency (T.R.F.) Receivers. Considerable improvement can be obtained in a regenerative detector receiver by including a stage of tuned radio frequency amplification before the detector stage: **Fig. 4.1(b)**. This provides greater gain, improves selectivity (mainly because the detector tuned circuit can be less heavily damped), isolates the regenerative detector from the aerial and so much reduces radiation from the receiver.

Although the selectivity of a t.r.f. receiver is deficient by modern standards and it is now rare to find such sets retained for normal amateur operation, it has some advantages for the home constructor:

(a) Simplicity of construction with no i.f. alignment or oscillator tracking problems.

(b) Good signal-to-noise ratio on weak c.w. signals.

(c) Complete freedom from spurious responses.

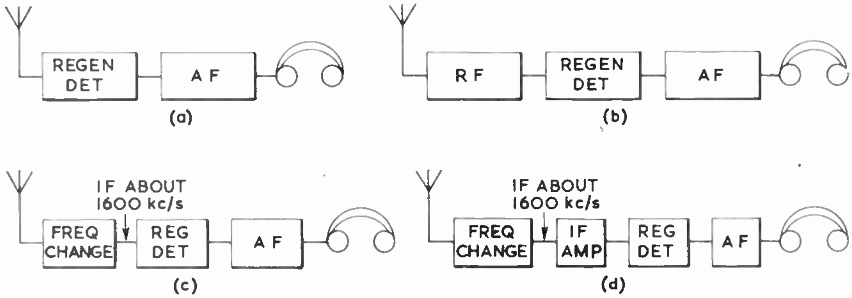


Fig. 4.1. Block diagram of simple h.f. receivers suitable primarily for Morse reception. (a) Regenerative detector followed by one or more stages of a.f. amplification. (b) The addition of a tuned r.f. amplifier improves selectivity and sensitivity and also reduces radiation from the receiver. (c) Simplest practical superhet receiver. By using triode pentode valves (ECF82/6U8) all stages can be accommodated in a two-valve arrangement. (d) The addition of one or more i.f. amplifiers greatly increases gain and permits the receiver to be used for telephony reception.

Overall gain when the detector is not oscillating is usually insufficient to provide good telephony performance on very weak signals.

Super-regenerative Detector Receivers. The high gain of a regenerative detector can be further increased and utilized for telephony reception by introducing a voltage at a supersonic frequency in such a way that r.f. oscillation ceases every half cycle of this second frequency, known as the *quench* frequency. This quenching voltage can be generated in a separate valve or, more commonly, in the regenerative detector stage. Although extremely high gain in a single stage is possible, this type of receiver has high interstation noise and poor selectivity while additional complications are necessary to receive c.w. or to avoid the emission of a rough signal capable of causing interference over considerable distances. Super-regenerative receivers are still of interest for simple portable equipment on 28 Mc/s and above but have little application on the lower frequency bands.

Superhet Receivers. By changing the frequency of the incoming signal to a lower fixed *intermediate frequency* in one step (single conversion), two steps (double conversion) or three steps (triple conversion), it becomes possible to build a high gain amplifier of controlled selectivity. Although the mixing process by which the frequency of the incoming signal is changed is inherently more noisy than amplification at the signal frequency, this disadvantage can be overcome by including one or more radio frequency amplifiers ahead of the first mixer.

The main practical disadvantages of the superhet are its susceptibility to various forms of spurious response, and the relative complexity of construction and alignment. The superhet principle is now used in the vast majority of amateur communications receivers and this class will receive more detailed treatment in this chapter than other types of receiver. The simplest forms of superhet are shown in **Fig. 4.1 (c)** and **(d)**. The block diagram of a typical superhet receiver for amateur communication is shown in **Fig. 4.2** with the main amplification in the i.f. stages. The importance of achieving high selectivity for Morse reception, permitting the rejection of the audio " image " signal, is illustrated in **Fig. 4.3.**

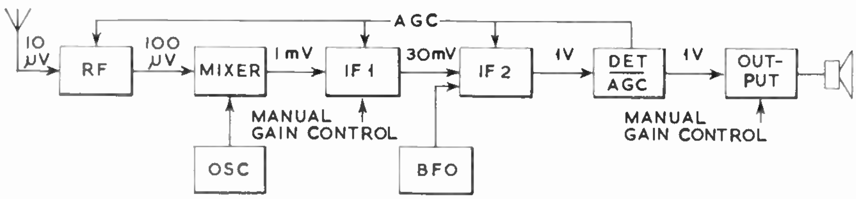


Fig. 4.2. Block diagram of a typical superhet h.f. communications receiver showing typical pattern of voltage gain. The intermediate frequency (i.f.) is usually of the order of either 470 kc/s or 1600 kc/s.

Selectable Sideband Receivers. As explained in Chapter 1 (*Principles*), conventional a.m. signals radiate sets of sidebands on each side of the carrier frequency, up to plus or minus the highest transmitted audio frequency. Since the two sets of sidebands are mirror images of one another, either set of sidebands can be completely removed without affecting the intelligence of the transmission, though (see Chapter 15—*Noise*) there will be a slight reduction of signal-to-noise ratio. In a crowded band it will be highly advantageous to restrict the receiver bandwidth so that the set responds to the carrier and only one set of sidebands and then to tune to whichever is the less affected by interference. Since the interfering stations may change at any moment, it will be clearly beneficial to arrange the set so that it can immediately be

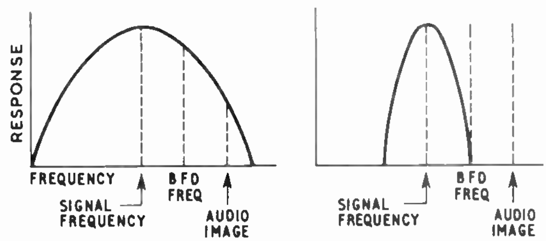


Fig. 4.3. How a really selective receiver provides single-signal reception of Morse signals. The broad selectivity of the response curve on the left is unable to provide substantial rejection at the audio image frequency whereas with the more selective curve on the right, the audio image is inaudible.

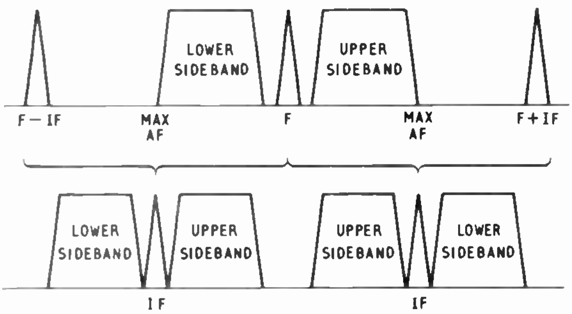


Fig. 4.4A. A local oscillator frequency higher than the signal frequency (i.e. *f* + *i.f.*) keeps the upper and lower sidebands of the intermediate frequency signals in their original positions. However, when the local oscillator is placed below the signal frequency (*f* − *i.f.*) the i.f. signals transpose the positions of the sidebands. By incorporating two oscillators, one above, the other below the input signal, sideband selection is facilitated.

switched from one to the other set of sidebands without having to adjust the tuning knob. There are various arrangements which permit this to be done, usually by providing two frequency conversion oscillators, one placed on each side of the signal frequency (Fig. 4.4A).

In order to receive without distortion one sideband and carrier of a double sideband a.m. signal the receiver bandpass must conform to certain requirements. For example, if a receiver with a sharply peaked bandpass is tuned to receive one sideband, the selectivity may be sufficient to reject the other sideband but when so tuned the carrier will almost certainly be placed well down the slope of the response curve. In these circumstances the sideband received will be too strong for the carrier; this results in distortion, as though the original signal was greatly overmodulated. If, however, the receiver bandpass approaches that of the ideal flat top shape (say, 2·5 to 3 kc/s wide) with steep sides, the receiver may be tuned so that the carrier is located at the edge of the bandpass or not more than about 5db below maximum response. The carrier and one set of sidebands can then be received without unduly restricting the a.f. response. Since flat-top response curves are difficult to achieve without the aid of effective bandpass crystal or mechanical filters, this form of reception is not always practicable without a fairly complex receiver, though reasonably effective results can often be achieved on receivers having a final i.f. of the order of 85 kc/s.

With an extremely carefully controlled bandpass characteristic it is possible to carry this process a stage further: the receiver can be set to receive either set of sidebands while suppressing both the other set of sidebands and the carrier, thus in effect converting an a.m. signal into an s.s.b. one (**Fig. 4.4B**). It is then necessary to re-insert a carrier as for s.s.b. reception. A great advantage of this form of reception is that it provides for a.m. signals the relative freedom from phase distortion enjoyed by s.s.b. signals. It is facilitated by the use of an asymmetrical bandpass filter with the sharper cut-off located on the carrier side.

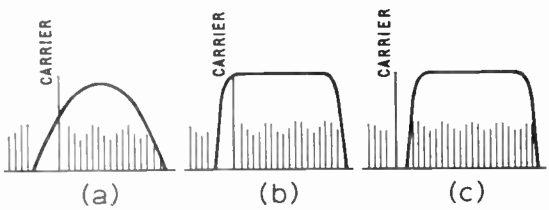


Fig. 4.4B. (a) Response curve which results in the carrier being received at reduced level compared with the sidebands. (b) Selectivity characteristics more suitable for carrier plus one set of sidebands. This permits reception though it results in quadrature distortion. (c) Reception of an a.m. transmission as s.s.b.

Transceivers. In recent years many combined transmitter-receivers, mostly intended primarily for s.s.b. operation, have appeared on the amateur market. When, as is usually the case, some of the circuits function during both transmission and reception, such equipments are termed *transceivers*.

Since some of the most expensive parts of an s.s.b. transmitter, including the high selectivity sideband filter and the high stability oscillator, are equally necessary for high performance receivers, there are clearly both space and economic advantages in an integrated unit. Transceivers have become fairly popular, despite the significant initial cost. The home-construction of multiband transceivers has been successfully achieved by many amateurs, but does represent a complex and advanced project.

The receiver circuits used in transceivers do not differ materially from those of separate receivers, although there is a tendency to use h.f. crystal filters for i.f. selectivity, rather than those at say 455 kc/s in order to reduce the number of stages necessary in the transmitter. However, 455 kc/s mechanical filters (see page 4.17) are often used. Some transceivers are designed so that it is possible to tune the receiver a few kilocycles away from the transmitter frequency without affecting the transmitter tuning; in other cases the receiver and transmitter can only be tuned together.

A related technique, found in some separate receivers, is to provide an outlet from the high stability tunable oscillator to provide a v.f.o. for an s.s.b. transmitter; this is often arranged to tune 5 to 5·5 Mc/s.

Although most transceivers incorporate a 3 kc/s bandwidth sideband filter for s.s.b. and some facilities for c.w. (and often for a.m. as well), it should be noted that few have a narrow band filter, of say 300 c/s bandwidth, desirable in any receiver required for frequent c.w. operation.

Synchronous Receivers

Forms of synchronous or coherent demodulation have recently become of particular interest to communications engineers, and would seem to have useful applications in amateur radio. A coherent system implies the incorporation in the receiver of an oscillator which can be locked in frequency and phase to the incoming signal (this technique is used in domestic colour television receivers and in stereo multiplex decoders). While true communications phase-locked receivers have been developed for certain forms of professional communications they are usually regarded as too complex for normal amateur application. There are, however, relatively simple forms of *homodyne* or *synchrodyne* receivers which have been known for many years but which appear to offer scope for further development for amateur applications.

For example, an effective s.s.b/c.w. homodyne receiver may comprise basically an r.f. amplifier, feeding directly a balanced mixer or product detector with tunable carrier insertion oscillator at the signal frequency, followed directly by the audio stages and without any form of i.f. amplifier. Provided that the balanced mixer is truly linear, then the effective selectivity of the receiver may be determined entirely in the a.f. circuits thus eliminating the need for relatively expensive crystal or mechanical filters of conventional superhet receivers.

Such a basic configuration can be extended to cover a.m. and d.s.b. reception (with substantial theoretical advantages over the use of envelope detection) provided that the local oscillator can be locked in frequency and phase to the incoming signal. One relatively simple way in which this has been shown to be possible in the past is the synchrodyne receiver, in which an incoming a.m. carrier causes the oscillator to be locked on to it.

Such receivers can be considered either as a " superhet " with an i.f. of 0 kc/s or alternatively as straight receivers with balanced linear heterodyne detectors. While a few experimental designs have appeared using these techniques, there would appear to be scope for much further work on these lines.

An interesting possibility is that of developing simple c.w. transceivers since the local oscillator on the signal frequency could be used directly as the transmitter v.f.o. without heterodyning.

PERFORMANCE REQUIREMENTS

Before typical circuitry can be discussed, it is necessary to consider certain of the basic performance requirements in some detail.

Sensitivity

Weak signals need to be amplified much more than strong ones in order to provide a satisfactory output level to headphones or loudspeaker. There are, however, definite limits to this process, and simply adding more amplifying stages will not provide a solution, the limitation being the noise generated both externally and within the receiver (principally in the early stages), which may mask any very weak signals present. This noise, which is described in detail in Chapter 15 (*Noise*), arises partly from sources within the early stages of the receiver (and is thus to a certain extent under the control of the designer) and partly from external sources. The weakest signal which can be satisfactorily used for communication purposes is therefore not governed ultimately by how much overall amplification is available from a receiver but by how weak a signal can be heard above the general noise level. This characteristic of an h.f. receiver is normally defined by stating the minimum signal voltage at the aerial terminals required to produce a specified output with a certain ratio of signal above the noise level at a specified setting of the selectivity control: this is known as the *signal-to-noise* ratio (or more strictly as the *signal-plus-noise-to-noise* ratio).

To give an idea of practical performance, a receiver with a 10db signal-to-noise ratio (at normal selectivity) with an input of 1 to 3 μV is in the high quality class while a receiver which will provide this ratio for a 5 μV signal will not miss many worthwhile signals. In most locations, the weakest c.w. signal readable on a high class receiver will be about 0·5 μV.

A receiver's sensitivity performance could thus be quoted in the following manner:

(a) *telegraphy:* 1 μV c.w. signal across the input sockets to give better than 20db signal-to-noise ratio with a passband of 1·5 kc/s; or

(b) *telephony:* 1 μV signal modulated 30 per cent to give better than 10db signal-to-noise ratio with a passband of 6 kc/s.

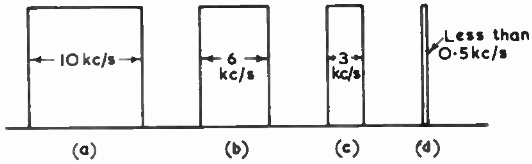


Fig. 4.5. The ideal characteristics of the overall bandpass of a receiver are determined by the type of signals to be received. (a) This would be suitable for normal broadcast reception (a.f. response to 5 kc/s); (b) suitable for a.m. communications (a.f. to 3 kc/s); (c) the bandpass can be halved for single sideband reception without affecting the a.f. response (still 3 kc/s); (d) extremely narrow channels (about 100 c/s) are occupied by manually keyed Morse signals, but some allowance must be made for receiver and transmitter instability—by adjustment of the b.f.o. any desired a.f. beat note can be produced.

For accurate assessment of a receiver's performance, all the factors quoted above must be specified since variation of any one will affect the others. The sensitivity of a receiver is increased by reducing the bandwidth to the limits imposed by the type of transmission; the effect on receiver noise of reducing the i.f. bandwidth is discussed in detail in Chapter 15 to which the reader should refer. Since individual operators possess to differing degrees the ability to copy a weak signal well down in the noise, it would not mean that the quoted signals would be the weakest that it would be possible to copy, but nevertheless form a very useful guide when comparing receiver specifications.

Unfortunately few amateurs have available the test apparatus needed to check manufacturers' figures. A simple but quite effective (and severe) test of sensitivity which can be applied to most well screened receivers is to remove the aerial and replace it by a resistor equal to the receiver's input impedance: if the receiver noise then peaks up when the first r.f. circuit is tuned through resonance (often possible by means of a panel-mounted aerial trimmer) on the highest frequency bands it is reasonably certain that the receiver possesses high sensitivity.

The sensitivity of a receiver is often defined, independently of the i.f. bandwidth, in terms of *noise factor*, usually quoted in decibels (db)—see Chapter 15 (*Noise*). Usable sensitivity on the h.f. band is limited to roughly 6-7db at the higher frequencies, and lesser sensitivity (higher noise factor) at lower frequencies. An h.f. receiver with a noise factor of 10db or less is likely to prove fully acceptable for amateur operation. Older models often have a noise figure of 12-16db or above due to the use of " noisy " valves in the first stages. Because of the very much lower external noise, far greater sensitivity can be used on v.h.f.

Selectivity

The ability of a receiver to separate stations on closely adjacent frequencies is determined by its *selectivity*. Chapter 1 (*Principles*) explains why it is generally necessary to use a superhet receiver with a low intermediate frequency to obtain a very high degree of selectivity—though today i.f. crystal filters are available up to about 10·7 Mc/s with excellent selectivity, they are not however cheap to buy.

The ultimate limits to practical selectivity are governed by the *bandwidth* of the type of transmission which is to be received. For high fidelity broadcast reception the response of a receiver needs to extend to some 15 kc/s on either side of the carrier frequency, equivalent to a bandwidth of 30 kc/s; for average broadcast reception this figure can be reduced to about 10 kc/s. For communications-quality telephony in the amateur bands 6 kc/s for the reception of both sidebands or 3 kc/s for the reception of one sideband will suffice; for telegraphy at manual keying speeds the total bandwidth could be reduced to less than 100 c/s, although to accommodate transmitter and/or receiver instability a 300 c/s bandwidth is generally regarded as more practical. These ideal receiver characteristics are indicated in **Fig. 4.5.**

To compare the selectivity of different receivers or the same receiver in different positions of its selectivity control, curves of the type shown in **Fig. 4.6** can be used. There are two ways in which these curves must be considered. The first is the *nose* figure which represents the bandwidth in kilocycles over which a signal will be heard with relatively little loss of strength; the other figure—regarded by most amateurs as the more important—is the bandwidth over which a really powerful signal is still audible (often taken as a reduction of 1000 times on the strength of the signal when correctly tuned in) and this is often termed the *skirt* performance. These two sets of figures are related by what is termed the *shape factor* of the receiver, found by dividing the bandwidth at the skirt by that at the nose.

The ideal receiver would have a shape factor of one on each position of selectivity; such a receiver would in practice be impossible to design at the present state of the art. Even the very best modern receiver is unlikely to achieve a shape factor much less than 1·5, and most good receivers range from about 2·5 to 5·5. A receiver which has a skirt bandwidth at −60db, of less than 10 kc/s can be considered to have quite good selectivity characteristics.

Spurious Responses

One of the major defects of simple superhet receivers is

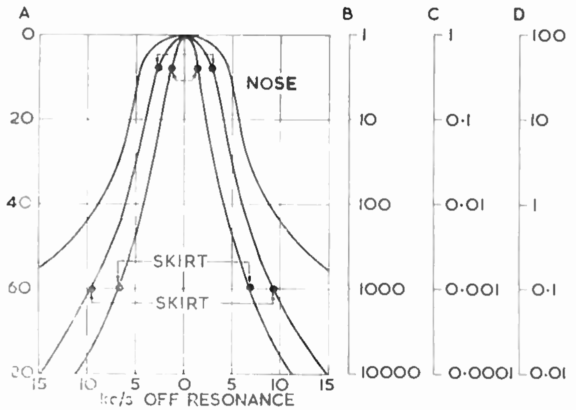


Fig. 4.6. The ideal vertical sides of Fig. 4.5 cannot be achieved in practice. The curves shown here are typical. These three curves represent the overall selectivity of receivers varying from the "just adequate" broadcast curve of a superhet receiver having about four tuned i.f. circuits on 470 kc/s to those of a moderately good communications receiver. *A, B, C* and *D* indicate four different scales often used to indicate similar results: *A* is a scale based on the attenuation in decibels from maximum response; *B* represents the relative signal inputs for a constant output; *C* is the output voltage compared with that at maximum response; *D* is the response expressed as a percentage.

that the same station can usually be received at more than one position of the tuning dial, or alternatively that the various oscillators within the set provide signals which can be tuned in at various points on the dial as though they were external carriers. In practice, this means that strong commercial stations are often heard as though they were operating within the amateur bands, while searching for weak stations is made more difficult by tuning through the carriers generated within the receiver, often referred to as *birdies*.

The most important causes of spurious responses are:

(a) The reception of signals on the image frequency of the receiver.

(b) The reception of harmonics of the various oscillators in a receiver.

These are considered in more detail later in the chapter.

All superhet receivers suffer from image response (see Chapter 1) which becomes progressively more important as the signal frequency increases. In practice the effect may be reduced to a very low figure by raising the intermediate frequency. In general terms, the image response of a given receiver compared with its response on its correct frequency needs to be reduced by at least 32 times (30db) at the highest frequency to which the set tunes (usually about 30 Mc/s). A full specification should give the receiver's response to image signals at various frequencies, the response at the highest frequency to which the set tunes is usually the most important.

In a single conversion superhet, birdies can be caused by harmonics of the beat frequency oscillator, but in practice this seldom happens. On the other hand, with double- and triple-conversion receivers (and with single-conversion receivers operated in conjunction with a converter making, in effect, double conversion), the elimination, or at least reduction of birdies becomes a major design problem. It is seldom possible in a wide coverage multi-conversion receiver to avoid birdies altogether and the problem then resolves itself to ensuring that they do not fall in frequently used portions of the amateur bands. The ideal specifications would be in the form "internally generated spurious responses are below the noise level in all cases."

Spurious responses are discussed in more detail on page 4.38.

Stability

The ability of a receiver to remain tuned to a particular frequency depends upon both electrical and mechanical stability of the tuned circuits, more especially those in the local oscillator(s). The primary cause of electrical instability in an oscillator (see Chapter 6) is the effect of heat on the tuned circuit and valve electrodes. Even in the best designs, there will usually be a steady variation appearing as drift in the first 10-15 minutes after switching on a cold set, but it should then settle down with little further change.

After undergoing a number of heat cycles, some components do not return precisely to their original values. This makes it very difficult to maintain accurate calibration of a bandspread dial over a long period. Many receivers therefore include a crystal controlled oscillator of high stability to provide 100 kc/s or 1000 kc/s marker signals on which the calibration can be regularly checked and adjusted. A calibration marker of this type is also extremely useful where the set has both a coarse (bandset) tuning knob and a fine (bandspread) tuning knob as it enables the main dial to be re-set accurately when changing bands.

Mechanical instability, which usually appears as a definite shift in frequency when the receiver is subjected to any form of mechanical shock or vibration, cannot easily be defined in the form of a performance specification. Sturdy construction, a heavy chassis and suitable mounting of components are required to reduce this form of instability. Receiver drift can be specified in terms of maximum drift in cycles per second over given periods of time, usually quoting a separate figure to cover the warming up period. For good s.s.b. reception short term stabilities of the order of a few cycles are desirable.

Tuning Rate

To tune accurately and rapidly to a suppressed carrier type of transmission (see Chapter 10), it is desirable to be able to set the receiver to within about 25 c/s or so of the station's frequency. Such accuracy of tuning, which is also most beneficial for c.w. reception, is unlikely to be achieved unless each complete revolution of the tuning knob represents only a moderate shift in the frequency of the receiver. A typical figure for a modern high grade receiver would be 5 kc/s or so per revolution, but equally important would be a tuning mechanism with a smooth action, free from backlash. The trend to very low tuning rates is a feature of the most modern designs. The use of step down gears to reduce the tuning rate of the control knob is termed *mechanical bandspreading*. The same effect can be achieved electrically by placing across the main tuning capacitors (which may vary in capacitance swing from 150 to about 500 pF) much smaller value variable capacitors of the order of 20 pF swing: see Fig. 4.7. Bandspreading is also possible by the use of voltage variable capacitance diodes. Permeability tuning makes possible very good bandspreading. In practice a combination of mechanical and *electrical bandspreading* is often used, while there is an increasing tendency to restrict the frequency coverage of receivers to the amateur bands only. Except for receivers having a tunable i.f. the tuning rate will usually vary on each waveband, and sometimes also within each band, although this problem can be overcome by the use of rather complex frequency synthesis or interpolation oscillator techniques.

Cross-modulation and Blocking

After a receiver has been made highly selective, preferably down to the −60db level, there may remain the problem of coping with extremely strong local or distant signals. Even when such a station is transmitting on a frequency well outside the i.f. pass band of the receiver, it may affect reception by *cross modulation* or *blocking*.

When a very strong signal reaches the input of a receiver whose gain control is set for the reception of weak signals, the relatively poor selectivity of the early signal frequency and intermediate frequency circuits (prior to selective filters) will not be sufficient to prevent the strong signal from being amplified, so that one of the stages may be driven into non-linearity. This results in the required weak signal being either reduced in strength (blocking) or becoming modulated by the strong signal (cross modulation).

Clearly, the more amplifying stages there are in a receiver before the circuits which determine its selectivity characteristics, the greater are the chances that one of them will be overloaded. From this it follows that in a double (or triple) conversion receiver—where there will be at least one r.f. amplifier and two mixer stages before the selective circuits—particular care is necessary to avoid overloading.

It is difficult for the amateur to assess in terms of actual figures the performance of receivers in this respect. Undoubtedly in many sets a really powerful signal can affect reception over a considerable portion of a band. Even on a high grade receiver costing several hundred pounds, it has been shown that cross-modulation effects can occur 50 kc/s away from an S9 + 60db signal and 15-20 kc/s away from an S9 + 40db one. At the lower levels, the stage most likely to produce these effects is the second mixer, but for really strong signals (S9 + 50db or more) the first mixer can be the prime source of this interference. Many double-conversion receivers will have a cross-modulation performance much inferior to the example quoted.

The susceptibility of a particular design to this form of interference depends on a number of factors: notably the pattern of gain distribution through the receiver and how this is modified by the action of a.g.c. or the manual gain control (some sets have a dual track potentiometer to provide differing gain tapers on r.f. and i.f. valves), as well as the types of valve used. One method of reducing the amount of r.f. gain needed to override the noise of the first mixer is to use a triode mixer, calling for a fairly high first i.f., and requiring a greater proportion of the receiver gain to be provided by the i.f. stages. Another technique is the use of beam deflection valves (e.g. 7360) as low-noise mixers. The fitting of an adjustable attenuator between the aerial and the first stage is also a useful measure.

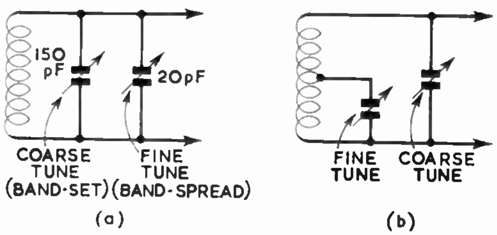


Fig. 4.7. Two common methods of providing bandspread tuning. (a) Small value capacitor connected in parallel across the main tuning capacitor. (b) Tapped coil permits different degrees of bandspreading to suit the various amateur bands, but can result in unwanted resonances.

FRONT-END CIRCUITS

R.F. Amplifier

In a superhet receiver, amplification at the signal frequency is highly desirable in order to overcome the noise contributed by the mixer. Also important is the protection provided by the additional tuned circuit against second channel (image) and certain other forms of spurious response. Excessive r.f. amplification, however, increases the likelihood of the receiver suffering from blocking or cross modulation.

It is possible, with low-noise mixers, to achieve maximum usable sensitivity without an r.f. stage, but it should not be forgotten that the maximum pre-mixer selectivity is a valuable aid in reducing spurious responses, and such selectivity is most easily achieved with one or more r.f. stages. Gain, however, should be limited to that necessary to override mixer noise.

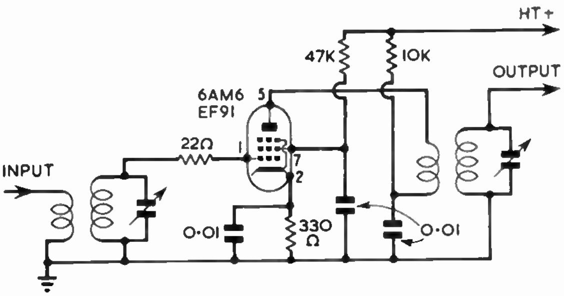


Fig. 4.8. Typical tuned r.f. amplifier.

The design of the r.f. amplifier stage(s) is governed by the type of mixer and the intermediate frequency of the receiver. With a conventional hexode or heptode mixer, the gain required from the r.f. amplifier will be of the order of 20db at the highest frequency (say 30 Mc/s) to ensure that the optimum signal-to-noise ratio is achieved. This can be obtained from a single good pentode amplifier. If the receiver uses a triode mixer, which has a much lower noise contribution, the gain of the r.f. amplifier becomes less important, but the stage is still useful in reducing spurious responses.

For a single-conversion superhet having an i.f. of the order of 470 kc/s (or for a double-conversion model in which the higher i.f. is of this order), two tuned r.f. amplifiers are needed to reduce image response to negligible proportions on frequencies above about 10 Mc/s, though as noted already, one stage is sufficient to secure optimum signal-to-noise ratio. If the intermediate frequency is increased to about 1·5 Mc/s, then a single r.f. stage should be able to provide good image rejection up to 30 Mc/s.

In practice, the r.f. amplifier usually consists of one or more high gain r.f. pentodes (valve types such as EF93/6BA6, EF85/6BY7, EF183/6EH7, EF95/6AK5, 6BZ6, 6SG7, 6CB6, EF91/6AM6 and 6DC6 are suitable) with a tuned input circuit between grid and cathode and with the valve operating into an inductive load which may be either the primary winding of an interstage r.f. transformer or a tuned circuit. A typical circuit is shown in **Fig. 4.8.** However, this

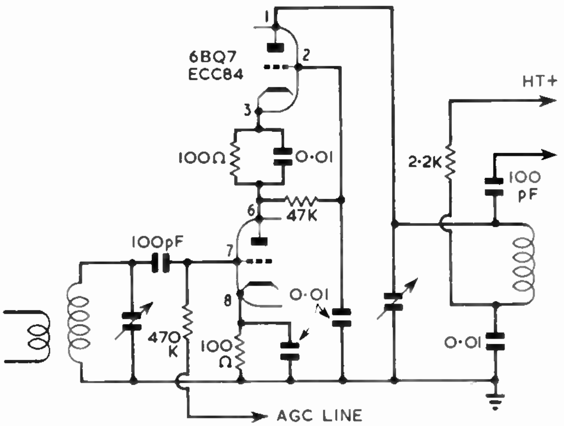


Fig. 4.9. Basic series-cascode r.f. amplifier. Although formerly used mainly for v.h.f. receivers, this arrangement is now found in h.f. receivers.

is not the only possible arrangement: of increasing popularity is the series-cascode circuit, while the grounded-grid amplifier is suitable for use with a triode mixer or other applications where high gain is not required.

In the matter of optimum signal-to-noise ratio and overall stage gain, there is little to choose—for frequencies below 30 Mc/s—between a correctly designed r.f. pentode and the double-triode cascode. However, the cascode is sometimes less susceptible to cross-modulation and some constructors consider it less critical in adjustment and operation. A typical series-cascode circuit is shown in **Fig. 4.9.** Suitable valves include the ECC189/6ES8, ECC84/6CW7 and 6BQ7.

The correct matching of the aerial to the valve input impedance is of great importance in securing optimum signal-to-noise ratio: the voltage gain obtained from the step-up ratio of the input circuit is not degraded by valve noise. General coverage receivers are often designed with a medium impedance input (300-600 ohms) with loose aerial coupling to the grid circuit, the aerial coil having a fair number of turns. For amateur-bands-only receivers it is usual to reduce the input impedance to 75 ohms, with the few turns of the aerial input coil overwound on the earthy end of the grid coil to provide very tight coupling. This system is to be recommended as it enables the maximum voltage step-up to be obtained. With tight coupling it is advisable to include a panel controlled aerial trimmer across the tuned grid circuit. When low impedance input circuits are to be used in conjunction with end-fed and similar aerials having a range of impedances, it is highly desirable to fit an external aerial matching unit: often that used for the transmitter can be employed by means of a coaxial change-over relay.

The first tuned circuit in a receiver is of especial importance in determining the sensitivity of the receiver and it should be noted that the alignment of this circuit to that of the mixer presents particular problems owing to the effects produced when coupling different types of aerial to the receiver. If the aerial is reactive its connection will usually cause detuning of the input circuit. A satisfactory method of overcoming this difficulty is to fit a panel control aerial trimmer to allow the circuit to be readily brought into accurate adjustment by the user when changing bands or aerials. If a receiver is designed solely for use with a non-reactive feeder of known impedance, this control would be unnecessary.

A short-base r.f. pentode will usually have a lower noise contribution than a variable-mu pentode, but in practice this is likely to be noticeable only above 21 Mc/s; a variable-mu type is often used for convenience of a.g.c. action, and the reduction of cross modulation. Of the present range of pentode valves, the 6CB6 and EF91 (6AM6) provide low noise but may prove susceptible to cross-modulation effects; the EF85/6BY7 or the American 6DC6 is better in this respect. Where a high slope, short-base r.f. pentode is used in an a.g.c.-controlled stage, the screen feed resistor should be of fairly high value (33 K to 47 K ohms) to lengthen the grid base, and so continue to give linear though reduced amplification as the negative bias is increased.

Controlled regenerative feedback in an r.f. amplifier can be of material assistance in increasing stage gain and in giving greater protection against image response. As explained in Chapter 15, this will be at some cost in signal-to-noise ratio. Thus although regeneration is sometimes useful for smaller receivers, it is to be avoided in high-performance designs. Optimum signal-to-noise ratios cannot be obtained from an

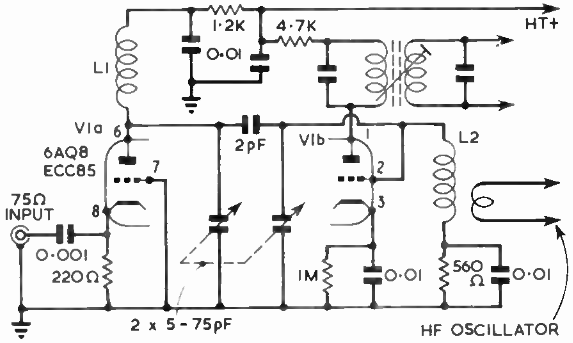


Fig. 4.10. Low noise r.f. amplifier-mixer using an ECC85/6AQ8 double triode valve. The first triode section functions as a grounded grid amplifier with untuned input; the second triode is the mixer inductively coupled to the h.f. oscillator. Note IM and 560 Ω resistors should be transposed.

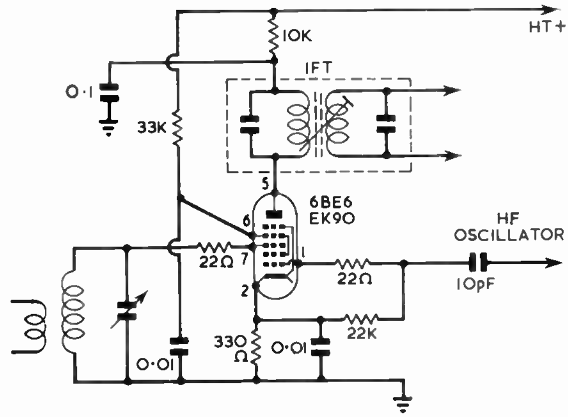


Fig. 4.11. Although heptode frequency changer valves such as the 6BE6/EK90 can be used as a combined mixer-oscillator, their use as a mixer with a separate h.f. oscillator is preferred for high performance receivers.

r.f. amplifier unless all forms of positive feedback due to stray capacitances, inefficient screen decoupling and the effect of common cathode lead inductance are reduced to the lowest possible figure. Coupling between input and output circuits must be kept extremely low; one form of coupling which is sometimes overlooked is that due to the common rotor spindle of a ganged tuning capacitor being slightly above earth potential. Some r.f. valves have two cathode pins (one for input connections, the other for output connections) to overcome the inductive coupling of a common cathode lead.

Where a low noise mixer is used, the r.f. gain can be appreciably reduced without loss of signal-to-noise ratio and with benefit to cross-modulation characteristics; with a triode mixer a single grounded-grid triode r.f. amplifier may suffice, particularly on 14 Mc/s and above. When fed from a low-impedance aerial transmission line, the valve input circuit need not be tuned. An arrangement of this type is shown in **Fig. 4.10.**

TABLE 4.1

Equivalent Noise Resistances (e.n.r.) of Typical Valves

R.F. Amplifiers			
6AC7	720	EF42	750
6AC7 (triode connected)	220	EF50	1400
6AG5	1900	EF54	700
6AK5	1880	EF85/6BY7	1500
6AK5 (triode connected)	385	EF91/6AM6	1200
6BA6	3520	EF183/6EH7	490
6BQ7A	390	EF184/6EJ7	300
6BZ6	1460		
6CB6	1440	**Mixers**	
6F23	670	6AK5	7520
6F24	370	6BA6	14080
6J6	470	6BA7	60000
6K7	16400	6BE6	190000
6SG7	3300	6J6	1880
6SH7	2850	6K8/ECH35	290000
6SJ7	5840	6L7	255000
6SK7	10500	6SA7	240000
6U8/EFC82 (pentode)	2280	6SB7Y	62000
6U8/ECF82 (triode)	295	6U8/ECF82 (pentode)	*9300
12AT7/ECC81	380	6U8/ECF82 (triode)	2000
12AU7/ECC82	1140	12AT7/ECC81	2400
12AX7/ECC83	1560	12AU7/ECC82	7280
ECC84/6CW7	*420	ECF86/6HG8	*2700
ECC85/6AQ8	500	ECH81/6AJ8	66000

* Calculated.

The above values are normally optimum manufacturers' figures and the e.n.r. will rise sharply when the mutual conductance is lowered by increasing the bias or due to valve ageing.

Mixer

In Chapter 1 it has been shown that any non-linear circuit element will act as a mixer, that is to say if frequencies f_1, f_2 are combined in the element, frequencies $f_1, f_2, f_1 + f_2$ and $f_1 - f_2$ will be present in the output. Thus almost any valve, crystal diode or transistor can function as a frequency converter. In h.f. receiver practice, however, the multi-grid valves, specially designed for the purpose (EK90/6BE6, ECF82/6U8, ECH81/6AJ8 or the older and much noisier 6SA7 and 6K8), are generally used because of the appreciable conversion gain which they provide and their relative freedom from interaction between the signal tuned circuits and the oscillator section. Occasionally the triode mixer is used because of its low noise contribution, but tends to be a little more critical in operation and provides much lower output; it is subject to oscillator " pulling " unless the intermediate frequency is high, but can be usefully employed in double conversion receivers with a high first i.f. as a means of reducing cross-modulation effects.

It is sometimes difficult with conventional mixer circuits to

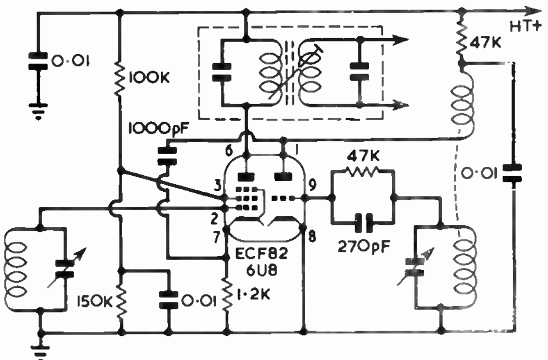


Fig. 4.12. One multiple valve that can be effectively used as a combined mixer-oscillator is the ECF82/6U8 triode-pentode in which two sections are isolated by means of internal screening.

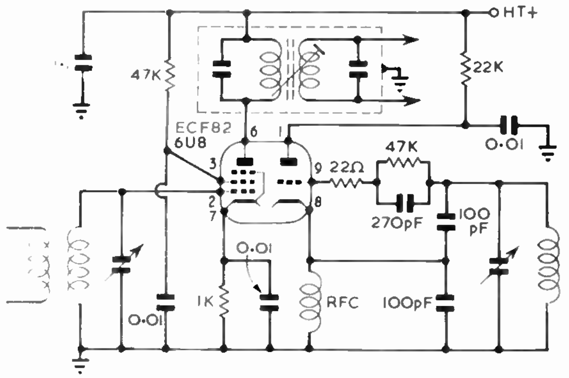


Fig. 4.13. A Colpitts oscillator for the ECF82/6U8 frequency changer. This requires only a two-terminal coil and is most suitable for use with independent oscillator tuning, as tracking problems do not then arise.

maintain constant conversion gain over the full tuning range of a receiver. This is partly because of the varying effect of inter-electrode and stray wiring capacitance and partly because of the difficulty of providing a constant oscillator injection voltage as the frequency is varied. A higher and more constant oscillator voltage can generally be obtained by using a separate valve for the oscillator in place of the oscillator section of the normal frequency changer valve. An exception is the ECF82/6U8 triode pentode which can provide efficient conversion in a single valve envelope, particularly at the higher frequencies. A suitable pentode mixer for use with a separate oscillator is the EF94/6AU6.

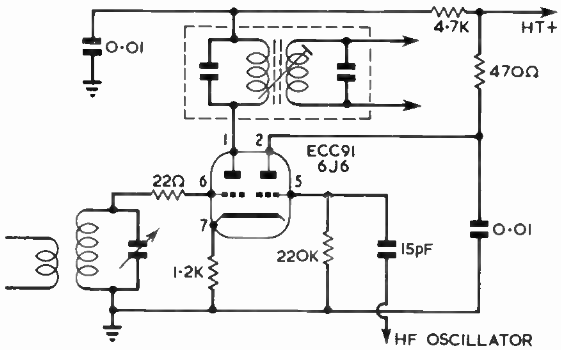


Fig. 4.14. A double triode valve used as a low noise mixer with the second section functioning as a cathode follower to provide isolation between the mixer and h.f. oscillator.

To improve oscillator stability both in combined mixer/oscillator valves and where a separate oscillator valve is used, the d.c. potentials applied to the grid and screen grid of the mixer should remain reasonably constant. It is inadvisable in a high performance receiver to apply a.g.c. to the mixer stage. Also for this reason, and because the screen voltage of a mixer is fairly critical for optimum conversion, the screen supply for the mixer should be derived either from a relatively low impedance source such as a potential divider network not containing high value resistors or from a regulated supply with no high value series feed resistors.

The cathode, screen and anode by-pass capacitors of the mixer have all to be effective at signal, oscillator and intermediate frequencies: a ceramic type of 0·01 μF capacitance will usually prove suitable. Typical mixer circuits are shown in **Figs. 4.10-4.14.**

A rather different twin-triode mixer arrangement that has been used in several recent designs is shown in **Fig. 4.15.** The first triode section forms a cathode follower, feeding the signal into the cathode of the mixer section, with the oscillator signal fed to the grid. The mutual conductance of the cathode-follower section must be higher than that of the mixer section; this can be achieved with identical twin triodes by reducing the voltage on the anode of the mixer section by feeding it through a resistor of the order of 33 K ohms or by a potentiometer network as shown.

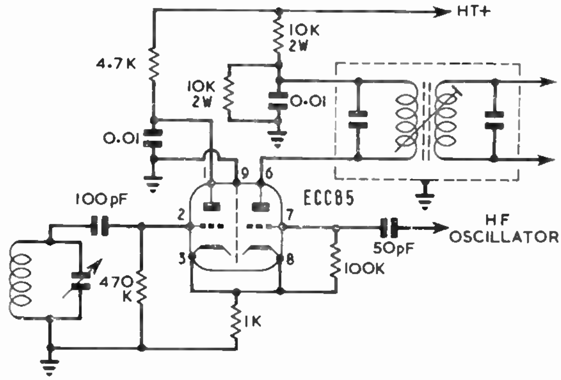


Fig. 4.15. Twin triode mixer in which the mutual conductance of the second section is reduced by a low h.t. supply. A relatively low injection voltage (1-2 volts) is required. The common cathode resistor may be optimized between 100 and 1000 ohms.

To assist in the reduction of cross-modulation and some forms of spurious response, the use of a balanced form of mixer has much in its favour. This may consist of twin triodes (e.g. B329, 12AU7, ECC82), two transistors, two or preferably four diodes (for example in the ring modulator circuit of **Fig. 4.16**), or a single beam-deflection valve (7360). All these techniques can be found in recent receivers for amateur or commercial communications. The low-noise balanced beam-deflection mixer (equivalent noise resistance (e.n.r.) about 1500 ohms) is shown in **Fig. 4.17 (a)**; an unbalanced version has also been used.

The H.F. Oscillator

The frequency to which a superhet receiver responds is governed not by the signal frequency circuits but by the

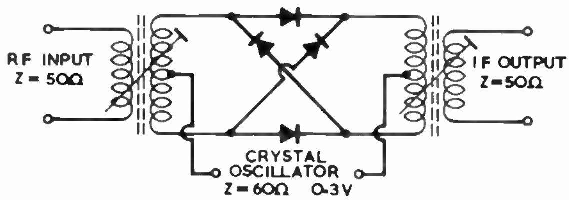


Fig. 4.16. Balanced ring modulator using four crystal diodes. This type of mixer is less susceptible to cross-modulation and to various forms of spurious responses. The particular circuit shown is a wideband v.h.f. first mixer but it can be adapted to h.f. operation. Provided an r.f. stage with a gain of about 20db precedes it, this form of mixer can be used in high performance valve or transistor receivers.

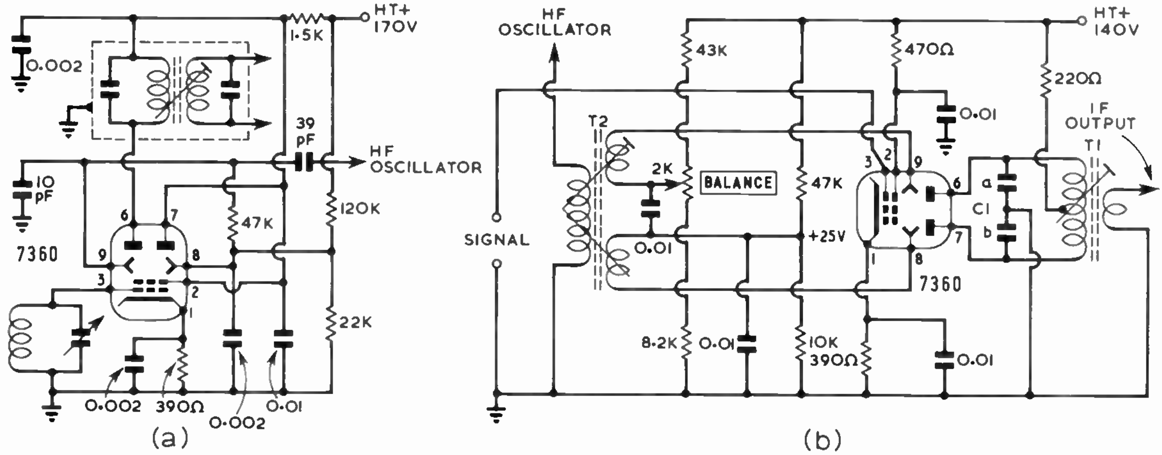


Fig. 4.17(a) Unbalanced version of the 7360 beam deflection mixer. (b) A balanced form of the 7360 beam deflection mixer designed for very low cross modulation characteristics. Oscillator voltage at the deflection plates is 1-10 volts r.m.s. T1 and T2 and the general layout should be arranged to maintain balance.

setting of the local oscillator. Any frequency variations of the oscillator are reflected in apparent variation of the received signal. The overall stability of a single-conversion receiver is determined largely by the design of the h.f. (local) oscillator: that of a double-conversion set by that of both local oscillators, but as oscillator stability is more difficult to obtain with increasing frequency, the first local oscillator (which normally functions on a higher frequency than the second local oscillator) will usually be the determining factor.

The prime requirements therefore of an h.f. oscillator are freedom from frequency changes resulting from mechanical vibration or temperature changes, sufficient but not excessive fundamental output for maximum conversion efficiency, low harmonic output to minimize spurious responses (particularly important in double- and triple-conversion receivers), and no undue variation of output voltage throughout the tuning range. For optimum and constant conversion gain, the oscillator injection voltage to the mixer is fairly critical (usually in the region of 9-10 volts).

Fig. 1.64 in Chapter 1 shows a typical combined mixer/oscillator frequency changer using a triode hexode or triode heptode valve. Modern multiple valves designed for frequency conversion are capable of good performance, but a separate h.f. oscillator is still preferred for high performance receivers. Apart from the advantages already given, this permits the use of a valve having high mutual conductance so that the amount of positive feed-back required for oscillation is less, resulting in a more constant output voltage and better stability. The use of a separate valve also allows the injection coupling to the mixer to be more accurately controlled and reduces interaction effects between signal frequency and oscillator tuned circuits. With double-conversion receivers, the second local oscillator, when operating on a fixed frequency, is often combined with the mixer.

High-slope r.f. pentodes, sometimes triode connected, are commonly used (e.g., EF94/6AU6). To reduce spurious responses the harmonic content of the oscillator output should be kept as low as possible, and for this reason a high-slope power triode valve, such as the EC90/6C4, is often used.

Basically, any of the standard forms of oscillator are suitable provided that the harmonic output is low. The most common arrangement, shown in **Fig. 4.18**, is the tuned grid circuit with anode positive feedback winding. The Colpitts oscillator shown in **Fig. 4.13** is also popular as the coil has only a single winding.

The stability requirements of an h.f. oscillator are akin to those for a variable frequency oscillator (see Chapter 6) but with the added disadvantage that the oscillator operates on its fundamental frequency to above 30 Mc/s. With the growing interest in single sideband reception, it has become even more important to ensure a very high order of oscillator stability. One result is a growing tendency to use a crystal-controlled h.f. oscillator, a typical circuit is shown in **Fig. 4.19**. Where the tunable h.f. oscillator is retained, many of the techniques of the v.f.o. have been introduced; for example the inclusion of an isolating stage between the oscillator and the mixer; see **Fig. 4.20**.

Most oscillators vary in frequency with a change in applied voltage. Since some variation in h.t. is likely when the gain of a receiver is changed, it is highly desirable to stabilize the oscillator h.t. supply by means of a voltage regulator tube. Some factory-built receivers also stabilize the heater current by means of a barretter.

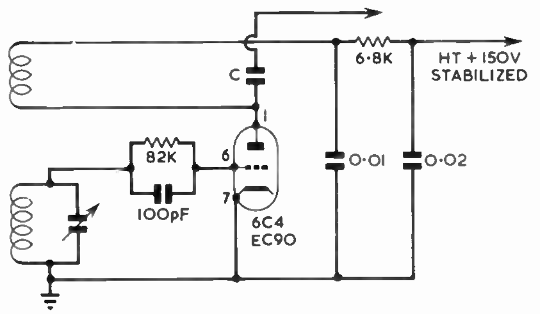


Fig. 4.18. Typical h.f. oscillator using a triode valve. Alternatively a pentode, such as the 6AU6 or 6BA6, may be used. The value of the coupling capacitor C will depend upon the type of injection used.

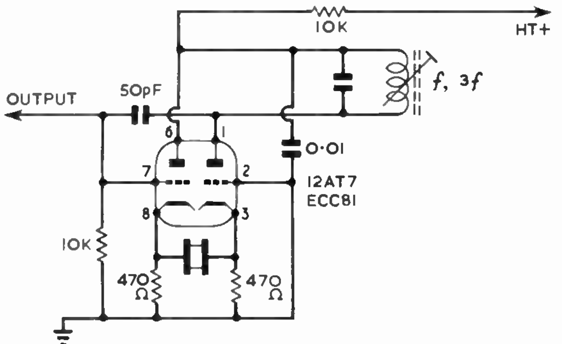


Fig. 4.19. Crystal oscillator circuit suitable for providing output on the series resonant fundamental frequency of the crystal or on the third overtone frequency (approximately three times the fundamental frequency) depending on the frequency of the tuned circuit.

Oscillator drift is caused primarily by the effects of rising temperature on the valve capacitances, the coils and the tuning capacitor. Drift can thus be minimized by such measures as:

(a) Adequate ventilation to keep operating temperatures low.

(b) Swamping of small capacitance changes by a large fixed capacitance across the tuning capacitor (part of this swamping capacitance can usefully take the form of a capacitor having a high negative-temperature co-efficient).

(c) Placing of heat sensitive components as far as possible from the main sources of heat.

(d) Careful choice of tuning capacitor and of the coil formers and winding technique.

(e) Rigid, vibration-resistant construction.

Considerable improvement can be obtained by using components which react in opposite directions to heat, thus providing a degree of automatic compensation.

The vanes of the oscillator tuning capacitor must be rigid enough to eliminate the vibration effects (*microphony*) which can otherwise occur when this component is near the loudspeaker. Appreciable r.f. current flows through the oscillator tuning capacitor and a poor rotor contact of varying resistance will cause noise and *twitter* (slight discontinuities and jumps in tuning). The type of wiper contacts

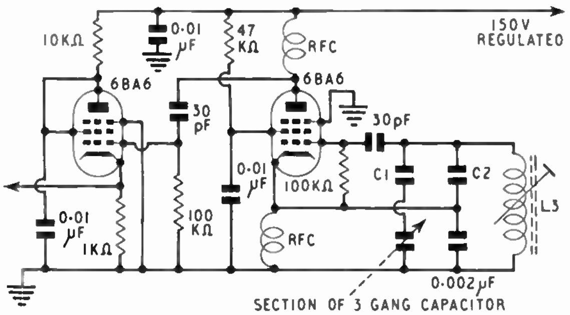


Fig. 4.20. A cathode follower isolating stage connected between a variable frequency oscillator and the mixer (as used in the receiver described on page 4.62).

generally fitted on gang capacitors can give trouble, and some designers prefer to use a separate variable capacitor mechanically linked to the gang capacitor used for the mixer and r.f. tuned circuits. For maximum stability, a really well-made capacitor, with stout, well-spaced vanes, and double bearings should be selected. Alternatively, the variable tuning capacitor can be eliminated by adopting permeability tuning in which a core is moved in and out of the coil (see later).

Hum in the oscillator stage results in the appearance of hum modulation, noticeable only when a station is tuned in. To reduce hum in the oscillator stage, it is recommended that the cathode should be connected directly to the earth line and a bypass capacitor should always be connected directly across the heater pins on the valve socket.

A circuit which is finding increasing favour both as a tunable h.f. oscillator and as a fixed or variable second conversion oscillator, is the cathode-coupled oscillator using a double-triode valve: see **Fig. 4.21.** This circuit offers the advantage of a two-terminal coil, is capable of good stability and has low harmonic output. The circuit is also most tolerant

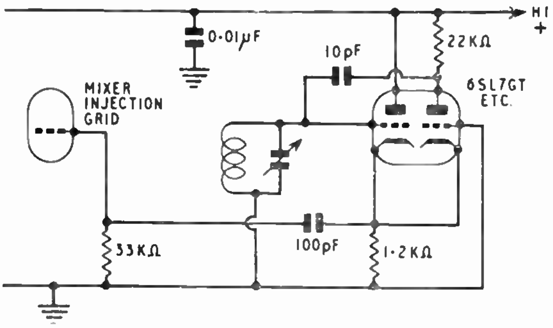


Fig. 4.21. The so-called cathode-coupled oscillator is suitable for either first or second conversion oscillators.

of high C operation, a useful feature where a stable or wide-range oscillator is required. Under certain conditions the circuit may "squegg," particularly if the grid coupling capacitor is increased much beyond the suggested 10 pF. This can be overcome by fitting a small non-inductive (carbon) resistor of between 10 and 50 ohms immediately at the grid pin of the first triode section. If a multi-band oscillator with switched grid circuit is required, it is advisable to set up the circuit first with a 3-30 pF trimmer in place of the fixed 10 pF capacitor and to experiment with different settings and/or different series grid resistors on the highest and lowest frequencies required, replacing the trimmer with a fixed capacitor when a suitable value has been found—it is unlikely that this will prove critical. The h.t. bypass capacitor should be connected directly between the anode of the first triode section and the earthing point of the oscillator to ensure low harmonic output. For harmonic suppression, the series inductance of the bypass capacitor is more important than its actual value and it may prove worth trying several types to achieve optimum harmonic suppression, or to use a combination of different types.

The rigid stability requirements of s.s.b. and some forms of telegraphy have led commercial designers towards the use of frequency synthesis techniques for the h.f. oscillator of receivers. Frequency synthesis is the name given to the formation by means of mixing (heterodyning), frequency division and multiplication of a required frequency from a single crystal oscillator or from a relatively small bank of crystals. The full synthesis of frequencies at intervals of 1 kc/s or less throughout the h.f. range with the necessary purity of output is a complex and costly process. However some factory receivers (HRO500, PR155) have used simplified forms.

In the HRO500 a 1 Mc/s oscillator with associated spectrum generator is used to provide 1 Mc/s markers to which a free running oscillator is stabilized by means of a phase-lock loop. In the PR155, an interpolation oscillator with a 1·2 Mc/s tuning span is also incorporated so that the phase-locked oscillator provides output throughout the required range. Another related, though not strictly synthesis, technique used in the RA17, RA217 receivers is based on the Wadley triple-mix system to eliminate automatically the effects of any drift of the local oscillator so that the overall stability is that of a stable crystal-controlled reference oscillator.

Home-construction of receivers using any of these techniques, though feasible, must be regarded as a project suitable only for those with very considerable experience. Factory receivers of this class inevitably cost several hundreds of pounds.

I.F. AMPLIFICATION

The i.f. amplifier is the heart of a superhet receiver, for it is here that the main voltage gain occurs (usually of the order of 80db) and the overall selectivity response is shaped.

Fig. 4.22 shows a typical pentode i.f. amplifier stage, very similar basically to those found in normal broadcast receivers. I.F. transformers comprising pairs of mutually-coupled tuned circuits are used to couple the valve to the preceding and succeeding stages. To permit the gain of the stage to be controlled by an automatic gain control system, the valve usually has variable-mu characteristics. Typical types are EF85/6BY7, EF93/6BA6, 6BZ6, EF183/6EH7, or the older octal-based types 6K7, 6SG7, 6SK7 and EF39.

The gain obtained in an i.f. stage depends upon the mutual conductance (g_m) of the valve, the Q and the L/C ratio of the resonant circuits in the i.f. transformers and the coupling (k) between the transformer windings. For a specified frequency, the gain will be maximum with the highest possible L/C ratio but if this is increased too much, it becomes increasingly difficult to maintain good stability and long term accuracy of alignment; this may also be affected by variations of the input resistance of the following stage with a.g.c. action.

Gain will be maximum when the product kQ is equal to one. I.F. transformers designed for this condition are said to be critically coupled. When the coupling is increased beyond this point (over-coupled) the maximum gain occurs at two frequencies equally spaced about the resonant frequency and there is a slight reduction in gain at exact resonance: this condition may be used in broadcast receivers to increase the bandwidth for good quality reception. If the coupling factor is lowered (under-coupled), the stage gain falls but the response curve is sharpened (see Chapter 1 page 1.20) and this may be useful in communication receivers. In the past some designs used mechanical methods of varying the coupling by actually moving the coils, but this is difficult to arrange. Alternative and more practical methods of varying selectivity are discussed later in this section.

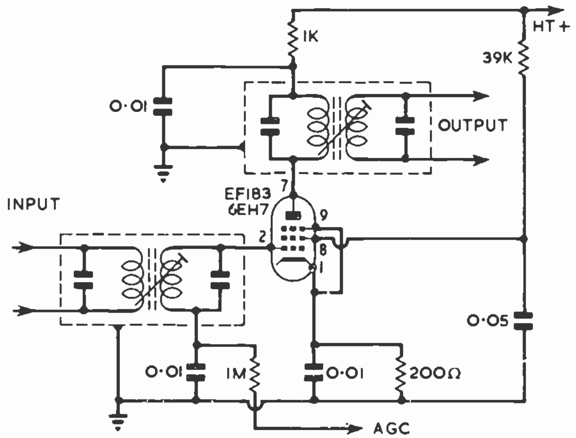


Fig. 4.22. Modern i.f. amplifier using high slope frame-grid variable-mu pentode.

Choice of I.F.

Choice of intermediate frequency is a most important consideration for the designer. The lower the frequency the easier it will be to obtain high gain and good selectivity. Conversely, the higher the frequency, the greater will be the difference between the frequencies of the signal and the h.f. oscillator, from which it follows that the greater will be the protection afforded against image (second channel) interference and the " pulling " of the h.f. oscillator by strong signals. Since these two basic considerations are directly opposed, the i.f. of a single-conversion superhet must therefore always be a matter of compromise. In practice, it will be governed by such considerations as the number of tuned signal frequency circuits before the mixer, the degree of selectivity required, the number of i.f. stages, the absence of strong signals on or about the i.f., and the availability of suitable i.f. transformers. By international usage, frequencies for single-conversion receivers are usually within the band 455-470 kc/s or around 1·6 Mc/s.

In a receiver having no (or only one) tuned r.f. stage, an i.f. of 1·6 Mc/s or above is required to reduce image response to a satisfactory figure at 30 Mc/s. With an i.f. of about 460 kc/s, three tuned circuits (two r.f. stages) are necessary to reduce the image to negligible proportions above about 15 Mc/s.

Two stages of 460 kc/s amplification (six tuned circuits) can provide a fair degree of selectivity without any additional measures such as those described later; typical bandwidths are 6 kc/s nose, 20-30 kc/s skirt. This would be passable for a.m. telephony, though only fair for c.w. reception, and adjacent channel interference would almost certainly be a major problem on crowded bands.

With an i.f. of 1·6 Mc/s, selectivity of the order necessary for modern amateur requirements would be extremely difficult to achieve without the use of crystal filters, though for simple sets selectivity can be considerably improved by using regeneration. A low i.f. of 85-100 kc/s can give with four tuned circuits (two i.f. transformers) a peaked nose bandwidth of the order of 3 kc/s and a skirt bandwidth better than 15 kc/s at −60db.

Double-conversion Receiver

The conflicting requirement of a low i.f. for good selectivity and a high i.f. for good image protection has led to the development of receivers having two or even three changes of frequency. In the double-conversion set, the incoming signals are first converted to a fairly high intermediate frequency, of the order of 1·6-5 Mc/s, making possible a reduction of image response to negligible proportions with only one r.f. stage, and so avoiding the complications with the ganged tuning of more than three circuits. Then in turn these i.f. signals are changed, usually with the aid of a fixed tuned local oscillator, to a lower i.f. where it is possible to obtain a high degree of selectivity. This second i.f. is often within the range 50-100 kc/s, but may be in the standard i.f. range of 455-470 kc/s. It is at the second i.f. that the main amplification is achieved and the main shaping of the response curve occurs. A block diagram of a representative arrangement is shown in Fig. 4.23. Thus a double conversion receiver can offer both good image protection and high selectivity, but at the cost of added complexity in design and construction. Great care is necessary to avoid the reception of spurious signals resulting from harmonics of one or more of the local oscillators; there is also the risk that the full benefits of the high selectivity may be lost owing to cross modulation resulting from the overloading of one of the early stages, since selectivity is not achieved until fairly late in the receiver, unless a second crystal filter (though not necessarily with such rigorous requirements) is inserted after the first mixer.

The use of a much higher first i.f.—above 30 Mc/s—has some attractions; including reduction of image response and allowing continuous coverage of all frequencies up to 30 Mc/s. This approach has been used in some commercial designs (e.g. Racal RA17, RA217, Plessey PR155).

The RA17 introduced the ingenious Wadley triple-mix tuning principle allowing the drift of the local oscillator to be automatically eliminated and stability determined by a 1 Mc/s crystal oscillator. Although a simplified form of this system has been used in some home-built sets, the requirement of the bandpass first i.f. filter (40 Mc/s ± 0·65 Mc/s) presents difficulties. A fixed first i.f. of 37·3 Mc/s is used in the PR155, allowing a selective filter to be inserted immediately after the first mixer.

There is one particularly interesting form of double-conversion technique used in many of the more elaborate sets and now finding increased favour with experienced amateurs. In this system the frequency of the first local oscillator is fixed for each band, usually by a crystal, and the output from the first frequency changer represents a band of frequencies which remains constant. The first i.f. section is then made tunable over a definite band, for example from 3·5 to 4 Mc/s, being converted to a lower intermediate frequency by means of a variable local oscillator, see Fig. 4.24. The entire receiver thus can be considered as a series of converters, one for each band, ahead of a single wave-range superhet receiver which, because of the absence of wave-change switching and with the local oscillator always working on a relatively low fixed h.f. range, can be made extremely stable and retain accurate calibration over a long period. A very wide amateur band, such as 28-29·7 Mc/s, would be split into two or more sections. This principle is also widely used for h.f. or v.h.f. reception in front of a lower frequency receiver: the converter is fixed tuned, with the actual tuning done on the main receiver, see Fig. 4.25.

In the conventional form of double conversion receiver with fixed first and second intermediate frequencies, the second oscillator is normally fixed tuned, but an effective form of fine bandspread tuning is possible by making the

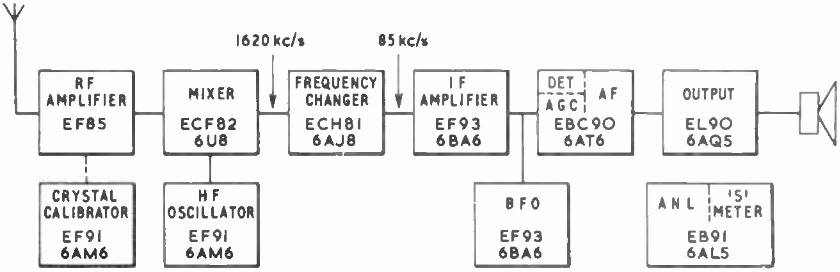


Fig. 4.23. Block diagram of a double conversion communications receiver.

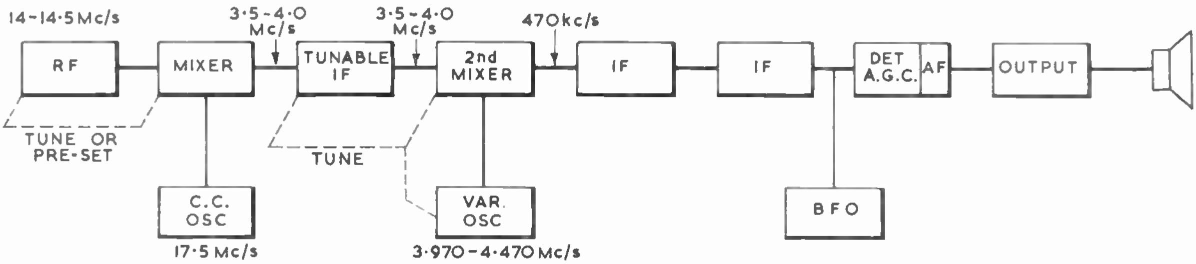


Fig. 4.24. Outline of double-conversion receiver with crystal-controlled h.f. oscillator and variable first i.f.

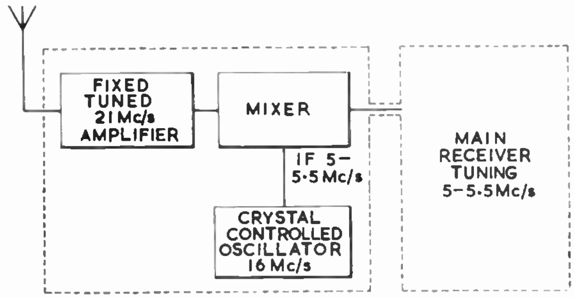


Fig. 4.25. How a fixed tuned converter is often used ahead of a communications receiver. All tuning is done on the main receiver.

second oscillator variable, by means of a panel control, over a range of about 5 kc/s: the bandwidth of the first i.f. section is usually more than adequate to permit this form of fine tuning without loss of gain.

Careful attention must be given to the screening and decoupling of the second oscillator if *birdies* and other spurious responses are to be avoided. Often the second frequency-changer (a combined mixer-oscillator such as the EK90/6BE6 is suitable) can be totally enclosed in a metal screening box and all leads emerging from the box decoupled to earth. The second oscillator must be designed for high stability although, as it will normally be working on a moderately low frequency, this is usually not so difficult a problem as the design of the first h.f. oscillator. So that the second oscillator does not operate on a frequency covered in the tuning range of the receiver, it is usual for this to be set on the low frequency side of the first i.f. In more complex receivers, this oscillator may be crystal controlled: sometimes two crystals are provided—one either side of the first i.f.—to permit side-band selection.

Occasionally designers include an amplifier stage at the first i.f. This arrangement, however, tends to increase the susceptibility of the receiver to cross modulation and blocking.

I.F. Selectivity

The degree of selectivity which can be achieved at 450 kc/s or above, solely by the use of two or three double-tuned i.f. transformers of conventional design, is not likely to prove sufficient for optimum results on both c.w. and telephony in the crowded amateur bands. In some designs, a higher degree of selectivity has been achieved by increasing the number of tuned i.f. circuits to eight or even twelve, each of high Q construction and under-coupled. Since the transfer of energy tends to be low because of the low transformer coupling factor, more overall valve amplification must be available, and three stages of i.f. amplification may be needed. Fig. 4.26 indicates some of the ways in which tuned circuits can be incorporated in the inter-valve couplings.

The selectivity which can be obtained is governed both by the Q of the i.f. transformer and its coupling factor, and good components should be used. Tuned circuits of higher Q than can readily be obtained in the conventional type of i.f. transformer (even when wound with Litz multi-stranded wire) can be obtained by using modern pot cores of high permeability ferrite materials. To provide switched degrees of selectivity, some receivers incorporate i.f. transformers having a low coupling factor but with a small additional primary winding tightly coupled to the secondary and which can be switched into circuit to broaden the response curve. This and the pot core systems are very effective for low frequency second i.f. stages.

Where selective i.f. filters are incorporated, it is most important that stray couplings and signal paths between the stages be kept to a very low figure, or signals will leak round the filter(s) and impair selectivity. Even a relatively small leakage becomes of consequence at −60 and −80db levels. Good layout and effective decoupling of the anode, screen, cathode and heater circuits are of particular importance. Decoupling capacitors should be chosen for low series inductance and with short connecting leads.

I.F. Regeneration

The gain of an i.f. amplifier can be considerably increased by the use of controlled positive feedback up to the point of oscillation. Such feedback also has the effect of providing an apparent increase of Q of the tuned i.f. circuits; this increases the selectivity of the stage. Since the signal-to-noise ratio will have already been largely determined in the early stages, the increase in noise contribution due to regenerative feedback will be of little practical consequence.

Fig. 4.27 shows a simple method of providing controlled regeneration without additional windings on the i.f. transformer. Positive feedback is introduced by varying the effectiveness of the screen decoupling capacitor.

In practice, the use of i.f. regeneration is confined mainly to smaller receivers where the extra gain is of most importance. This is partly because the improvement in selectivity is impaired in the presence of strong signals (where it is most

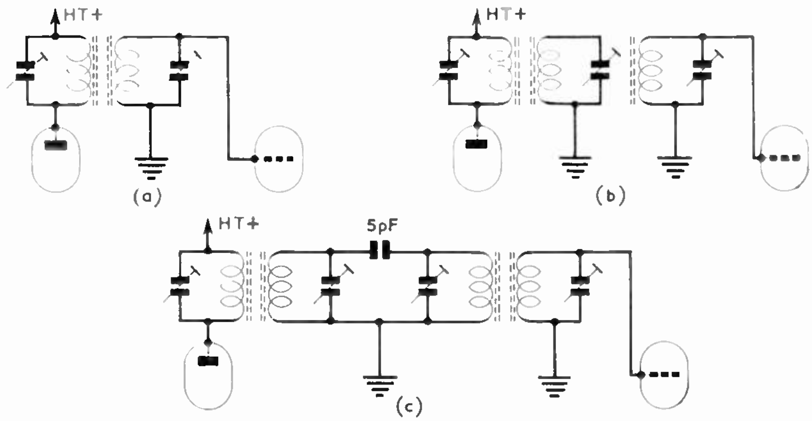


Fig. 4.26. I.F. inter-valve couplings. (a) The conventional double-tuned i.f. transformer. (b) Special triple tuned i.f. transformers have been used in some designs to provide a useful improvement in skirt selectivity. (c) Two conventional i.f. transformers can be connected in this way to provide four tuned circuits per stage.

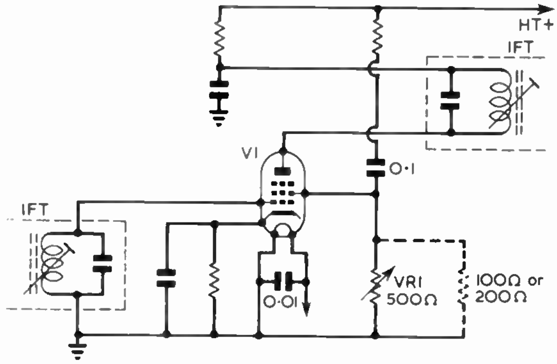


Fig. 4.27. Controlling regeneration of an i.f. amplifier by varying the efficiency of the screen decoupling. The fixed resistor in parallel with the control may have to be added if the regeneration control is too coarse.

needed) and partly because of the difficulty of maintaining accurate alignment at all positions of the regeneration control.

Crystal Filters

The selectivity of a tuned circuit is governed by its frequency and by its Q (ratio of reactance to resistance). There are practical limits to the Q obtainable in coils and i.f. transformers. In 1929, Dr. J. Robinson, a British scientist, introduced the quartz crystal resonator into radio receivers. The advantages of such a device for communication receivers were appreciated by James Lamb of the American Radio Relay League and he made popular the i.f. *crystal filter* for amateur operators.

For this application a quartz crystal may be considered as a resonant circuit with a Q of from 10,000 to 100,000 compared with about 300 for a very high grade coil and capacitor tuned circuit. From Chapter 1, it will be noted that the electrical equivalent of a crystal is not a simple series or parallel tuned circuit, but a combination of the two: it has (a) a fixed series resonant frequency (f_s) and (b) a parallel resonant frequency (f_p). The frequency (f_p) is determined partly by the capacitance of the crystal holder and by any added parallel capacitance and can be varied over a small range.

The crystal offers low impedance to signals at its series resonant frequency; a very high impedance to signals at its parallel resonant frequency, and a moderately high impedance to signals on other frequencies, tending to decrease as the frequency increases due to the parallel capacitance.

While there are a number of ways in which this high Q circuit can be incorporated into an i.f. stage, a common method—providing a variable degree of selectivity—is shown in Fig. 4.28. When the series resonant frequency of the crystal coincides with the incoming i.f. signals, it forms a sharply tuned "acceptor" circuit, passing the signals with only slight insertion loss (loss of strength) to the grid of the succeeding stage. The exact setting of the associated parallel resonant circuit, at which the crystal will offer an extremely high resistance, is governed by the setting of the *phasing control* which balances out the effect of the holder capacitance. Such a filter can provide a nose selectivity of the order of 1 kc/s bandwidth or less, while the sharp rejection notch which can be shifted by the phasing control through the pass band can be of the order of 45db. Fig. 4.29 shows the improvement which can be obtained by switching in a filter of this type in a good communications receiver (a simple method of switching the filter is to arrange the phasing trimmer to short-circuit at one end of its travel). Inspection of the response curve will show that the improvement in the skirt selectivity is not so spectacular as at the nose, and even with a well-designed filter may leave something to be desired in the presence of strong signals.

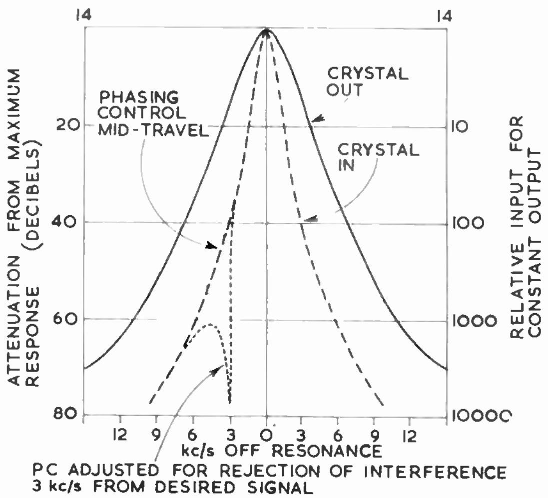


Fig. 4.29. A graph showing the improvement in selectivity which can be obtained by the use of a crystal filter of the type shown in Fig. 4.28.

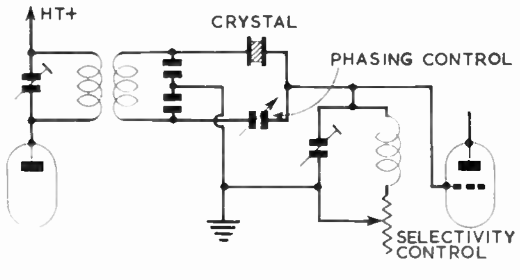


Fig. 4.28. Variable selectivity crystal filter. Selectivity is greatest when the impedance of the tuned circuit is reduced by bringing the variable resistor fully into circuit. For optimum results there must be adequate screening to prevent stray coupling between the input and output circuits which would permit strong off-resonance signals to leak round the filter.

The degree of selectivity provided by a single crystal depends not only upon the Q of the crystal and the i.f. but also upon the impedance of the input and output circuits. The lower these impedances are the greater will be the effect of the filter, though this will usually be accompanied by a rise in insertion loss. To broaden the selectivity curve and make the Q of the filter *appear* less, it is only necessary to raise the input or output impedances. In Fig. 4.28 the input impedance may be lowered by detuning the secondary of the i.f. transformer. The output impedance will depend upon the setting of the variable resistor which forms the selectivity control. With minimum resistance in circuit, the tuned

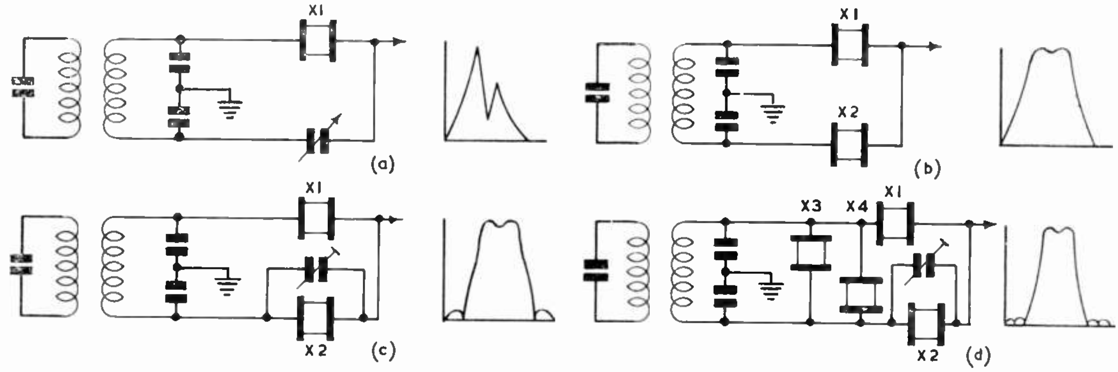


Fig. 4.30. Half-lattice crystal filters showing the improvement in the shape of the curve which can be obtained when the crystals are correctly balanced or when extra crystals are used to reduce the 'humps.' Typical crystal frequencies would be X1 464·8 kc/s, X2 466·7 kc/s, X3 463 kc/s, X4 468·5 kc/s.

circuit will offer maximum impedance, which can be gradually lowered by bringing more resistance into circuit. Maximum impedance corresponds with minimum selectivity.

A disadvantage of the single crystal filter is that when used in the position of maximum selectivity, it may introduce considerable " ringing," rendering it difficult to copy a weak c.w. signal: this is due to the tendency of a high *Q* circuit to oscillate for a short period after being stimulated by a signal, producing a bell-like echo on the signal.

Since the minimum nose bandwidth of a single crystal filter may be as low as 100-200 c/s, it is not possible to receive a.m. signals through the filter unless its efficiency is degraded by a high impedance or by " stenode " tone correction (see page 4.56). If this is done, such a filter will often prove most useful for telephony reception through bad interference, though at some cost to the quality of reproduction. Even the most simple form of crystal filter, consisting of a crystal in series with the i.f. signal path, without balancing or phasing, can be of use.

Bandpass Crystal Filters

As already noted, the sharply peaked response curve of a single-crystal filter is not ideal, and has by modern standards a relatively poor shape factor: improved results can be achieved with what is termed a *half-lattice* or bandpass filter. Basically, this comprises two crystals chosen so that their series resonant frequencies differ by an amount approximately equal to the bandwidth required; for example about 300 c/s apart for c.w., 3-4 kc/s apart for a.m. telephony or 2 kc/s apart for s.s.b., see **Fig. 4.30(b)**. This form of filter, developed in the 'thirties, has in recent years come into widespread amateur use. Although it has a much improved slope over the single crystal filter, in its simplest form there will still be certain frequencies, just outside the main pass band, at which the attenuation is reduced. Unless a balancing trimmer is connected across the higher frequency crystal the sides of the response curve tend to broaden out towards the bottom. As the capacitance across the crystal is increased, the sides of the curve steepen, but the side lobes tend to become more pronounced. Capacitance across the lower frequency crystal broadens the response and deepens the trough in the centre of the passband.

To eliminate the " humps " in the response curve additional crystals may be included in the filter, these may be up to about six in number, **Fig. 4.31**. Examples of bandpass filter response characteristics are shown in **Fig. 4.30**. Alternatively, additional filter sections may be incorporated in the i.f. section in cascade. An advantage of using several cascaded filters is that less critical balancing and adjustment are needed. Provided that there is no leakage of i.f. signals around the filter due to stray capacitances or other forms of unwanted coupling (an important consideration with all selective filters), extremely good shape factors of the order of 1·5 can be achieved with about three cascaded filters on about 460 kc/s. This system

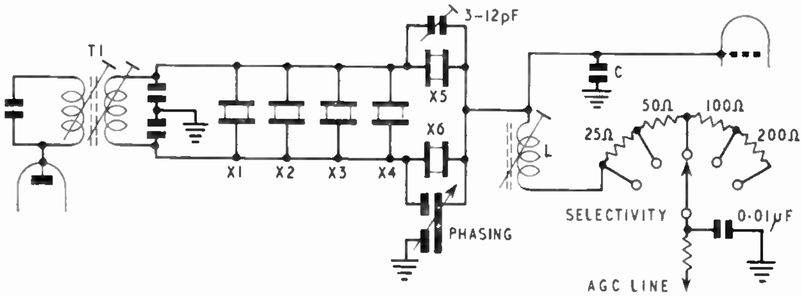


Fig. 4.31. Up to about six crystals may be used in a single half-lattice filter. Here is a variable selectivity unit for 465 kc/s using FT241 crystals. X1 461·1 kc/s (49); X2 462·9 kc/s (50); X3 468·5 kc/s (53); X4 470·4 kc/s (54); X5 464·8 kc/s (51); X6 466·7 kc/s (52). Numbers in brackets refer to channel numbers for FT241 crystals. T1 and L/C should be tuned to mid-filter frequency.

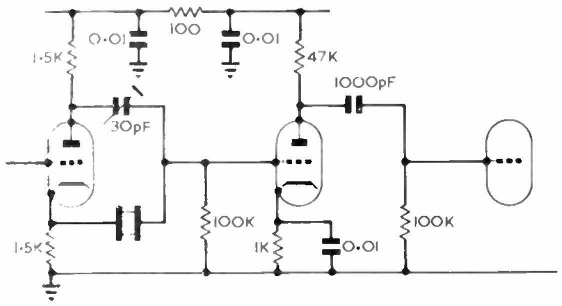


Fig. 4.32. A crystal may be used in conjunction with an otherwise aperiodic (untuned) amplifier. In practice several amplifiers may be cascaded. Typical valves would be the triode ECC85/6AQ8 or the triode-pentode ECF80/6BL8.

is used in the receiver described on page 4.56. There will be an insertion loss of the order of 6db per section.

To provide a sharp variable rejection notch, it is advantageous to incorporate some form of bridged-*T* filter or *Q* multiplier (see later) in order to be able to eliminate steady heterodyne interference.

To reduce susceptibility to cross-modulation or blocking, crystal and other selective filters should be placed as near as possible to the front-end of the receiver.

Where the first fixed i.f. is too high for a crystal filter to be completely effective, it is nevertheless advantageous to include a preliminary filter at this point, with the more selective filter later in the receiver (but still at the earliest practicable stage).

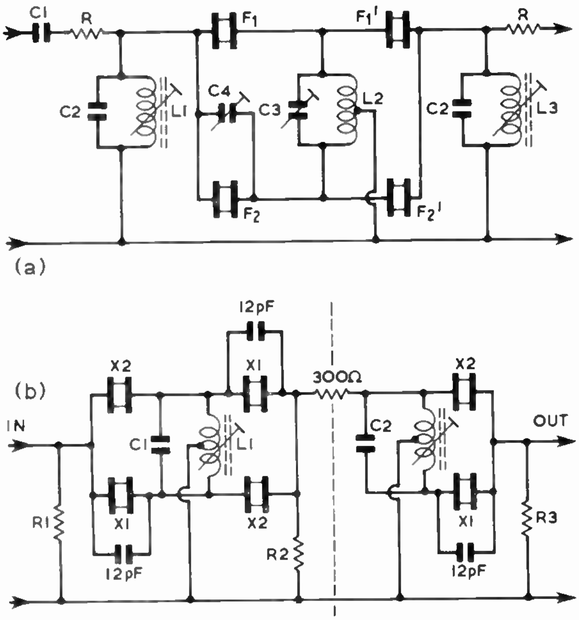


Fig. 4.33. Typical h.f. crystal filters using FT243 crystals.
(a) Filter using four crystals (F1 = F1′; F2 = F2′ = F1 + 1500 to 2500 c/s); C1, 2000 pF; C2, 47 pF; C3, 3–30 pF; C4, 3–10 pF. L1 and L3 to resonate at filter frequency with C2. L2 to resonate with about 15 pF setting of C3. R should be 2000 ohms. (b) Filter using six crystals and capable of "nose" passbands of 2·4 3 kc/s and −60 db skirt bandwidth of 6 7 kc/s. The sets of X1 and X2 crystals should be separated by 1·7 kc/s in series resonant mode. R1, R2 and R3 are 560 to 820 ohms.

Crystals for Bandpass Filters

For bandpass filters, most amateurs use surplus type FT241 crystals. These are in two groups labelled in frequency and with a channel number. The fundamental frequencies are from 370 to 540 kc/s. The first group are marked in frequencies from 20 to 27·9 Mc/s with channel numbers up to 79. The actual crystal frequency may be found by dividing the frequencies marked on the holder by 54. As the channels are spaced 0·1 Mc/s apart, the crystal separations are 1·85 kc/s. The second group of FT241 crystals are marked from 27 to 38·9 Mc/s in 0·1 Mc/s steps, with channel numbers from 270 to 389; the fundamental frequency can be found by dividing the marked frequency by 72. The spacing of this group is about 1·49 kc/s. FT241 crystals of similar channel numbers are likely to show some slight spread of frequencies. Should the precise frequencies required not be available, it is possible to raise the frequency of a crystal by edge grinding and to lower it by plating.

AN/TRC-1 crystals in similar holders to the FT241 series are available in fundamental frequencies from 729 to 1040 kc/s and marked in channel numbers from 70 to 99·9.

High Frequency Crystal Filters

As the frequency increases, it becomes more difficult to obtain entirely satisfactory characteristics from half-lattice crystal filters. Whereas extremely good results can be achieved on 450-470 kc/s, bandpass crystal filters for intermediate frequencies of 1·6 Mc/s and above may prove less effective when viewed by the highest modern standards, particularly when using the readily available surplus crystals. Since many of the problems associated with superhet design would be solved if sufficient selectivity could fairly easily be obtained with high intermediate frequencies, the development of suitable but inexpensive h.f. crystal filters is a matter of concern to amateurs. Excellent factory-built 9 Mc/s filters developed for s.s.b. applications are available at reasonable cost. Filters at v.h.f. have been used in commercial designs.

One current line of development of interest to home-constructors is the use of a number of crystals in place of i.f. transformers in an aperiodic (untuned) amplifier. This type of selective amplifier is particularly suited for receivers intended primarily for c.w. reception. A typical circuit of a single filter is shown in **Fig. 4.32.** In practice at least two filters would be cascaded.

Effective h.f. crystal filters using FT243 crystals between 5·5-6·5 Mc/s have been built successfully by a number of amateurs. Typical designs are shown in **Fig. 4.33.**

The possibility of using bandpass crystal filters at signal frequencies has been shown to allow a number of transmitters to be operated in close proximity.

Mechanical Filters

The development of mechanical filters during the past ten years or so has provided the amateur designer with a convenient method of obtaining—at intermediate frequencies between 60 and 600 kc/s—a bandpass characteristic having almost any desired bandwidth combined with a flat top and

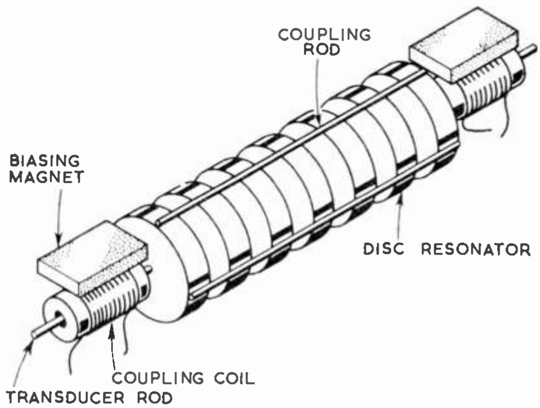


Fig. 4.34. The Collins mechanical filter. I.f. signals are converted into mechanical vibrations by a magneto-strictive transducer and passed along a series of resonant discs, then finally reconverted to i.f. signals by a second magneto-strictive transducer. The bandwidth of the filter is governed by the number of resonant discs and the design of the coupling rods.

steep sides. These filters are smaller than the usual i.f. transformer and provide in one small unit a stable filter having the characteristics of a multiple crystal network. A functional diagram of a mechanical filter is shown in **Fig. 4.34.**

The filter consists of three basic elements:

(a) Two magneto-striction transducers which convert the i.f. signals into mechanical vibrations and vice versa.

(b) A series of metal discs mechanically resonated to the particular frequency.

(c) Disc coupling rods.

Each disc is equivalent to a series resonant circuit with a high Q (of the order of 5000 or above). Varying the mechanical coupling between the rods—by making them larger or smaller—varies the bandwidth of the filter. The vibration of the discs cannot be observed, and the entire filter is enclosed in a hermetically sealed case. The insertion loss of one of these filters is about 10db, and the only adjustment required is to tune the input and output transducer coils with the appropriate capacitance. Normally, non-critical high impedance circuits are suitable at the input and output ends; for use with transistors however, the filters can be matched into low impedance circuits by using a series-resonant termination. To obtain the full benefit of the filter, the input and output circuits must be effectively screened from each other.

Mechanical filters cannot at present be home-constructed, and must be purchased complete. A pair of filters, one for c.w. the other for telephony—adds many pounds to the cost of a receiver, and an amateur can often construct crystal filter networks at appreciably lower cost.

In recent years, the availability of both American and Japanese mechanical filters has tended to lower prices, and make them more competitive with crystal filters. For s.s.b. applications there is some advantage in the use of assymmetrical filters, with the steepest slope on the " carrier " side.

Bridged-T Filters

A further method of providing a rejection notch which can be used for the elimination of steady heterodyne interference is the *bridged-T* filter.

The basic filter is shown in **Fig. 4.35** with its associated response curve, which indicates that the filter provides high attenuation at the resonant frequency. This attenuation will be maximum (providing the deepest notch) when the resistor is equal to one-quarter the impedance of the tuned circuit. By providing a variable resistor it thus becomes possible to adjust the notch depth.

Filters of this simple form are to be found mainly in receivers having a low i.f., usually below 100 kc/s. However, effective use of this type of filter can be made at 450-470 kc/s by increasing the Q of the filter coil by means of the Q multiplication technique, as indicated in **Fig. 4.36.** An advantage of the bridged-T filter over the phasing control of a crystal filter is that it causes less distortion of the i.f. response on either side of the deep, narrow notch.

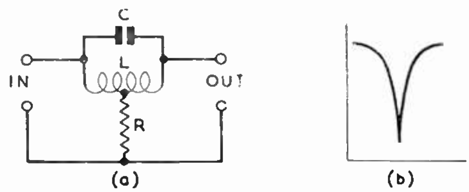


Fig. 4.35. The basic bridged-T filter circuit (a) and associated response characteristics (b) showing deep notch at resonant frequency. The notch attenuation will be maximum when *R* is equal to one-quarter of the resonant impedance of *LC*. In practice two equal capacitors are often used in a capacitive potentiometer instead of a centre-tapped coil.

Ceramic Filters and Transformers

Piezoelectric effects are not confined to the classic materials such as quartz and Rochelle salt crystals. In recent years increasing use has been made of certain ceramics, such as

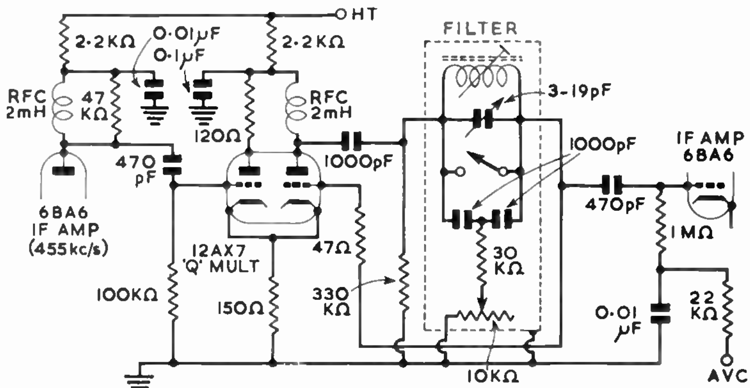


Fig. 4.36. Bridged-T circuit combined with a Q Multiplier. The notch can be tuned across the receiver bandpass by means of the 3-19 pF trimmer and the depth of the notch adjusted by the 10 K variable resistor.

barium titanate (for gramophone pickups) and the lead zirconate titanate (PZT) series. Small discs of PZT ceramics which resonate in the radial dimension to the intermediate frequency can be used in bandpass filters in the same way as quartz crystals, though at the present state of development the *Q* is considerably lower. PZT ceramics can also be used directly to provide an impedance transformation: by providing two sets of electrodes a complete replacement for an i.f. transformer can be obtained, giving improved selectivity characteristics compared with conventional components. To date this type of piezoelectric ceramic i.f. transformer has been used mainly for transistor receivers, as it provides a very convenient means of obtaining the low impedances required for transistor input circuits. While having a good nose selectivity the skirts tend to be too broad for ceramic filters to be much used in advanced amateur receivers at present.

Q Multipliers

The apparent *Q* of a tuned circuit and hence its selectivity can be increased by connecting it to an amplifier with regenerative feedback. Such an amplifier can be regarded as a negative resistance which will cancel some of the loss resistance in the tuned circuit thereby raising the *Q*. This arrangement is familiar to many in t.r.f. receivers. In recent years it has been applied to a variety of devices generally known as *Q* multipliers and of great value in superhet receivers. Basically these devices comprise a simple valve or transistor amplifier with adjustable positive feedback. This is coupled to an early i.f. coil in such a way that regeneration can be applied up to the threshold of oscillation. A more elaborate version contains a tunable circuit which allows the narrow peak to be varied in frequency within the normal band-width of the receiver. A further useful feature of the more elaborate version is that the tuned circuit (with its *Q* raised by regeneration) can be re-connected, by switching, in the negative feedback circuit of a triode amplifier. The triode with negative feedback performs the opposite function to that described above and presents the i.f. transformer with additional loss resistance depending upon the tuning of the *Q* multiplier tuned circuit. At resonance drastic damping occurs and a sharp rejection notch results which is tunable over the pass band similar to the action of a crystal filter with phasing control.

The *Q* multiplier is a worthwhile addition to any superhet receiver, even those already possessing a bandpass crystal or mechanical filter.

It should be connected to the earliest possible stage in the receiver so that the enhanced selectivity can give the maximum protection against cross modulation in the later stages which operate at higher input voltage levels. The point usually chosen is the anode of the mixer in single superhets or the anode of the second mixer in double superhets.

The *Q* multiplier technique can be extended to the use of crystals as the basic high *Q* tuned circuit (suitable primarily for c.w. reception); alternatively use may be made of the existing i.f. transformer to provide *Q* multiplication without additional coils. The principle has also been successfully adapted for improving pre-mixer selectivity at signal frequencies.

Constructional details of a *Q* multiplier are given later in this chapter.

THE DETECTOR

The main types of detector in modern receivers are:

(a) The diode detector, using either a thermionic or crystal diode, found in the vast majority of receivers.

(b) The regenerative leaky-grid detector, relatively little used in superhets but capable of high gain and good selectivity.

(c) The infinite impedance detector, offering extremely low loading of the final i.f. transformer.

(d) Various forms of heterodyne detector developed primarily for the reception of s.s.b. and c.w. signals.

Diode Detector

The conventional diode envelope detector functions directly on a.m. signals and in conjunction with a b.f.o. or carrier insertion oscillator on telegraphy and s.s.b. It is an uncomplicated device presenting a stable but rather low impedance load to the final i.f. transformer, and for many years was used almost exclusively. It has the disadvantage that it provides no gain, and may introduce distortion at low signal levels. Because of its relatively low dynamic impedance and consequent a.c. shunting on the final i.f.transformer it is advantageous for the secondary winding of this transformer to have a lower *L*/*C* ratio than for the other windings, and to be more tightly coupled than for the earlier transformers; however, in practice, satisfactory results can be obtained using similar i.f. transformers throughout a set. To minimize a.f. distortion on weak signals having a high modulation percentage, care should be taken to keep the capacitance across the diode load as low as possible.

The diode detector is often combined in a single multiple valve with a second diode used as a rectifier to provide a negative a.g.c. bias and also a triode which is generally used for the first stage of a.f. amplification: see Fig. 4.37. This practice can introduce undesirable coupling between the various circuits, but is economical.

Suitable valve types for the detector and a.g.c. diodes would be the EB91/6AL5, EB34/6H6 or crystal diodes types

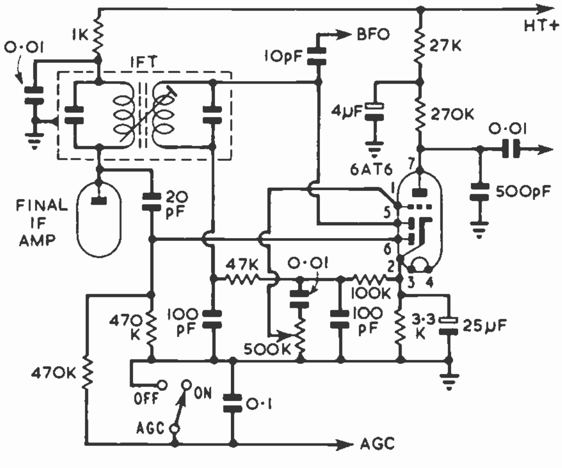


Fig. 4.37. Typical use of a double triode valve for detection, a.g.c. and a.f. amplification. A.G.C. delay is obtained from the cathode bias but this is not applied to the detector diode.

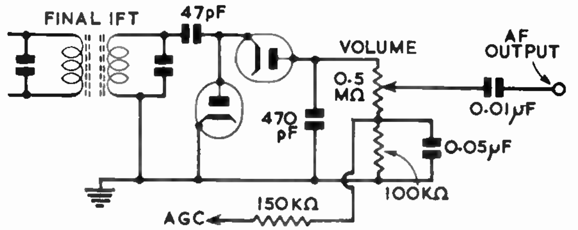


Fig. 4.38. Diode integrator detector. Either thermionic valves (6AL5, 6H6) or semi-conductor diodes (IN41, OA79, OA81, etc.) can be used.

OA79, OA81, IN41. Suitable double-diode and triple-diode triodes include the 6AT6/EBC90, 6AV6/EBC91 and the 6AK8/EABC80.

It is possible to use two diodes in a voltage doubler arrangement, offering some advantages for weak signal reception at low distortion, see Fig. 4.38.

Regenerative Detector

The *regenerative detector* offers the advantages that a high gain can be obtained (permitting its use on weak i.f. signals) and that it can be made to contribute to the selectivity of the receiver. Its main disadvantage is that it is difficult to use in conjunction with an a.g.c. system; furthermore it needs very careful adjustment if full benefit is to be obtained. It is particularly useful for simpler receivers intended primarily for c.w. reception; it dispenses with the need for a separate b.f.o. and can do much to overcome a deficiency of i.f. gain.

I.F. transformers incorporating a feedback coil suitable for this purpose are not readily available. One method is to remove the i.f. transformer from its screening can and to add a feedback winding, finding the correct number of turns for smooth regeneration by trial and error. An alternative method is described on page 4.47.

Infinite Impedance Detector

The *infinite impedance* detector, Fig. 4.39, has certain advantages but has until recently only rarely been used in communications receivers. At audio frequencies, there is considerable negative current feedback across the relatively high value cathode resistor, reducing distortion to a very low level. As the triode valve is not driven into grid current, there is little damping of the i.f. transformer, enabling a higher voltage to be developed across the secondary winding than would be possible with a diode detector. Output from the b.f.o. or c.i.o. is fed to the grid of the detector via a small capacitance. It is difficult to use in conjunction with an a.g.c. system unless a separate a.g.c. i.f. amplifier is incorporated, although circuits to overcome this problem have been published (e.g. *Technical Topics for the Radio Amateur*—RSGB).

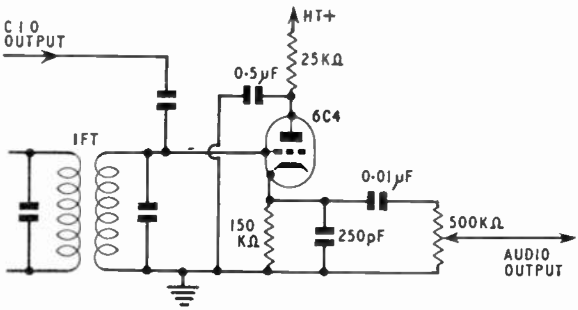


Fig. 4.39. Infinite impedance detector for use in a communications receiver.

Product or Heterodyne Detector

For the reception of s.s.b. signals various forms of mixer or heterodyne detector have been introduced to reduce intermodulation distortion at low signal levels. In the form

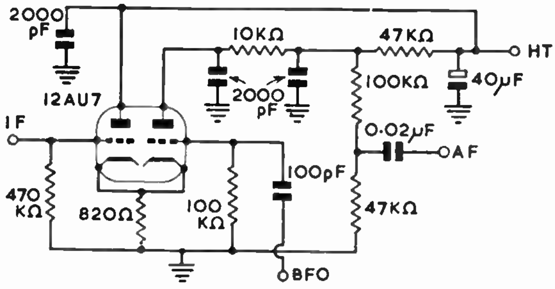


Fig. 4.40. The product or heterodyne detector is useful for the reception of s.s.b. and c.w. signals but is unsuitable for a.m. telephony signals. Suitable valves include 6SN7, 12AU7 and ECC82.

generally termed the *product detector* two input signals are fed to what is basically a mixer stage:

(a) The incoming signal at the intermediate frequency; and
(b) A signal from the carrier insertion oscillator (c.i.o.) or beat frequency oscillator (b.f.o.).

The difference in frequency represents the a.f. output and, after filtering to remove the r.f. components of the original signal, is fed to the a.f. stages. **Fig. 4.40** shows a typical circuit using a double-triode valve, though alternative arrangements using 6BE6 or ECH81 mixer valves are also commonly used. The reader will appreciate that these detectors can be used both for s.s.b. and c.w. but are unsuitable for the reception of a.m. telephony transmissions,

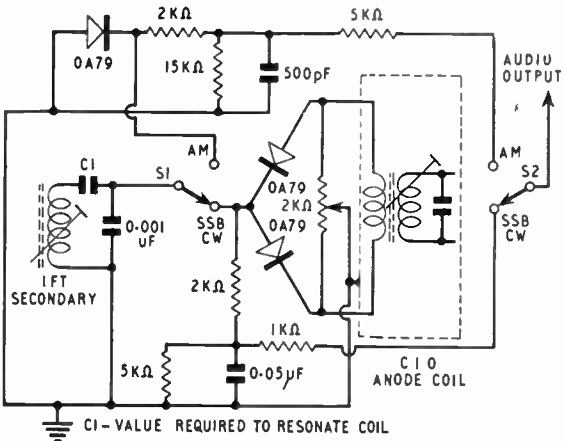


Fig. 4.41 (a). This arrangement using three germanium diodes provides demodulation for s.s.b., c.w. and a.m. signals and is incorporated in the receiver described on page 4.62.

except when the signals are converted to s.s.b. by selective filters.

Another form of heterodyne detector, capable of excellent s.s.b. performance, is shown in **Fig. 4.41a.** This uses two germanium crystal diodes in a bridge ring modulator (as for s.s.b. generation) in conjunction with a further single diode which can be switched into circuit for a.m. reception. The resonating capacitance across the secondary of the final i.f. transformer in the receiver comprises two fixed capacitors with a capacitance ratio of approximately 10: 1 to form an a.c. potentiometer, providing a low impedance load to the diodes. The c.i.o./b.f.o. voltage is fed in push-pull across a balancing potentiometer from a low impedance link winding in the oscillator anode circuit. When correctly balanced by means of the 2 K ohm potentiometer, there is no output when the c.i.o. voltage is removed. This type of demodulator is used in the advanced amateur receiver shown on page 4.63.

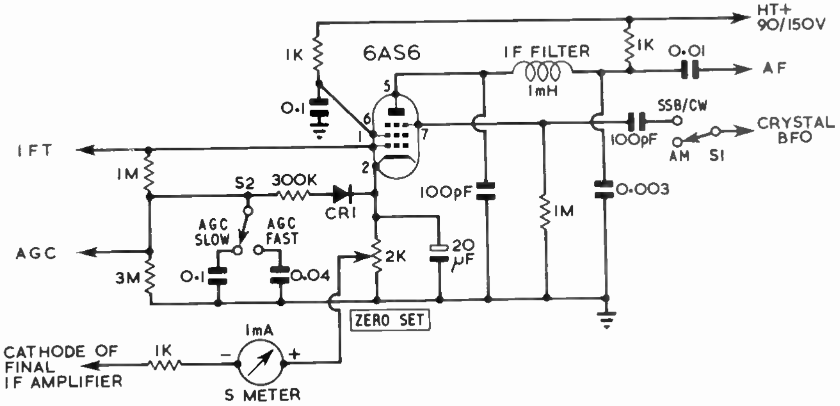


Fig. 4.41(b) Philco circuit for combined a.m./s.s.b./c.w. detector/a.g.c. system. CR1, 1N91 or similar crystal diode.

Beam deflector valves (e.g. 7360) used as product detectors can provide much greater audio output than conventional circuits at the expense of greater circuit complexity.

A detector/a.g.c. circuit designed for a.m./c.w./s.s.b. service without the need for independent detectors with other product detectors and permitting a.g.c. to be used on all modes is shown in **Fig. 4.41b.**

The arrangement consists basically of a 6AS6 valve used as a grid-leak power detector. When S1 is in the A.M. position, the b.f.o. is disconnected. The incoming i.f. signals are rectified in the grid/cathode circuit and the detected signal amplified in the anode circuit. A.G.C. is taken off from the cathode circuit with the 1N91 crystal diode used to delay the operation of the a.g.c. until the signal strength reaches a satisfactory level. S2 selects the a.g.c. time-constant.

On SSB/CW the b.f.o. signal is fed to the suppressor grid of the 6AS6 and mixes with the incoming signals in the anode circuit. Since the b.f.o. (which may or may not be crystal-controlled) is isolated from the detector circuit, it should not affect the a.g.c. level. The 1 K ohm resistors in the anode and screen leads are for isolation. To some extent exact values of components will depend on supply voltage, and the associated stages.

The Beat Frequency Oscillator

The *beat frequency oscillator* (b.f.o.), termed for s.s.b. reception the *carrier injection oscillator* (c.i.o.), exercises considerable influence on receiver performance on c.w. and s.s.b. signals. Any variation in oscillator frequency will be reflected in variation of the a.f. beat note on a c.w. signal, and will have the effect of detuning the receiver on s.s.b. Since the b.f.o. is tuned to a frequency close to that of the receiver's final i.f., it is an easier matter to obtain the required stability with an L/C oscillator than for earlier frequency conversion oscillators. Nevertheless the requirements are fairly stringent, and modern s.s.b. designs tend to favour the use of crystal controlled oscillators.

An output of several volts is required for effective s.s.b. reception, but both harmonic output and stray coupling to receiver circuits other than the detector stage should be low, this calls for effective screening of the oscillator and good decoupling, including the heater circuit. The stage should be supplied from a stabilized h.t. line. There is some advantage to be gained from providing variable coupling to the detector, though this is rarely found in practice.

Fig. 4.42 shows a typical circuit. The variable capacitor C2 provides a control panel variation of about ± 5 kc/s, the knob often being labelled *pitch control*; the circuit should be carefully adjusted so that at mid-travel of this control, the oscillator frequency is at zero beat with the intermediate frequency. This permits easy change of a.f. beat note on c.w. signals, and provides for simpler adjustment of the b.f.o. when receiving s.s.b. Output from the b.f.o. can be taken from the cathode, a conveniently low impedance point, to minimize pulling of the oscillator by strong signals.

To permit immediate selection of the upper or lower sidebands in s.s.b. reception two carrier insertion frequencies—spaced approximately 3 kc/s apart—are useful, since this allows the carrier to be inserted on either side of the response

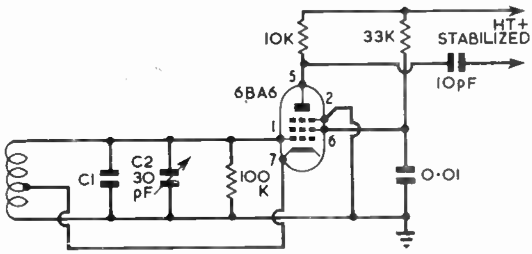


Fig. 4.42. Typical b.f.o. circuit using an electron-coupled oscillator. A high value of C1 will help stabilize the oscillator. C2 is the pitch control. When a b.f.o. is used with a product detector for s.s.b. it is sometimes called a c.i.o. (carrier insertion oscillator).

curve. This can be done by using two crystals, one above and the other below the i.f.

An even more elegant system is possible in multi-conversion sets by using two crystals in the final conversion oscillator. For example with a set having i.f.'s of 1·6 Mc/s and 450 kc/s suitable crystal frequencies would be 1150 kc/s and 2050 kc/s. This has the effect of reversing the sidebands and allows the use of an assymmetrical filter.

AUTOMATIC GAIN CONTROL

The more r.f. and i.f. stages there are in a superhet receiver, the greater will be the available gain and the greater the possibility of overloading by strong signals unless this gain can be reduced. This may be done manually by altering the bias applied to variable-mu valves by varying a common cathode bias resistor (there is a tendency in commercially-built receivers to use dual track resistors for this purpose with differing tapers for r.f. and i.f. valves). Since however, h.f. signals are subject to considerable slow and rapid fading, it is advantageous for the gain to be controlled automatically by the strength of the incoming signal. This can be done fairly simply for a.m. telephony signals; the basic circuits can also be adapted with some additional complexity for c.w. and s.s.b. reception. In simple amateur communications receivers, *automatic gain control or a.g.c.* (sometimes *automatic volume control* or *a.v.c.*) may be regarded as a luxury rather than an essential, for a poor a.g.c. system is often much worse than no a.g.c. at all.

The basic function of a conventional a.g.c. system is to hold the a.f. output from the detector reasonably constant over wide variations of signal level. This can be done by rectifying a portion of the amplified i.f. signal to obtain a negative d.c. voltage which is then applied as bias to one or more of the earlier stages. An increase in signal at the a.g.c. rectifier causes greater negative bias to be developed and so reduces overall gain; conversely, a fall in signal causes bias to be lowered and gain increases. The application of the a.g.c. voltage should be *delayed* until after the signal has reached a level sufficient to provide a good signal-to-noise ratio.

The greater the number of stages controlled by the *a.g.c. line* and the greater the gain within the control loop, the better will be the efficiency of the a.g.c. system—that is to say, the more constant will be the output for wide variations of signal strength.

It is not good practice to apply an a.g.c. potential to the mixer valve(s), although this is often done in simpler receivers. The efficiency of the mixer is reduced and the change in input capacitance with change of operating point can lead to pulling of the oscillator frequency. In receivers with two r.f. stages, the first stage may sometimes be operated under fixed bias conditions to prevent degradation of the signal-to-noise ratio.

A typical a.g.c. system is shown in **Fig. 4.43.** In this a portion of the i.f. signal is taken via a capacitor from the anode of the final i.f. amplifier to the a.g.c. rectifier diode and a negative d.c. voltage builds up across the load resistor. As it is important that no signal modulation should appear on the a.g.c. line, the output is fed through one or more sections of resistance-capacitance a.f. filters. The values of these filter components affect the rate at which the a.g.c. system follows the variations in signal strength. The larger the time-constant the slower will the a.g.c. respond to signal fluctuations. For a.m. telephony a time-constant of the order of about 0·2 second is commonly employed.

When an i.f. valve is controlled by an a.g.c. line, the screen is usually fed through a fairly high value resistor (about 22 K to 47 K ohms). This ensures that the screen voltage rises as the negative bias to the valve increases and provides a longer grid base for the valve. It also reduces the chances of a valve being driven into non-linear amplification, which would make cross-modulation or blocking more likely.

There are two main reasons why a conventional a.g.c. system, such as that described above, is unsatisfactory on c.w. and s.s.b. signals. First, in neither of these modes of transmission is there a continuous carrier on which to base the a.g.c. action. Secondly, the output from the b.f.o. acts on the a.g.c. system as though it were a strong external carrier so reducing the gain of the receiver.

To overcome the absence of a continuous carrier, as in telegraphy, the time constant of the a.g.c. system is lengthened so that the gain no longer rises and falls between each Morse symbol. The a.g.c. action is then based on the average signal received over a given period: whereas a discharge time constant of the order of 0·2 seconds is satisfactory for a.m. telephony, a more suitable figure for c.w. or s.s.b. is about 0·5—1·0 second. The charge time for the a.g.c. system

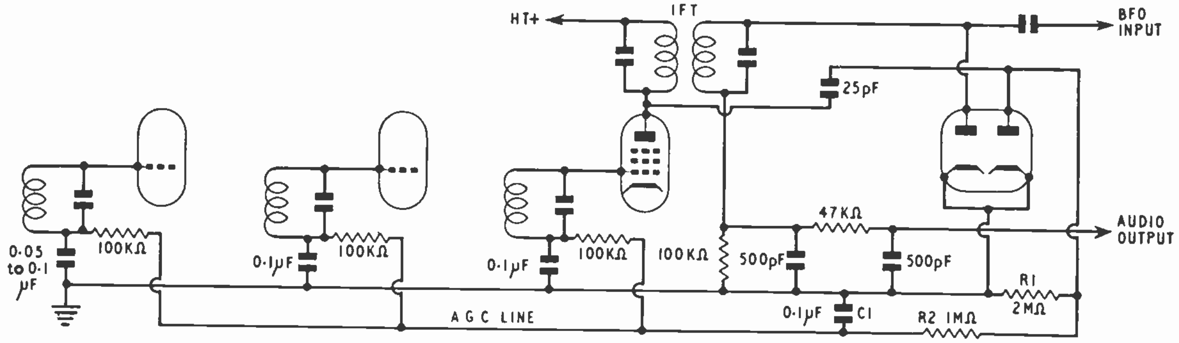


Fig. 4.43. Typical a.g.c. system and diode demodulator used for a.m. reception. The negative bias is used to control the gain of r.f. and i.f. amplifying stages.

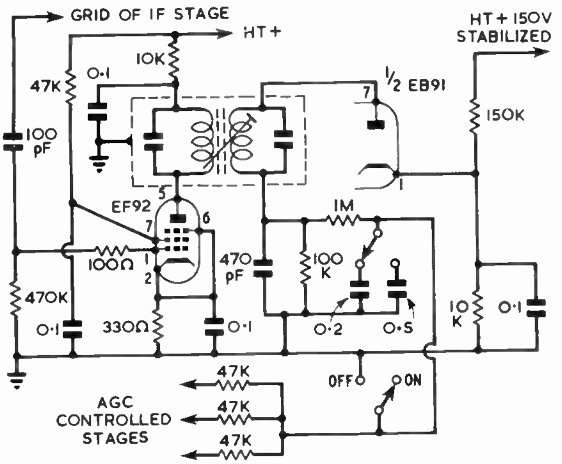


Fig. 4.44(a). The use of a separate a.g.c. amplifier and switched time constants enables the a.g.c. system to be used for both a.m. and c.w. reception.

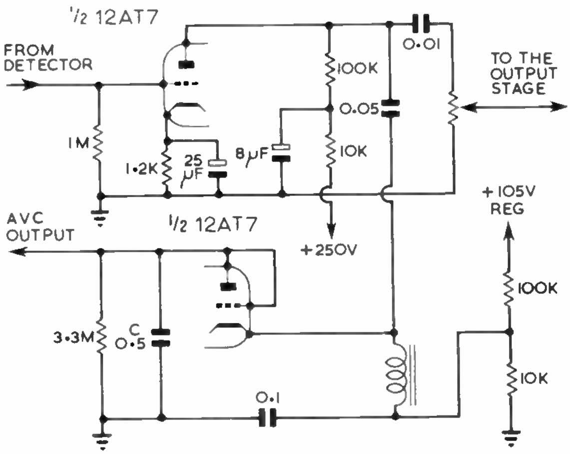


Fig. 4.44(b). Audio derived a.g.c. circuit.

should preferably be as short as possible, for example 25 milliseconds for a.m., 200 milliseconds for c.w./s.s.b.

The use of a " hang " a.g.c. system, with a brief rise time of about 10 milliseconds, a hang time (that is a time before any discharge begins in the absence of a carrier) of 400–500 milliseconds, and a discharge time of about 50 milliseconds, can allow the use of the same time constants on c.w., a.m. and s.s.b.

To prevent the b.f.o. from reducing gain, it is necessary to screen the b.f.o. thoroughly and then arrange a low level injection direct to the signal detector diode but it is difficult to overcome completely the effects of stray coupling. A more practical system is to feed the a.g.c. diode from an entirely separate a.g.c. amplifier, the input of which is in parallel with that of the final i.f. amplifier valve: see **Fig. 4.44(a).** With this system the signal for the a.g.c. system is thus taken off before the injection of the b.f.o. In some designs the take-off point is earlier and the selectivity characteristics of the a.g.c. system are made broader than the main i.f. channel; this permits the reduction of gain by strong adjacent-channel signals which might otherwise cause splatter or cross-modulation.

An alternative automatic *volume* control system which is applicable to all types of transmission consists of rectifying a small portion of the a.f. signal and using the voltage so obtained to control the output of the receiver.—see **Fig. 4.44(b).** The arrangement in its simplest forms tends to be unsatisfactory, however, as it is generally necessary to take the voltage from a point after the a.f. gain control. A more sophisticated arrangement which does not suffer from this disadvantage is shown in **Fig. 4.44(c).**

S METERS

A meter which indicates the approximate strength of the signals applied to the input terminals is a useful adjunct to a communications receiver. It can be used for comparing the strengths of different stations, for observing the results of tests on receiver or transmitter aerials, for checking the directional characteristics of a rotary aerial, and for use as a tuning indicator.

Many different circuits have been developed for this purpose, but all depend for their operation on the voltage developed across the a.g.c. line of the receiver. Since this bias voltage controls the current flowing in one or more i.f. stages, the simplest *S meter* comprises a meter of about 10 mA f.s.d. connected in the anode circuit of one of the a.g.c. controlled stages. This arrangement however, provides a reverse-reading meter with minimum current indicating maximum signal strength. To overcome this difficulty a more sensitive meter (usually 1 mA f.s.d.) is often connected in a bridge circuit with zero-setting adjustment, so that a decrease in anode current to one or more a.g.c. controlled stages produces an increase in current flowing through the meter. A simple

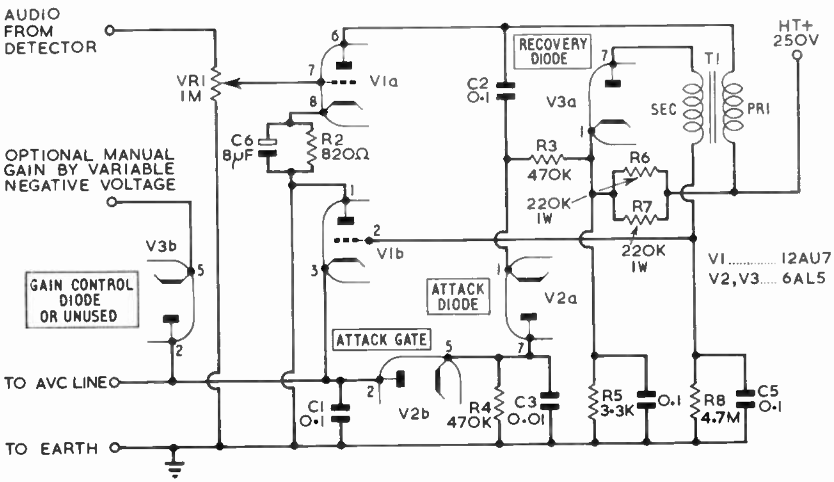


Fig. 4.44 (c). The audio derived "hang" circuit. V1 is a 12AU7 and V2, V3 are 6AL5 valves.

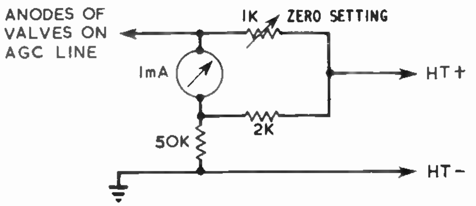


Fig. 4.45. Basic bridge circuit for a forward reading S meter. In some designs, variation of current to one or more of the screen grids of the a.g.c. controlled valves is used.

form of bridge circuit is shown in **Fig. 4.45** but many variations are found in practice, mainly intended to overcome the difficulty of obtaining a fairly linear calibration of what is essentially a logarithmic scale over a wide voltage range of the order of 100db (voltage ratio of 100,000 times).

There is considerable advantage in employing a separate valve to measure the potential developed across the a.g.c. line, and a recommended circuit using a double-triode valve is shown in **Fig. 4.46.** This is essentially a valve voltmeter and can be added as a small external unit to almost any existing receiver having an a.g.c. line. R1 is the zero adjustment to compensate for different levels of receiver noise. The input voltage taken from the receiver a.g.c. line is controlled by R2.

The calibration scale on a 200 μA f.s.d. meter provides adequate spread for the lower S units but also accommodates readings up to about 20db over S9. Mid-scale reading should be adjusted by R2 to represent about S8 or S9, and the scale calibration to suit the receiver concerned. The best relation between S units and decibel ratios has long been a matter of controversy, and to set up a scale accurately to a predetermined ratio such as one S point for 6db requires the use of a signal generator having an accurately calibrated attenuator; but unless the gain of the receiver is approximately equal on all bands such calibration will be of limited accuracy. For the average amateur, it is probably best to regard the calibration of an S meter in terms of purely arbitrary units, chosen to suit the actual equipment and the operator's own requirements. For instance, with the unit shown, S1 could be the minimum signal detectable above the noise level and S9 a very strong but non-local signal, with the scale divided evenly for intermediate strengths.

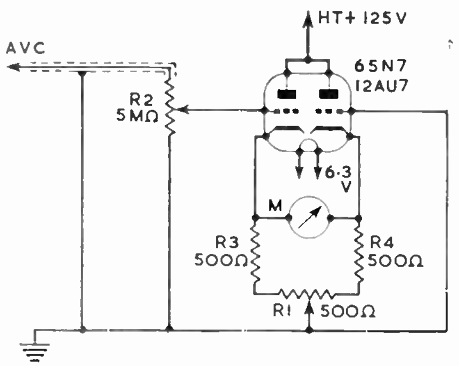


Fig. 4.46. Valve-voltmeter type of S meter. A suitable meter would be 200 μA full scale deflection. R3 and R4 should be 10 per cent tolerance.

A.F. STAGES

The a.f. output from a diode detector is usually of the order of 0·5–1 volt. A single a.f. amplifier will be adequate for headphone reception, but a further stage of power amplification will be needed to operate a loudspeaker. Since the signal-to-noise ratio is determined in the early stages of the receiver, there is little point in providing excessive a.f. amplification, though some reserve may be useful to overcome insertion losses if an audio filter (described later) is fitted. In practice, the basic circuit arrangements in this section of the receiver differ little from those found in broadcast receivers, except to provide for the connection of headphones in such a manner that the full output of the receiver is not applied across them, and the loudspeaker muted. For communications purposes there is no need to provide a wide a.f. range, indeed any high modulation frequencies will normally be removed by the selective i.f. stages. It is useful also to restrict bass response, and this can

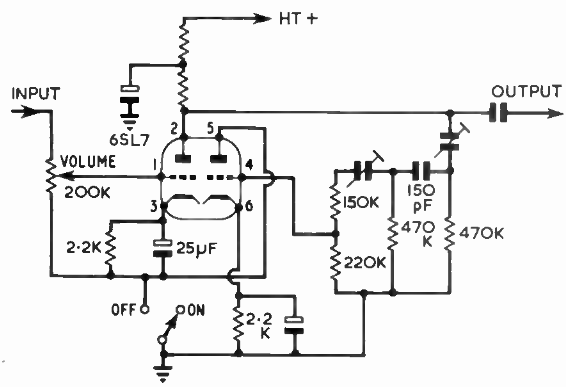


Fig. 4.47. Audio filter, using conventional components, which employs negative feedback to attenuate signals other than those over a narrow band at about 950 c/s.

be done by reducing the value of the interstage coupling capacitors from the conventional value of the order of 0·01 μF to about 0·001 μF or less; this will also help remove any traces of mains hum.

A triode-pentode can conveniently be used to provide both a stage of a.f. voltage amplification and power output: one example is the ECL86/6GW8.

The restriction of a.f. response can be carried to extreme limits for c.w. reception by peaking the response sharply between say, 800 and 1000 c/s, the frequency range selected by most operators for telegraphy reception. The design of a filter for a passband of only 100–200 c/s bandwidth on a.f. signal offers much less difficulty than one for operation at intermediate frequencies. An a.f. filter is effective in reducing interference to c.w. reception, but should not be regarded as a substitute for good i.f. selectivity for the following reasons: (a) it is unable to distinguish between the upper and lower sideband beat notes and so cannot in itself, provide true single-signal reception; (b) being placed so late in the receiver, it does nothing to reduce the susceptibility of the receiver to blocking. A further disadvantage of a sharply peaked a.f. filter is that many operators find it fatiguing to listen for any length of time to a single-tone note free of harmonics. Where a similar degree of selectivity has been achieved in the i.f. stages, a.f. harmonics will usually be reintroduced in subsequent stages so that the peaking is not so noticeable.

An inductance/capacitance tuned circuit can be used as an a.f. filter but the inductor requires a special winding technique if a high Q is to be achieved. Many amateurs use an item of surplus equipment (the type FL8 filter) which is tuned to about 1000 c/s.

By using frequency selective negative feedback and an extra valve it is possible to construct a sharply tuned filter requiring only preferred values of resistors and capacitors. **Fig. 4.47** shows one of the many possible arrangements. In the OFF position the first half of the double triode acts as an a.f. amplifier. With the filter ON the second triode section is switched in series with the cathode of the first section, and provides considerable negative feedback, attenuating all frequencies other than a small band at about 1000 c/s, the precise frequency being determined by the constants of the *RC* network between the anode of the first triode and the grid of the second triode.

The fatigue induced by tuned a.f. filters can be avoided by providing a sharp upper cut-off frequency, but with only a gradual attenuation curve for lower frequencies: this can be arranged by using a standard choke or a.f. transformer as the inductive element. A typical arrangement is shown in Chapter 16 (*Mobile Equipment*).

In the section on Q multipliers, it was explained that a sharply peaked filter can be used either to provide a narrow response curve or arranged so that it results in a sharp rejection notch. This principle can be applied equally at a.f. and forms the basis of a "selectoject" (select or reject) filter though this is now rarely used.

Noise Limiters

The h.f. range, particularly above about 15 Mc/s, is susceptible to electrical interference from car ignition systems and electrical appliances using commutation, vibrating contacts or any other mechanism whereby electric sparks, no matter how minute, are created. The signals radiated—or carried along electric wiring—from such appliances are in the form of high amplitude, short duration pulses covering a wide range of frequencies. In many urban districts, these interference signals set a limit to the usable sensitivity of a receiver.

Because the interference peaks, though high, are of extremely short duration, a considerable improvement can be

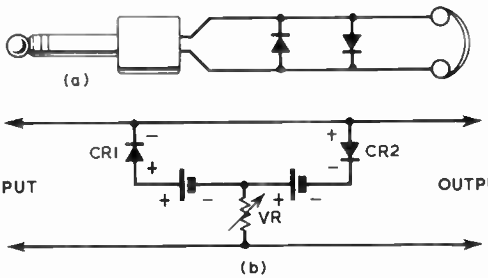


Fig. 4.48. Simple a.f. peak limiters. (a) Westector metal rectifiers can be used without back-biasing potentials to provide suitable limiting for headphone reception. (b) The effectiveness of this limiter with crystal or metal rectifiers is governed by the setting of the variable resistor.

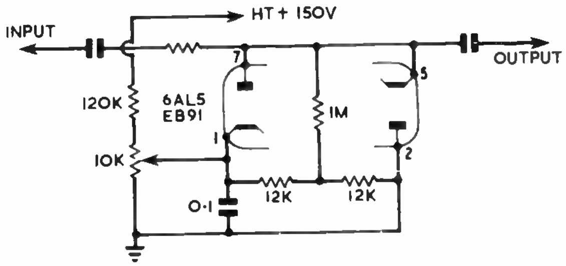


Fig. 4.49. Positive and negative peak limiter that can be built into a receiver. The degree of limiting is determined by the setting of the 10 K ohms potentiometer.

obtained by clipping or "slicing" all signals which are of appreciably greater amplitude than the desired signal. This can be done by relatively simple a.f. limiters as shown in **Figs. 4.48 and 4.49.** Alternatively the noise pulses can be used to provide a fast-acting biasing pulse used to reduce momentarily the receiver gain. During the period of the noise pulse, the output of the desired signal will also be reduced but this is of little practical consequence, since the

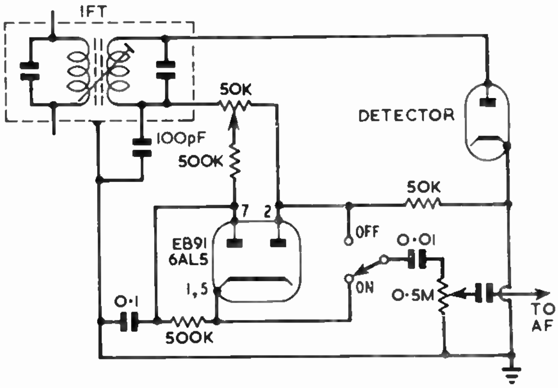


Fig. 4.50. This type of automatic noise limiter is most suited to a.m. telephony reception. The noise pulses provide fast-acting biasing pulses which cut off signal output during noise peaks.

ear is much less disturbed by "holes of silence" than by peaks of noise. A number of effective noise limiters of this type have been developed, and a typical circuit is shown in **Fig. 4.50.**

Unfortunately, since the noise pulses contain appreciable high frequency components, the effect of passing these pulses through highly selective i.f. stages will be to distort the pulses, and this will render noise limiting much less effective on communications receivers than on a wideband receiver such as those used for television reception. To overcome this difficulty, noise limiting systems have been developed which derive the cut-off or blanking bias from noise pulses which have not passed through the receiver i.f. stages but are received on a broad-band fixed tuned receiver. Such systems however are fairly complex and to date have been used only in conjunction with elaborate factory-built equipment.

The commonly used series and shunt limiter of the Dickert type, whilst very effective on a.m. signals when

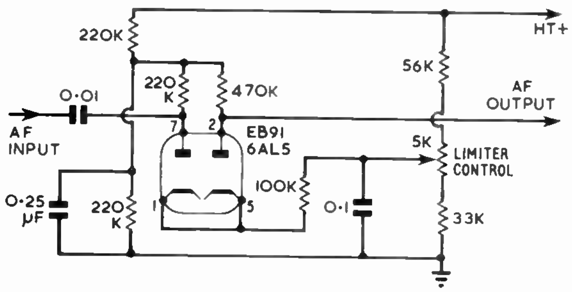


Fig. 4.51. Variable peak limiter suitable for all types of signals. This is an alternative arrangement of the type shown in Fig. 4.49.

following a diode signal detector, is not suitable for use on s.s.b. or c.w. nor to follow a detector of another type.

Separate demodulators are often employed for s.s.b./c.w. and a.m. and it is convenient to place a noise limiter in the audio input to the first a.f. amplifier. An arrangement which is known as the variable peak limiter, will permit the clipping of both negative and positive half cycles is shown in Fig. 4.51. This limiter is not an automatic following type but is manually adjusted by the panel mounted control. Correct operation of this control will be obtained by adjusting the values of the resistors on either side of the potentiometer so that all output is removed at the fully clockwise position.

A number of additional practical circuits for various forms of noise limiter can be found in *Technical Topics for the Radio Amateur* (RSGB) and in the *Amateur Radio Circuits Book* (RSGB).

POWER SUPPLIES

H.F. receivers for use in the home station are normally designed for operation from the a.c. mains supply using a double-wound transformer to provide 6·3 or 12·6 volts for the valve heaters, and full-wave rectification to give an h.t. line of 200–250 volts. The current ratings will depend on the number of stages and the valves used, and it will be most advantageous to have available reasonable reserves for operation of associated external units such as pre-amplifiers and converters. The rectifier valve and mains transformer should be most generously rated to avoid excessive heating with continuous operation, particularly in the tropics.

Although it is the practice in most commercial designs to accommodate the power pack on the receiver chassis, it represents a major source of heat and can produce stray magnetic fields around the transformer and smoothing choke which can be the cause of hum. The home constructor who may not wish to incorporate accurate thermal compensation of oscillator circuits or magnetic screening will often find it a decided advantage to place the power pack in a completely separate unit.

A great reduction of heat is possible by using silicon power diodes rather than a rectifier valve. Suitable silicon diodes are readily available and their use is recommended for all sets in which the power pack forms an integral part of the receiver. Where silicon diodes are used some precautions will usually be necessary to avoid the effects of mains over-voltage transients and switch-on inductive surges; this question is considered in greater detail in Chapter 17—*Power Supplies*.

Apart from those stages for which closely regulated h.t. supplies are required (oscillators, mixer screen supplies), the actual value of h.t. is not critical. Provided the power pack is run well within its capabilities, sufficient regulation will usually be provided by the load imposed by the receiver. An h.t. rail of 150 or 200 volts, instead of the more common 250 volts can materially reduce heat generated in the receiver.

Heater voltage is more critical; poor performance or short valve life can often be traced to under- or over-running the heaters. Constructors should ensure that the heater voltage—*measured at the valveholders*—is within 5 per cent of the nominal voltage (that is, within the range 6–6·6 volts for a nominal 6·3 volt supply).

The circuits described in this chapter are intended for use with double-wound mains transformers, which completely isolate the chassis from the supply mains, with heaters connected in parallel and operating at 6·3 or 12·6 volts. Some economy is possible by using what is known as a.c./d.c. technique—widely employed in broadcast receivers—in which the mains transformer is omitted and the valve heaters connected in series. This is necessary for operation from d.c. mains, and is sometimes used in competitively-priced, factory-built communications receivers. However, it requires one side of the supply mains to be directly connected to the chassis of the receiver, and stringent precautions are necessary to avoid risk of shock. It is strongly recommended that

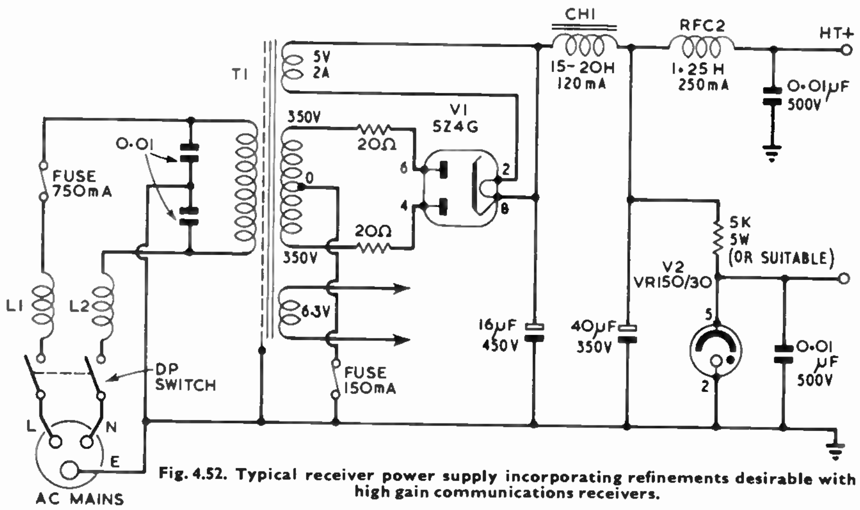


Fig. 4.52. Typical receiver power supply incorporating refinements desirable with high gain communications receivers.

a.c./d.c. techniques should not be used for home-constructed amateur receivers or for modified equipment.

A typical receiver power supply circuit using a rectifier valve is shown in **Fig. 4.52**. The mains filter connected on the input side of the mains transformer is useful in preventing electrical interference and r.f. currents from entering the receiver. R.F. chokes L1 and L2 must be capable of carrying the receiver mains current without heating (0·5 amp rating is usually suitable), and the filter capacitors should be rated to withstand 300 volts a.c. (see page 4.36). These components should preferably be mounted close to the mains transformer, or alternatively in a separate screened and earthed metal box. Protective fuses in the mains input circuit and also in the h.t. circuit are advisable. An indirectly heated rectifier valve is preferable to avoid h.t. being applied to the receiver before the valves have warmed up. Surge limiting 20 ohm resistors limit the peak current through the rectifier. The smoothing choke should be very adequately rated, and of the order of 15 to 20 Henrys. An average receiver will require about 80 to 100 mA (small receivers with headphone output much less) and it is useful to design the power pack to supply about 120 mA, leaving a reserve for the operation of external converters and other devices.

A stabilized voltage rail is required for the oscillators and this is normally obtained by means of a voltage regulator tube. Calculation of the value of the series voltage resistor to provide the necessary voltage drop while maintaining, in all operating conditions, the minimum current of the regulator tube, is discussed in Chapter 17—*Power Supplies*. To provide low operating temperatures this resistor should be very generously rated.

If, with a well-smoothed supply, there is still some residual mains hum, it may prove advantageous to use a heater supply balanced about earth. This can be done by means of a centre-tapped heater winding on the mains transformer (or by two equal low-value resistors across the winding with their junction connected to earth). All leads in the receiver carrying a.c. (heater, pilot lamp and a.c. primary wiring) should be twisted and routed away from grid circuits: a.c. loops should be avoided (see **Fig. 4.53**). Occasionally hum may be introduced by electronic conduction between cathode and heater; this can be overcome by putting a positive bias of about 50 volts on the heater line rather than connecting this to earth.

Standby Switching

When a communications receiver is located immediately adjacent to a transmitter, it will usually be severely overloaded when the transmitter is radiating. Most receivers therefore incorporate a STANDBY or SEND/RECEIVE switch to mute the receiver during transmission periods. In some

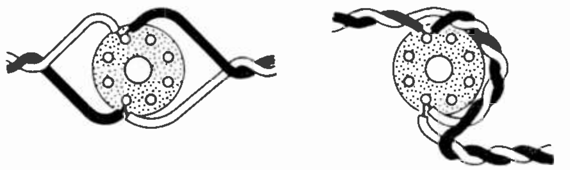


Fig. 4.53. Incorrect heater wiring can leave a loop under the valve holder which could increase hum level. The correct method of heater wiring is shown on the right.

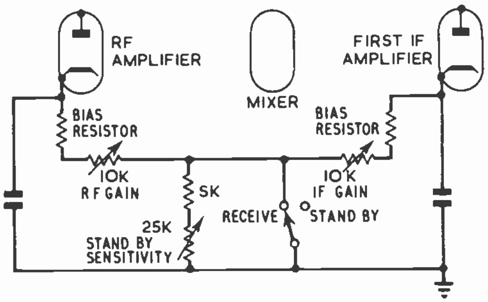


Fig. 4.54. Variable sensitivity standby control which facilitates monitoring of outgoing signals from the transmitter. In the standby position extra bias is applied to the r.f. and i.f. stages.

receivers this takes the form of a switch in the main h.t. line, though this may lead to cathode poisoning of the valves unless a resistor is connected across the switch to permit the valves to continue operating, but with reduced h.t. This system is not recommended practice for high performance receivers as it produces a thermal change in the operating conditions of the h.f. oscillator, causing possible drift during operation. A preferred method is to increase substantially the bias applied to the r.f. and sometimes the i.f. valves during transmission periods. This can be done by switching in an additional pre-set cathode bias resistor, set to the level at which the receiver will provide useful monitoring of the transmitter (**Fig. 4.54**). Alternatively, the additional bias can be supplied from a negative voltage supply or from an external negative potential, such as is provided for receiver muting in break-in control systems.

TRANSISTORIZED RECEIVERS

During recent years, high grade communications receivers based entirely on semiconductors have become available and are increasingly used in the commercial communications field. They offer the advantages of compact size, greater reliability, less change of characteristics with age, and low power consumption permitting optional use of batteries. Not only are semiconductors significantly more reliable than valves, but the associated low operating voltages and reduced heat make for increased reliability in the other components. The small size of semiconductor devices makes feasible more sophisticated circuitry; for example, the use of various forms of stabilized oscillators or frequency synthesizers, and complex a.g.c. techniques.

Such receivers can have a performance every bit as good as or better than the best valve designs; but high performance can be achieved only with careful and knowledgeable design, and in some respects calls for considerably greater complexity. For example one compact general-purpose commercial receiver (the Plessey PR155) uses no less than 160 semiconductor devices.

In general, while transistor communications receivers of medium performance can be achieved rather more easily and with fewer constructional problems than for a comparable valve receiver, really high grade transistor sets tend to present great difficulties, and must be considered an advanced project. In practice the majority of domestic amateur receivers continue to use valves, though there are many occasions when a " hybrid " approach—that is the mixing of some valves with transistors represents a currently attractive solution (for example, a built-in crystal calibrator or sideband selection crystal oscillators can be more conveniently designed around transistors). When mains supplies are not available as in mobile operation, the attractions of semiconductors are overwhelming. Where a unit is to be formed basically around semiconductors but valves are required for say the front-end, an attractive solution is to use the small Nuvistor valves.

The main problem with readily available conventional transistors is that of achieving a really wide dynamic range of the front-end and i.f. stages needed to handle both very weak and very strong signals without introducing various forms of cross-modulation and blocking. This is in part because there is no near equivalent bipolar transistor to the variable-mu pentode valve. This makes it difficult to obtain really effective automatic gain control and good mixer characteristics unless unipolar devices such as the various forms of field effect transistors (FETs) mentioned below are used.

Other problems include:

(a) Increased circuit loading due to lower input impedances.
(b) Feedback capacitances that may require neutralizing,
(c) The variation of semiconductor characteristics with temperature.
(d) The fact that, unless protection circuits are fitted, transistors can easily be damaged or destroyed by large input voltages due to static build-up on the aerial, or impulse interference from a nearby transmitter.
(e) The greater characteristics spread between devices bearing the same type number than is usual with valves.
(f) And the need to pay greater attention to accurate impedance matching through all stages of the receiver.

While this may seem a formidable list, in fact most of the problems are being overcome or avoided with current types of semiconductors. Silicon devices can withstand greater voltages and are less susceptible to temperature than germanium devices (though they tend to be a little noisier for equivalent structures). Devices intended for forward rather than reverse a.g.c. allow improved a.g.c. characteristics without introducing cross-modulation. Another development which is already affecting the design of the front-end of h.f. receivers and converters is the low-noise junction-gate field effect transistor. Various special forms known as IGFETs (insulated gate field effect transistors), MOSTs (metal oxide semiconductor transistors), and the more recent dual-gate IGFETs (combining the equivalent of two valves and thus suitable for use in cascode circuits) are also likely to have a profound effect on design.

These unipolar devices have high input impedances and characteristics which make them less susceptible to cross-modulation; in this respect they are nearly equivalent to variable-mu pentode valves. Because their characteristics are nearly square-law, they can act as extremely efficient low noise mixers. At the time this edition is being prepared, the prices of the various forms of field effect transistors are appreciably higher than more conventional bipolar transistors but even these prices have dropped to a level where the use of FETs should be seriously considered for the front-ends of receivers.

Front-end dynamic range with conventional transistors is a formidable problem, though by use of attenuators between the aerial and the first stage, or by additional local a.g.c. loops (sometimes used in conjunction with a voltage-controlled transistor stage) and the use of forward a.g.c. receivers can be designed to handle even wider dynamic ranges than are normal with valves.

Development in the semi-conductor field is very rapid and the next few years are likely to produce devices which will revolutionise receiver design. Prices of newly developed components are usually prohibitively high but fall rapidly as production techniques improve.

The Field Effect Transistor

As already mentioned, the unipolar field effect transistor and associated devices have emerged as a most important device for h.f. and v.h.f. receivers and converters. They offer substantial advantages over the more usual bipolar transistors in allowing r.f. stages and mixers to handle an appreciably wider range of signals without introducing cross-modulation. Also eliminated are those problems of transistor design which arise from the lower impedances of bipolar transistors, and their being voltage rather than current operated devices.

A unipolar device is simply one in which only one kind of majority carrier is involved (in contradiction to bipolar devices in which the basic action depends upon both majority and minority carriers in any given device).

Fig. 4.55 is a diagrammatic representation of the FET, from which it will be noted that the carriers flow in a narrow channel between source (s) and drain (d) with the flow determined by the field (that is the voltage) applied to the gate (g). The action is thus voltage controlled and is closely analogous to the valve, but since the carrier can be either holes or electrons the control voltage may be positively or negatively biased according to whether the device is a *p*-channel or *n*-channel device (rather like the difference between a *p-n-p* or *n-p-n* transistor). For those who think in terms of valve operation, the gate roughly corresponds to the grid, the source to the cathode and the drain to the anode.

In the application of bias potentials to the gate, the polarity of the biasing voltage will clearly depend upon whether a *p*-channel or *n*-channel device is being used; in the case of the IGFET there is a further difference between *depletion-mode* and *enhancement-mode* devices. An IGFET device having appreciable channel conductance with zero gate bias, and which is then reverse biased to reduce drain

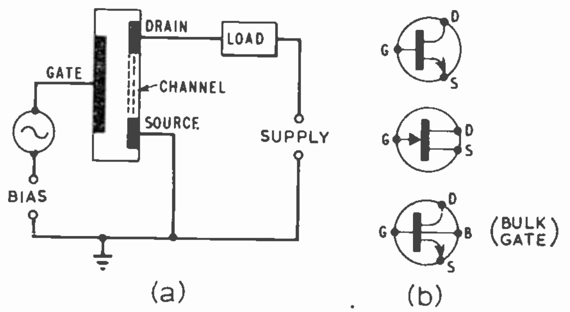


Fig. 4.55. (a) Diagrammatic representation of the h.f. junction FET in a common source amplifier. (b) Some of the varied symbols used for the FET. With *n*-channel devices the bias and supply polarities are the same as for valves.

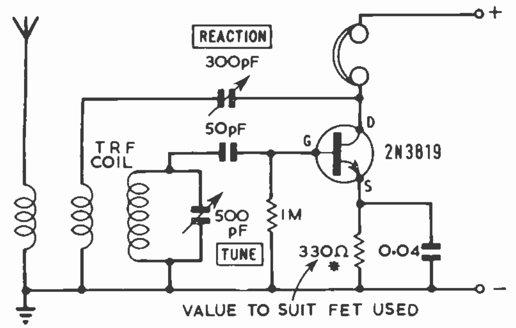


Fig. 4.56. A single-FET receiver.

current and hence amplification is termed a depletion-mode device. However IGFET devices are more commonly designed to utilize an enhancement mode, in which under zero bias there is little channel conductance but which can be forward biased to increase the control current.

Junction-gate FET's are always reverse-biased as in the depletion-mode IGFET.

Because of the extremely high input impedance of IGFET'S, considerable care must be taken when handling the devices to prevent the build-up of static charges which can puncture the insulated layer. Shorting wire or foil should be kept wrapped around leads of devices whenever possible when out of circuit, and always during storage. When the shorting wire is to be removed, hold the leads between the finger and thumb of one hand to provide a leakage path. Never allow a gate lead to float since there is then no bleed path for static electricity. Junction-gate FETs are less susceptible to this form of damage, but even so reasonable precautions should be taken.

The input impedance between gate and source is very high (extremely high in the case of the IGFET), and the devices can be used in circuits which are almost exact low-voltage equivalents of valve circuits. As an amplifier the FET can be used in the common-source configuration (similar to the grounded-cathode valve circuit) or as common-gate (similar to the grounded-grid), and there are some advantages in the use of common-gate amplifiers on v.h.f. and u.h.f.

The IGFET/MOST is, for practical purposes, similar to the h.f. junction-gate devices but have an additional insulated layer interposed between gate and channel, resulting in extremely high input impedances, but making it rather easy to damage a unit by static voltages up to the stage where it is safely wired into circuit having bleed resistors.

A problem with currently available FET devices is the wide production spread of characteristics, and this will often make it necessary to select the biasing resistor to suit an individual specimen; this however need present no great difficulties to the constructor of single units. For many stages, a technique is to include initially a 10K ohms potentiometer set at maximum resistance in the source lead in the place of the bias resistor, and then to apply power and reduce the source resistance gradually until the required drain current is obtained. The potentiometer is then removed, its value measured, and replaced by a fixed resistor of the closest corresponding value.

The slope of the FET characteristic depends upon the operating point, as in a variable-mu valve. Square-law characteristics mean that this slope varies linearly with the bias, gradually decreasing as the device goes towards its cut-off point (in this case called the *pinch-off* voltage) rather like the variable-mu pentode. This characteristic means that the device will have relatively little difference in noise performance when used as a mixer as it has as an amplifier, apart from its value in providing good a.g.c. These devices have lower susceptibility to temperature variation than bipolar transistors, a feature making them attractive as oscillators.

The relatively high input impedance implies that good pre-selector selectivity can be obtained, since the compromise between power transfer and Q necessary with bipolar transistors does not arise. It should be noted that although an FET amplifier is an appreciably higher impedance arrangement than one with bipolar transistors, this does not mean that the source impedance must always be very high. For example at 100 Mc/s the optimum input impedance for a common source 2N3823 junction-gate FET is about 1000 ohms, and in the common-mode configuration about 300 ohms.

Representative of junction-gate FETs suitable for h.f. receiver applications are the 2N3823 and lower-cost devices such as the 2N3819, 2N4223, 2N4224 and TIS34. Suitable h.f. amplifier arrangements are the neutrode (single neutralized device in common source configuration), the cascode (two devices or single dual-gate unit), and for u.h.f. the unneutralized common-gate configuration.

FET Applications

Although it would be possible to use FETs for many receiver stages, the additional cost means that in practice these devices are usually fitted only in the r.f. stage(s), mixer, and, less frequently, as oscillator or in i.f. stages.

Fig. 4.56 shows the simplest possible arrangement using a single junction-gate FET as a regenerative 0-V-0 type of receiver; this can give results comparable to an equivalent valve receiver which it so closely resembles, though such a receiver would be of limited use for amateur working under modern conditions.

An experimental front-end using two junction-gate FETs and a bipolar transistor oscillator is shown in **Fig. 4.57.** This circuit is based on a design by G3UMF intended to cover about 10 to 30 Mc/s and to provide an i.f. output on 1·6 Mc/s.

The square-law relationship between transconductance and drain current makes FETs and IGFETs attractive for such applications as product detectors and as balanced mixers.

A further problem with semiconductor front-ends is the need to provide protection from extremely strong signals from the station transmitter or static charges built up on the aerial or other forms of impluse inference. Crystal diodes, connected back-to-back across the first tuned circuit, can give useful protection; and silicon transistors are less susceptible than germanium types to damage from this cause.

Transistor R.F. Stage

The design of an h.f. receiver is always a compromise between minimum noise figure, minimum spurious responses and minimum cross-modulation or blocking. Low noise means high gain before the mixer; minimum spurious responses require a number of tuned circuits before the (first) mixer (or alternatively carefully designed wideband filters);

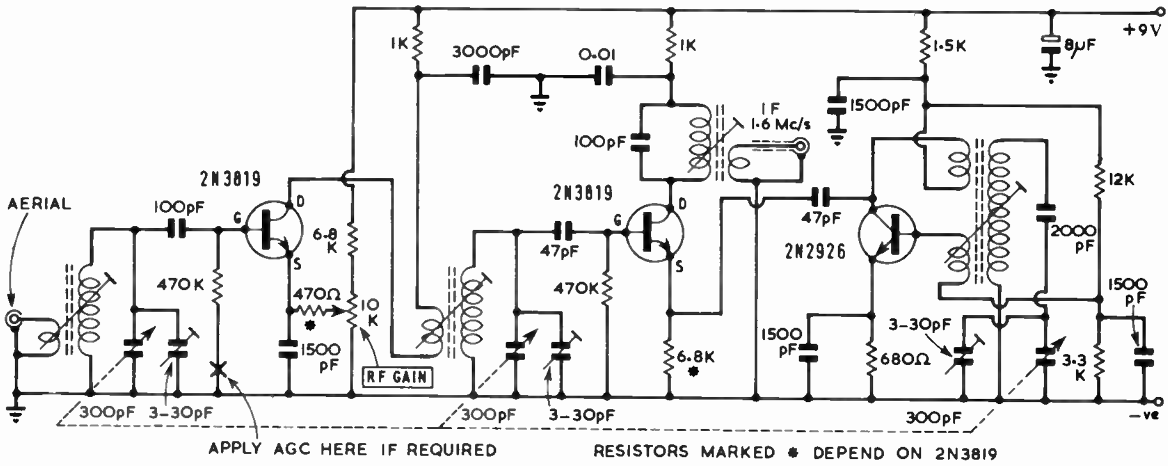


Fig. 4.57. Circuit of the G3UMF FET front end.

low cross-modulation, on the other hand, requires the least possible gain before the mixer.

Since external noise limits usable h.f. sensitivity, extremely low noise is much less important than on v.h.f., and a noise figure of say 6–10db is quite adequate. This can be achieved readily with modern transistors. In the absence of a variable-mu device, a stepped attenuator between aerial and the first transistor is important, unless this stage is protected by a local a.g.c. loop in which the control characteristic is derived from the signals present in this stage, rather than, as in conventional a.g.c. those present after the selectivity response has been shaped. Germanium devices have better noise performance than silicon devices, but sufficiently low noise can be achieved with both types.

Feedback capacitance between input and output signals may require neutralization, although this may not be needed if the device is operated well below its upper frequency limits. An alternative approach, which effectively eliminates the need for h.f. neutralization, is the transistor cascode circuit.

As already stated the provision of really good a.g.c. is more difficult with transistors than with valves; in practice both *reverse* and *forward-acting* a.g.c. techniques may be found.

Reverse a.g.c. is a term applied to a transistor amplifier which is gain-controlled by holding the collector voltage relatively constant and changing the bias current and hence collector current through the device. It is thus close to the technique used with valves, except that gain is current rather than voltage controlled. A disadvantage is that the linearity of the amplifier is reduced, increasing susceptibility to cross-modulation; for this reason use of the system is deprecated for high-performance receivers.

Forward a.g.c. is an increasingly popular system in which the collector-base (or collector-emitter) voltage is made to vary in accordance with collector current, the gain being directly dependent upon voltage across, and current through, the transistor. The usual technique is to include a resistor in the collector circuit so that the collector voltage falls with increasing emitter current.

Transistor Mixers

Any non-linear device can be used as a mixer, and a transistor can be operated as a mixer by applying an oscillator voltage to the emitter circuit and the incoming signal to the base circuit; alternatively both signals may be applied to the base circuit. Mixing results from the non-linearity between collector current and base-emitter voltage. It should be noted that the optimum oscillator voltage is much less than that required for a valve, about 0·1 volt peak. In effect the oscillator signal modulates the non-linear impedance of the emitter-base diode. Since the r.f. signal is also present the usual sum and difference frequencies are produced, and amplified by transistor action.

In the majority of small broadcast transistor receivers a single transistor functions as a combined mixer/oscillator using what is termed the *additive self-oscillating* arrangement. The transistor acts both as a commmon-emitter mixer and also as a common-base oscillator. To obtain ready oscillation when the receiver is first switched on, the bias is chosen so that the oscillator circuit is initially biased for class A operation; subsequently rectification of the oscillator voltage at the emitter provides additional bias current, driving the transistor into class B operation. This helps to stabilise the amplitude of oscillation, providing a more equal performance over the tuning range; in some sets a germanium diode is connected across the collector feedback coil to limit oscillator output and so prevent blocking oscillation.

While this type of circuit can be used in amateur-band receivers (particularly for the lower frequencies), there are a number of advantages, including reduction of oscillator pulling, in using a separate transistor as an oscillator.

Conversion gain of a transistor mixer is defined as:

$$\frac{\textit{I.F. power available at the mixer output}}{\textit{R.F. power available at the mixer input}}$$

Conversion gains of 20–25db are typical.

Transistors used as mixers should have efficient emitter-base diode characteristics; low emitter input capacitance; and good power gain at the intermediate frequency. Although a transistor can be used as a mixer with a separate oscillator up to about twice its alpha cut-off frequency

(f_α), it may be advisable, in order to keep the noise figure low, to use transistors with an alpha cut-off frequency some six or seven times the maximum frequency of oscillation. This implies the use of a transistor with an f_α in the region of 200 Mc/s for an h.f. receiver with a maximum operating frequency of 30 Mc/s.

Noise figures of 17db at 30 Mc/s are representative of a good low noise mixer. Since this is well above the optimum noise figure (6–10db) it shows the need for pre-mixer r.f. amplification. But in view of the possible poor cross-modulation and blocking characteristics of mixers it is always advisable to aim at the lowest possible mixer noise in order to reduce as far as possible the need for pre-mixer signal amplification. A measure of negative feedback applied to the mixer (for example the inclusion of a low-value, unbypassed emitter resistor) can improve cross-modulation characteristics but at some degradation of the noise characteristics.

For high performance it is thus important to limit the signal applied to the mixer, either by an aerial attenuator or efficient a.g.c. on the r.f. amplifier. A form of semiconductor mixer which appears to offer considerable advantage where the loss of conversion gain is not important is the diode ring balanced modulator using four crystal diodes; this can result in significantly improved intermodulation and blocking characteristics, and reduction of spurious signals. This technique can be applied to receivers throughout the m.f., h.f. and v.h.f. range; for optimum noise performance on h.f., it requires pre-mixer r.f. amplification of the order of 20db on the weakest signals. (See Fig. 4.16.)

The optimum approach is almost certainly the use of field effect devices which have near ideal square-law characteristics making them inherently more suitable for this application than either transistors or diodes.

Transistor Oscillators

In terms of frequency and amplitude stability, transistor oscillators vary from very bad to very good. Transistor characteristics are both temperature and frequency dependent, impedances tend to be much lower and also contain reactive components. To obtain really good stability these points must be taken into account.

With a tunable oscillator, the resonant circuit should be of high Q and arranged so that as little loading as possible is imposed by the transistor. This means that the tuned circuit is isolated as far as possible from the effects of changes in transistor characteristics caused by changes of supply voltage and/or temperature. The d.c. operating point of the

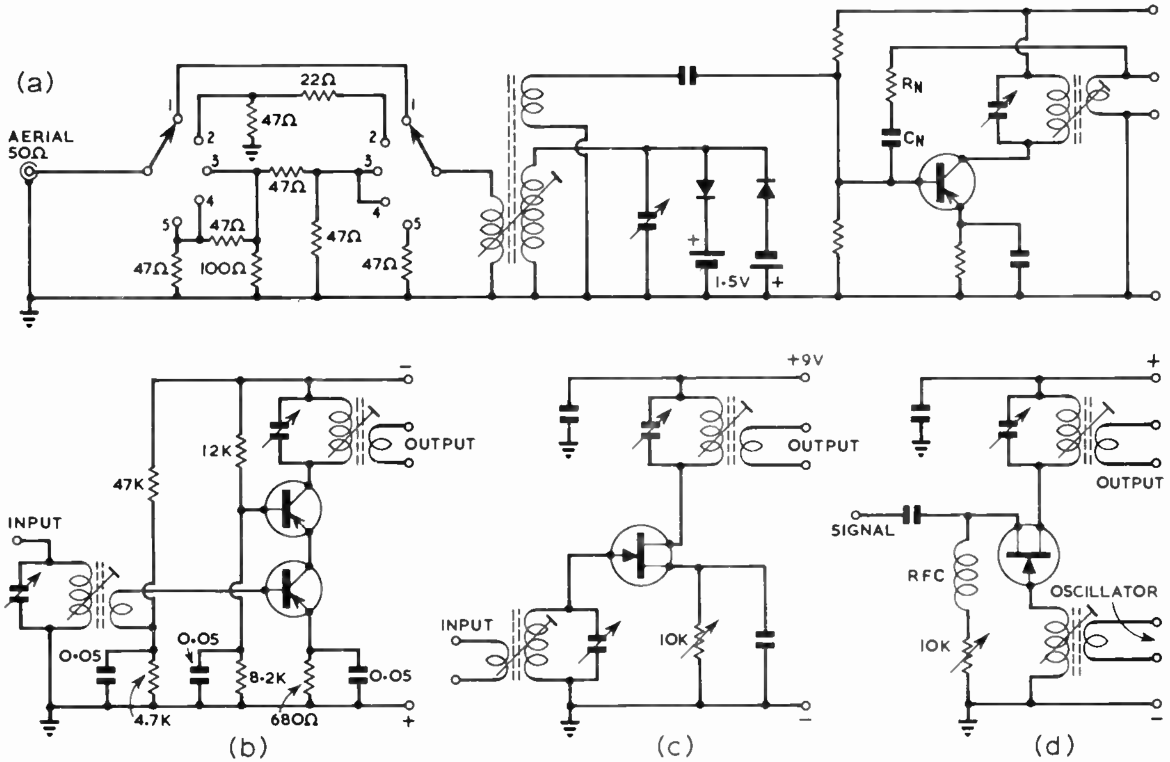


Fig. 4.58. (a) Simplified circuit of input r.f. amplifier circuits showing the type of precautions which may be advisable including a stepped attenuator, back-to-back biased crystal diodes to provide protection against local transmitters and (C_N, R_N) unilateralized "neutralization." (b) Cascode r.f. amplifier for operation up to 30 Mc/s. Typical transistors AF114, 2G414, Neutralization is unnecessary. (c, d) N-channel field effect transistors such as the 2N3823 offer appreciable advantages for low cross modulation, low noise r.f. amplifier and mixers. With very high input impedances, the circuits closely resemble those found in valve practice. A typical r.f. amplifier is shown at (c) and an FET mixer in (d). The various forms of FETs, MOSTs and IGFETs may prove valuable for the front-ends of communication receivers.

transistor should be stabilized, since any variation may affect the output frequency and waveform.

To overcome the problem of low impedances, the tuned circuit should be either of extremely high C, or the transistor connected to a tapping (or step-down winding), or via a capacitive potentiometer. Input impedance of the transistor can be raised by the use of the super-alpha pair (Darlington compound) connection with two transistors. Capacitive potentiometer and super-alpha techniques have both proved very successful in transmitter v.f.o.s. A degree of stabilization of the emitter current can be obtained by feeding the oscillator stage from a relatively high supply voltage through a high value resistor.

Some authorities suggest that a transistor with the highest possible f_α characteristics should always be used in a variable oscillator; but, provided that the f_α is high enough not to cause reduced output at the highest operating frequency, this precaution appears to have less effect than is sometimes suggested.

Many of the steps required to improve oscillator stability tend to reduce output and a buffer stage may be needed; this can conveniently be a common-collector circuit with a low impedance output taken from the emitter, with the signal injected into the emitter circuit of the mixer stage.

Transistor oscillators are considered in greater detail in Chapter 6—*H.F. Transmitters*.

Integrated Circuits

The increasing availability of low-cost monolithic semiconductor integrated circuits (SICs) is likely to have a growing influence on h.f. receiver design, not only in the use of devices intended specifically for linear circuits (r.f., i.f. and a.f. amplifiers) but also in the adaption of SICs originally intended primarily for digital (two-state) circuits but which can often be used as linear amplifiers at low signal levels, utilizing the differential amplifiers often incorporated. With the aid of integrated circuits, it will be possible to tackle complex equipment more cheaply, more quickly and with more certainty of achieving satisfactory results.

The integrated circuit can be considered as a group of active devices (transistors, diodes, field effect devices) and passive components (resistors and capacitors), and all necessary interconnections to form complete circuit networks packaged into one unit. In practice, these building blocks will normally be used with external conventional components for such applications as communications receivers.

In practice, two rather different forms of integrated circuit have been developed, plus some " hybrid " forms in which these techniques are combined together. The main groups are thin- and thick-film ICs and semiconductor ICs (SICs). In thin-film form, the connections, resistors and capacitors are deposited on insulated substrates such as ceramic or glass; the active devices are at present usually attached separately. Devices such as transistors which are

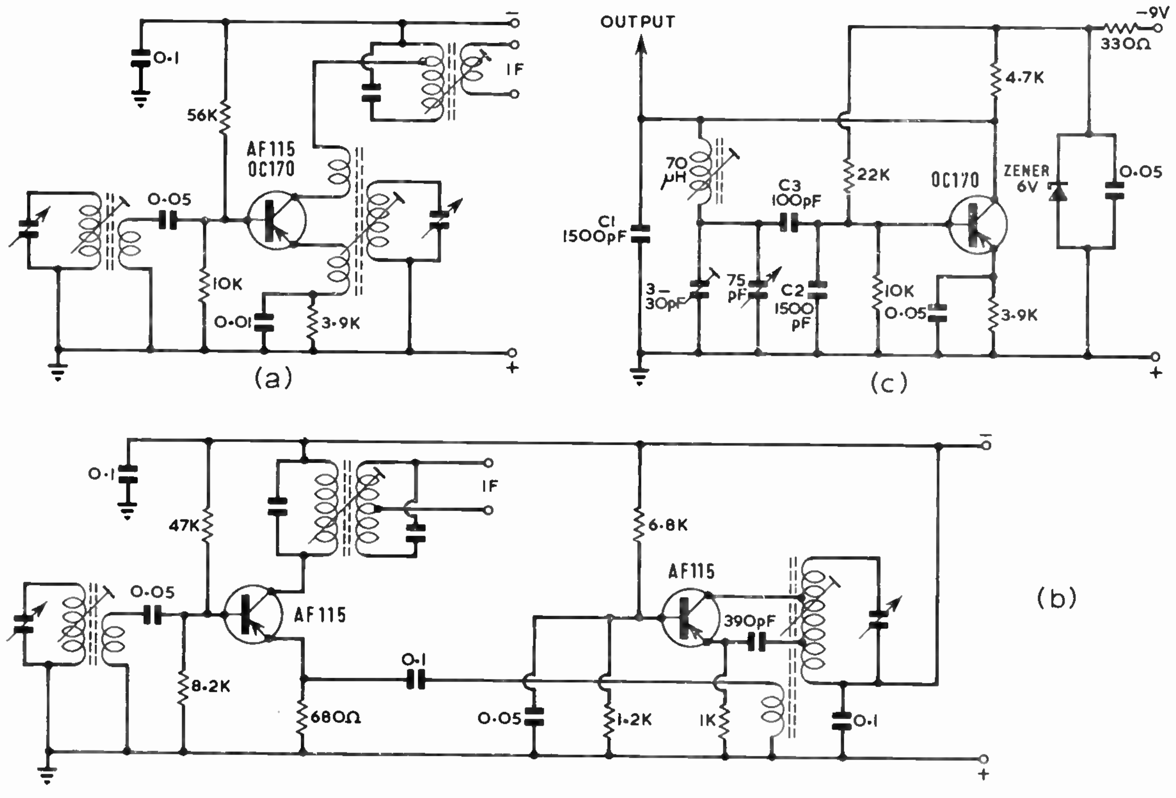


Fig. 4.59. Typical transistor mixer circuits. (a) Self-oscillating additive mixer. (b) On the higher frequency bands there are advantages in using separate mixer and oscillator transistors. (c) A Vackar oscillator, with values to provide a tuning range of 2.0 to 2.5 Mc/s, C1 and C2 should be polystyrene and C3 silvered mica.

manufactured separately and then attached to the substrate are often in a special form known as *flip chips*. In the case of thick-film units, the basic unit of the thickness of the film is of the order of 0·001 in. instead of the few hundred Ängstrom units of thin films.

The SICs fall into two main categories. In *multichip* or *chip* form, the individual components are produced on separate pieces of semiconductor material, and then assembled into a single package. On the other hand, the *monolithic* SIC (the type in which price reductions have been most significant) is entirely fashioned out of a single piece of semiconductor material, with all components, devices and interconnections and isolations formed within one tiny slice of the semiconductor. This is usually done by various diffusion planar processes; since planar techniques are easier with silicon than with other semiconductor materials, this is usually the material involved. The only additions are the leads brought out to connect the SIC into circuit.

The device is often referred to by the type of package in which it is enclosed: this is often the TO5 metal can originally developed for transistors the corresponding epoxy encapsulation, a rectangular *flat-pack*, or the *dual in-line* form in which the leads are bent to facilitate assembly on printed boards.

The SIC is not simply a module containing a large number of components but a complete circuit network having a finite number of leads brought out, so that there is a limited number of uses to which any particular device can be put. Since there are limitations to the component values which can be readily achieved, the use of integrated circuits tends to lead to a rather different approach to circuits in general.

For instance, a high value resistor or large capacitance takes up much more of the basic silicon slice area than does a transistor, so that it may often be preferable to use a whole series of transistors just to avoid the need for a single resistor. Typically it may be worth designing a circuit using five transistors just to eliminate the need for a single 47K ohm resistor, and capacitors impose even heavier penalties in slice area.

Most of the low-cost integrated circuits thus contain quite large numbers of transistors and diodes, but the resistor and capacitor values are kept as low as possible. There is no easy way of incorporating inductors within an SIC, and these are normally added in the external circuit, or eliminated altogether by, for example, the use of d.c.-coupled stages in i.f. amplifiers, following a single highly selective filter.

Many low-cost units can be readily used in multivibrator circuits, and this provides a satisfactory form of crystal calibrator arrangement. Another useful application is as a product detector or as an i.f. amplifier.

The incorporation of microelectronic devices in amateur receivers is unlikely to lead directly to substantial reductions in size or weight since many of the more bulky components cannot readily be integrated; there are also limits to equipment size imposed by the need for control knobs, dials and switches. They do however offer considerable scope for using more complex circuitry without increasing size, constructional complexity, power drain and cost. Their ultimate appeal will be on grounds of price, convenience in construction, and reliability in operation.

An important consideration is that more predictable and reproducible results become possible. For example, one of the inherent advantages of housing a complete circuit network in a metal TO5 can is better stability. Such common problems as regeneration in i.f. stages should be appreciably reduced.

For the amateur constructor, it will often be a matter of using a few integrated circuits for particular sections of a receiver, most of which will continue to be based on individual (discreet) components and semiconductors, but with ICs gradually taking over more and more of the total circuit.

CONSTRUCTION

The performance and reliability of an h.f. communcations receiver are governed as much by the choice of components and by the care taken in construction as by the basic design. Mechanical as well as electrical considerations must also receive attention.

The chassis layout is of particular importance in keeping r.f. and i.f. wiring short to minimize losses and stray couplings. In a multi-valve receiver, the layout should be arranged so that the main signal path follows logically through stages placed adjacent to one another. For most band-switched receivers, the placing of the coil pack, wavechange switching, ganged tuning capacitors and associated valves is decided first, usually occupying the main central part of the chassis; once this is fixed the layout of the remainder of the stages can be determined. Components and valves likely to run at high temperatures—particularly h.t. rectifiers and the a.f. output valve—should be placed as far as possible from the oscillator tuned circuits and other heat-sensitive components, or located in a separate unit. Adequate ventilation should be provided to keep operating temperatures as low as possible to reduce the time and extent of warming-up drift. The chassis material should be the heaviest gauge available in order to obtain maximum rigidity. If a completely rigid chassis is not obtainable, it may prove advisable either to reinforce the chassis or to mount the entire tuning section on a floating sub-chassis mounted on rubber grommets.

Screening is also important. In some instances for the second and third oscillators in multi-conversion receivers and the b.f.o., complete screening of the stage with an aluminium can may be advisable. For most other circuits, sufficient screening can be provided by an 18 s.w.g. (0·048 in.) brass or aluminium plate interposed between the circuits. Such screens must be rigidly fixed and make good electrical contact with the chassis. The chassis itself can form an effective screen by mounting the relevant components above and below it.

In high gain amplifying stages it is important that input and output circuits should be electrically isolated from each other to the maximum extent possible to avoid unwanted feedback which could cause instability or degrade the signal-to-noise ratio. Some valves, including r.f. pentodes, r.f. double triodes for cascode operation and certain modern frequency changers have internal screening; others—mainly the older octal-based types—are of metal construction. For many of the miniature valves external screening cans may be needed. Close-fitting aluminium screens raise the temperature of a valve appreciably and should preferably have a matt-black finish to assist heat radiation. Special heat-dissipating contact-type cans are also available. Where the valve-holder has a centre spigot (common on miniature types) it should be earthed to decrease stray coupling between grid and anode circuits. Sometimes such coupling can be reduced by fitting a small plate across the underside of the valve socket, though this will usually be necessary only at the higher frequencies. Above 14 Mc/s or so there is some

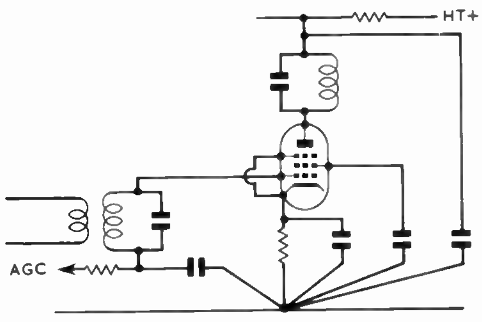


Fig. 4.60. To avoid regenerative or degenerative feedback in r.f. stages at the higher frequencies, it is advisable to bring all earth-return leads to a single point as close as possible to the valveholder.

advantage in using single point earth return wiring (see Fig. 4.60), all earthy connections associated with one particular stage being connected to chassis at a single point close to the valveholder. The object of this is to avoid circulating currents in the chassis, which may lead to regenerative or degenerative feedback outside the control of the user.

At an early stage in the design of the receiver a decision must be made between providing continuous coverage of all frequencies between say 1·8 and 30 Mc/s, possibly with the aid of a separate bandspread tuning control for use on the amateur bands, or alternatively covering only the amateur-bands (sometimes making provision for the reception of standard frequency transmissions on 2·5, 5, 10, 15 or 20 Mc/s). With a receiver covering amateur bands only there is much less difficulty in obtaining good tracking of the oscillator and signal frequency circuits, with consequent improvement in the efficiency and ease of alignment; it also facilitates the provision of the slow tuning rate highly desirable for c.w. and s.s.b. reception.

Coils and Wavechanging

The method adopted to alter the tuning coils when changing bands has considerable influence on the general construction of a receiver. The main methods in use are:

(a) Plug-in coils, either separate or combined into a single plug-in assembly for each band.

(b) All coils mounted on the chassis, with a Yaxley type wafer switch to select the coils required for each band.

(c) The coils mounted in a rotating turret assembly, so that although all coils are contained in the receiver only those required for one band are connected in circuit at one time.

(d) Individual front-ends for each band, usually constructed in the form of interchangeable sub-units.

Plug-in coils have the great merit of simplicity and high efficiency; they can be recommended for experimental work and for reducing the constructional complexity of receivers. The coils are usually wound on low-loss moulded formers,

TABLE 4.1

Guide to Coil Windings

Figures are given as a guide only and are based on a tuning capacitor with a maximum capacitance of 160 pF. The maximum frequency limit will depend largely on the value of stray capacitances, whilst such factors as closeness of turns, lengths of lead, position of dustcore (where used) will materially affect the frequency coverage. The reaction winding for t.r.f. receivers should be close to the lower end of the main winding. The aerial coupling coil, if of low impedance, should be wound over the earthy end of the main winding. If of medium impedance or for intervalve use, the coupling coil should be spaced a little way from the lower end of the main winding. Where an h.t. potential exists between windings, care should be taken to see that insulation is adequate. Small departures from the quoted wire gauge will not make any substantial difference. Generally, reaction and coupling windings can be of moderately fine wire.

The number of turns for intermediate ranges can be judged from the figures given.

Diameter of former	Number of turns				S.W.G. (main winding)	Approximate frequency range (Mc/s)		Remarks
	Main tuned winding	Low impedance aerial	Medium impedance aerial	Reaction		Minimum	Maximum	
1½ in. ribbed air-core	3	1	2	2	20	13·5	31	Turns spaced two wire diameters
	5	1	2 or 3	2 or 3	20	11·5	23	Turns spaced one wire diameter
	9	2	4	3	22	6·5	14	Slight spacing
	17	3	5	4	24	3·4	6·8	Close wound
	42	6	10	10	30	1·6	3·3	Close wound
⅞ in. ribbed air-core	8	2	3	3	24	16	30	Close wound
	18	3	5	5	24/26	7	16	Close wound
	40	6	10	10	30	3·5	8	Close wound
⅜ in. dust-iron core	8	2	3	3	26	13·5	31	Close wound
	14	3	5	4	28	7·0	15	Close wound
	26	5	8	6	30	3·2	7	Close wound
	40	6	10	9	32	1·6	3·6	Slightly pile wound

the bases of which carry pins designed to make good electrical contact with sockets mounted on the receiver chassis. The coils thus occupy only a small space on the chassis and this permits short r.f. wiring. The main disadvantages are a degree of operating inconvenience, and the difficulty of varying the inductance for alignment purposes, since plug-in formers with variable cores are not always available.

In those cases where the coils are mounted on the chassis, it is usual to wind them on small formers with variable brass or ferrite cores. The wavechange switch should have sufficient contacts to allow coils not in use to be short-circuited, otherwise unwanted resonances and absorption effects may occur. The coils are often mounted below the chassis, either directly underneath or to one side of the ganged tuning capacitor. The switch should be mounted close to the appropriate valve sockets to keep wiring short; similarly, the highest frequency coils should be placed nearest to the switch. For low losses on the highest frequencies, the wavechange switch insulation should be ceramic (particularly important where a single wafer carries sets of contacts at both earth and h.t. positive); otherwise, since wafer losses form only a small proportion of the whole, Paxolin type insulation will normally prove reasonably satisfactory. The switch action should be positive and the contacts preferably kept coated with silicone lubricant to prevent oxidization. Electrolube No. 1 (Green) is a useful proprietary lubricant for switches.

Turret coil assemblies are electrically very efficient but mechanically difficult to arrange. Stray capacitance and inductive coupling can be kept low, and the main receiver wiring simplified. The various coils are usually mounted on a framework which is rotated by the action of the wavechange knob. Sets of contacts fitted along strips holding the coils press against spring contacts connected to the main circuit. Television turret tuners have been successfully adapted for this purpose.

The interchangeable front-end or converter unit for each band is particularly well suited for use with a receiver having a tunable first i.f. The unit generally comprises a broad-band r.f. stage, first mixer and crystal-controlled h.f. oscillator. This system avoids the difficulties of switching h.f. circuits and allows each unit to be designed for optimum performance on the particular band concerned. It is however, a relatively expensive form of construction, amounting to a single-range receiver with a series of fixed-tuned converters. Some typical designs for the h.f. bands are described in Chapter 16.

Table 4.1 provides a guide to typical coil windings. For use with ganged circuits, care should be taken to wind coils as similar as possible; coils with adjustable cores permit accurate matching of inductances. Single-layer coils for the h.f. bands can easily be wound by hand, ready-made coils are also available.

Tuning Mechanisms

The tuning drive arrangements can make or mar the performance of a receiver, particularly on s.s.b. and c.w. signals. Points to consider are smooth continuous action and absence of backlash in the reduction gearing. Some operators find a moderately heavy flywheel action of assistance.

Assuming a total bandwidth of the order of 500 kc/s and a 180° tuning span, a reduction ratio of 100 : 1 would be needed to give a tuning rate of about 10 kc/s per revolution. This would generally be considered a good tuning rate, but where the receiver is intended primarily for c.w. or s.s.b. reception it may be advantageous to reduce the tuning rate still further, to around 5 kc/s or even 2·5 kc/s per revolution. With very low tuning rates it will be found advisable to fit a small handle to the tuning knob to facilitate tuning from one end of a band to the other. With receivers having a fairly high tuning rate (on many well-known models this may exceed 100 kc/s per revolution and considerably more on the 28 Mc/s band) it is sometimes possible to fit a further reduction drive on to the existing tuning spindle.

The knob with which tuning is normally done (the bandspread knob on a general coverage receiver) should have a fairly large diameter (at least 2 in.), to facilitate careful setting and should be mounted in the most convenient position for the operator. A very suitable drive mechanism for incorporation in a home-built receiver is the Eddystone type 898.

Permeability Tuning

While it is the usual practice to tune h.f. circuits with a variable capacitor, it is equally possible to do this by varying the inductance by means of sliding cores. This system, known as permeability tuning, offers considerable advantages in the maintenance of correct L/C ratios and permits closer aerial input coupling. Since the screw principle can be adopted to alter the position of the cores it makes possible almost unlimited mechanical bandspread without the problems of gear reduction drives. Although permeability tuning has been most successfully used for many years by one major American company, the mechanical and electrical complications involved in band-switched receivers have resulted in little progress in this field by amateur constructors. There is considerable scope for further development of permeability tuning systems, particularly for covering single wave ranges, as for example in receivers with fixed tuned front-ends and tunable first i.f. ranges.

Permeability tuned oscillators can be significantly more stable than those using capacity tuned circuits.

Fixed Capacitors

Several different types of fixed capacitor are found in receivers, and it is important that the constructor should be aware of the uses and limitations of the various types. At one time capacitors fell conveniently into three main categories: paper capacitors for a.f. applications, mica capacitors for r.f. circuits and electrolytic capacitors for smoothing.

While paper capacitors are still used for almost all a.f. circuits, several different forms of container are now available. The conventional waxed cardboard paper tubulars are still widely used, but should be avoided for any position where a very high insulation resistance is necessary. After a few years' use—and less in the tropics—the d.c. resistance of the cardboard container may amount to no more than about 5 Megohms. If used for inter-valve coupling this may result in a positive bias being applied to the grid of the succeeding valve. A low insulation resistance should also be avoided in a.g.c. circuits. To reduce insulation leakage many different types of containers have been introduced, including better forms of waxed-cardboard tubing and metal tubes with insulation seals at the ends.

Various forms of plastic-film or polyester capacitors have been developed and are being increasingly used. They have very low dielectric loss, a much higher insulation resistance than paper capacitors and have also proved much more reliable in use.

Quite high a.c. voltage peaks may occur in a.f. stages and capacitors subjected to these voltages should be rated to withstand the peaks plus any direct voltage which may be across them. Capacitors subjected to continuous a.c. stress must always be rated specifically for a.c. working since an a.c. working voltage of 300 volts is roughly equivalent to a 1000 volt d.c. rating. For such applications, petroleum jelly or liquid impregnants are better than wax.

When a fixed capacitor is used as part of a tuned circuit, for example to pad the oscillator circuit, it is essential to use a type of capacitor which introduces very little loss. Moulded mica capacitors can be used but the smaller size of silvered mica types have made these the normal choice. The silver coating is attached firmly to each side of the mica plates and the capacitance remains stable over very long periods; it is also possible to obtain this type to close tolerance.

Ceramic capacitors have taken over many of the tasks for which mica capacitors were formerly used, except where a very high order of stability is necessary. The various high-permittivity types are suitable and economical for r.f. or i.f. decoupling and similar purposes. Modern disc type ceramic capacitors are excellent for r.f. bypass applications, combining high capacitance in small space. Because they can often be mounted close to the valveholders or other main components, lead inductance can be kept extremely low.

With low-permittivity ceramic capacitors, use can be made of their sensitivity to temperature variations in order to provide compensation for changes in other component values as the receiver warms up. By using a capacitor with a negative temperature coefficient such as N750 (meaning 750 parts/million/degree Centigrade) to form a small part of the oscillator tuned circuit capacitance, frequency drift can be much reduced. It will usually be necessary to find the correct proportion of the tuned circuit capacitance which should be of the N750 type by experiment. A useful though somewhat expensive component for temperature compensation is the "Tempa-trimmer" made by Oxley which has a variable coefficient but fixed capacitance.

Electrolytic capacitors are valuable where high capacitances (above about 1 μF) are required. They are relatively inefficient for r.f. applications, and have a limited shelf-life. Care must be taken not to exceed the voltage ratings. Their main use is for ripple reduction in power supplies and for a.f. decoupling.

Where a capacitor forms part of a tuned circuit, close tolerance types—to about 1 per cent—may be needed. For most decoupling and interstage coupling applications, 10 or 20 per cent tolerance types may be fitted.

Diode Capacitors

A reverse-biased junction diode (and to a lesser extent point-contact diodes) has a capacitance which varies with a change in the reverse voltage. This characteristic can be used to provide what in effect is a variable capacitor tuned by adjusting the d.c. bias potential applied across it, if necessary from a physically remote position. Silicon junction diodes developed for use as capacitors are generally termed *varactors*, or by various trade names such as 'Varicaps.' An important application of such devices is for microwave parametric amplifiers and harmonic generators but they also find uses in h.f. receivers.

A representative varactor diode (Hughes HC7005) may have a change of capacitance of from 24 pF to 81 pF as the reverse bias is reduced from 21 to 1 volts. Although the capacitance continues to increase below this figure, the Q factor tends to fall and temperature coefficient degrades. The Q of such devices is considerably below that of conventional capacitors but is sufficiently high to allow their direct use in various tuning applications in h.f. receivers; the relatively low Q allows ganged tuning without calling for extremely close tolerances in capacitance characteristics. In practice, however, the true varactor diode is a relatively expensive device and use in amateur equipment tends to be confined to single tuned circuits or bandspread tuning of an oscillator where the facility of a panel mounted potentiometer is required.

Diode Switching

An increasingly important application of semiconductor diodes is to act as switches in receiver and transmitter applications. The switching properties of diodes have long been used in specific circuit configurations such as diode ring and bridge modulators/mixers.

Diode switches offer a number of advantages over mechanical ones, particularly for switching at r.f. There is no contact tarnishing or other forms of dirt, switching is almost instantaneous, switch capacitance may be as low as 1 pF, the diode can be mounted anywhere and not confined to a shaft, and the contacts do not arc.

Disadvantages are that a diode switch in the "open" state still imposes a finite resistance and thus cannot be used in cir-

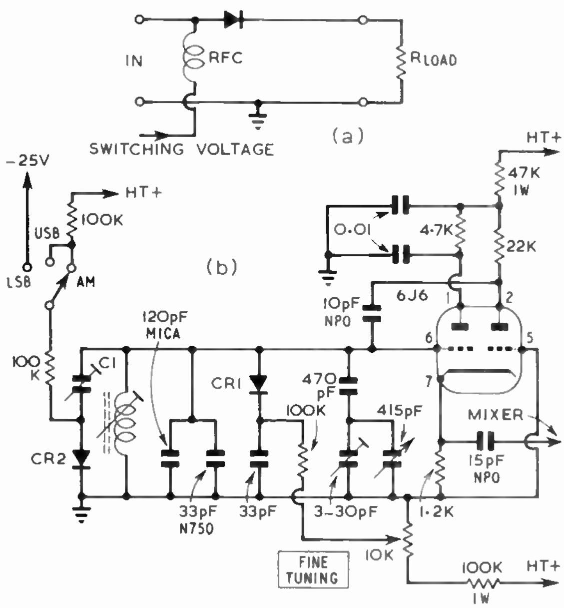


Fig. 4.61 (a). Basic diode switching circuit.

Fig. 4.61 (b). A 2-3 Mc/s v.f.o. for a multiple conversion receiver illustrating both the use of diode switching (CR2) and the use of a germanium crystal diode (CR1) for fine tuning.

cuits demanding extremely low loss, a power source is required to bias the diode and the switching currents must be sufficient to ensure a low enough differential resistance in the closed state.

Diode switching circuits are designed so that either a forward bias (" on ") or reverse bias (" off ") is applied across the diode. This bias potential is usually connected through chokes or resistors to isolate the d.c. circuit from the circuit being switched: see **Fig. 4.61**. The diode characteristics required are: low capacitance, high reverse resistance and low forward resistance. In some applications the p.i.v. rating of the diode must be considered. Point-contact germanium diodes such as the OA81 are often used.

Resistors

Consideration must always be given to the power which will be dissipated in a resistor, in order that a component of ample wattage rating is fitted. When little or no current flows through a resistor it is permissible to fit the miniature ¼-watt types, though some constructors prefer to keep to a minimum rating of ½ watt. Where appreciable d.c. flows through a resistor (cathode bias, screen feeds, anode load, etc.), the power dissipated should be calculated ($I^2 \times R$, where I is in amperes and R in ohms) but no attempt should be made to fit a resistor likely to run near its wattage rating—always fit a generously rated component for reliable operation.

A resistor operating near its wattage limit will run warm and successive heat cycles will tend to cause its resistance gradually to increase. Where the wattage dissipated exceeds about 1 watt it will be advisable to fit wire-wound resistors; otherwise the carbon composition type is generally suitable. though for a few applications where long-term stability is required cracked carbon types—available to tolerances of 1 per cent—may be preferable. Receiver resistors can usually be of 20 per cent tolerance, though for screen feeds and bias resistors it is advisable to fit 10 per cent types.

SUPERHET TRACKING AND ALIGNMENT

The superhet principle, as noted earlier, depends on the provision of a locally generated signal differing from the incoming signal frequency by an amount equal to the intermediate frequency. This local signal may be placed either higher or lower in frequency to the incoming signal, though in practice the oscillator usually operates on the higher frequency side of the signal frequency to avoid difficulties from harmonics.

In simple superhet receivers, the tuning of the signal and local oscillator circuits can be carried out independently using separate variable capacitors. Clearly, this is inconvenient as the operator must continuously adjust two or more knobs when tuning across a waveband. It is usual therefore to fit a ganged variable capacitor with from two to four sections to tune both the oscillator and signal frequency circuits. From this arises the problem of so arranging the tuned circuits that the oscillator and signal frequency circuits *track* together (keep accurately in step) so that for all settings of the control knob the frequency difference between them stays nearly the same. For example, suppose the band 5000–8000 kc/s is to be covered on one waveband with an i.f. of 470 kc/s, the oscillator tracking on the high frequency side of the signal. Then it will be necessary for the oscillator to operate on 5470 kc/s when the set is tuned to 5000 kc/s and 8470 kc/s when tuned to 8000 kc/s. Furthermore the oscillator must track throughout the tuning span; so that, for example, the oscillator is on 6470 kc/s and 7470 kc/s when the set is tuned to 6000 kc/s and 7000 kc/s respectively. Unfortunately, because the resonant frequency of a tuned circuit depends on the square root of the product of the inductance and capacitance, it is not possible simply to reduce one of these factors in order to shift the oscillator tuning span to 5470–8470 kc/s. Both the L and C components must be reduced in the correct proportions. The inductance can be reduced by using fewer turns on the oscillator coil. The capacitance is usually reduced by including in series with the variable tuning capacitor a fixed or variable capacitor known as the *padding* capacitor. If there is no means of adjusting the coil it is usual to make the padding capacitor variable but with a coil having an adjustable core it is more common to use a fixed capacitor of fairly close tolerance. Typical values for padding capacitors in general coverage receivers are given in **Table 4.2.** With amateur-bands only receivers, it is generally unnecessary to use padders. The padding or coil variation has most influence on the lower frequency end of the tuning span.

To permit ready adjustment of the circuits on the higher frequency end of the tuning range, it is usual to connect small pre-set variable capacitors across the coils, known as *trimmers*. Typical arrangements are shown in Fig. 1.66 in Chapter 1.

Unfortunately, even with full padding and trimming adjustments, it is not possible to obtain exact tracking of the oscillator with the signal frequency circuits over the entire sweep of a ganged tuning control. The best compromise provides three frequencies—one towards each end of the tuning span and one fairly central position—where the signal frequency circuits will be precisely the i.f. away from the oscillator frequency, with some slight error having to be tolerated at all other positions. For instance, in the above example the circuits would be accurately trimmed at around 7750 kc/s and padded at about 5250 kc/s, it would then be found that the third point of accurate alignment would be in the region of 6250 kc/s.

These unavoidable alignment errors are small but require some consideration for really high performance receivers. The errors can be minimized by keeping the total tuning span of each waveband as narrow as possible; the situation will be

TABLE 4.2

Guide to Values of Padding Capacitors for General Coverage Receivers

Signal frequency range	Approximate maximum capacity of tuning capacitor	Intermediate frequency	Value of oscillator padder
13 to 30 Mc/s 6 to 13 Mc/s 3 to 6 Mc/s 1·5 to 3 Mc/s	200 pF to 220 pF	450/465 kc/s	3000 pF 2000 pF 1425 pF 650 pF
10 to 30 Mc/s 3·5 to 10 Mc/s 1·4 to 4 Mc/s	360 pF	450/465 kc/s	3625 pF 2825 pF 1050 pF
12·5 to 32 Mc/s 4·5 to 12·5 Mc/s 1·7 to 4·5 Mc/s	360 pF	1600/1620 kc/s	2100 pF 960 pF 380 pF

improved if the amateur bands coincide with the accurately tuned portions. Because of this, despite the added complications in the coil assembly, it is preferable that a general coverage receiver should have as many wavebands as possible. In receivers which tune the amateur bands only the alignment errors can be kept to a very low figure. Furthermore, it is possible in such sets to overcome the whole problem of tracking by making the signal frequency circuits sufficiently broad-band to provide adequate response at least over the higher frequency bands without retuning, this allows the entire tuning to be carried out on a single small variable capacitor in the oscillator circuit or by permeability tuning this circuit only.

For general coverage receivers, the calculation of the correct coil windings to permit accurate tracking is difficult. This is mainly because the average constructor is seldom able to measure exactly the inductance of hand-wound coils or the various stray wiring capacitances which play an important role in determining the tuning range. Although surprisingly good results can be achieved by intelligent guesswork, there is little doubt that the difficulties encountered in this section of superhet receiver design have been the main reason why so many amateurs use factory-built sets. One solution is the use of the complete tuning units or coil packs, available from a number of manufacturers; another is to follow published designs in which full coil winding data is given. Yet another approach has already been suggested, this is the construction of amateur bands only models with broadly tuned signal frequency circuits or with signal-frequency and oscillator circuits tuned independently (this is much less inconvenient than on a general coverage model, as the adjustment of the signal frequency control can generally be limited to a final touching up on a signal). From almost all viewpoints, the amateur-bands-only approach can be recommended.

A most useful instrument for the preliminary adjustment of the tuned circuits of a receiver is the grid dip oscillator (see Chapter 19—*Measurements*) which permits rough checks to be made on the tuned circuits of a receiver during actual construction.

Spurious Responses

One of the most important criteria of a superhet receiver is the presence, or absence, of spurious signals. These spurious responses take three main forms:

(a) External signals which cannot be tuned out.
(b) External signals heard on frequencies other than those on which they are operating.
(c) Continuous carriers heard in the tuning range of the receiver but which are caused by internal oscillators.

Type (a) is almost invariably due to breakthrough of signals into the i.f. stages, though in a double superhet having a tunable first i.f. such breakthrough will provide tunable signals of type (b). I.F. breakthrough is generally caused by insufficient pre-mixer selectivity, and in this form is usually troublesome only on receivers having no r.f. stage. It can be reduced by series- or parallel-tuned circuits at the i.f. connected as wave-traps to reject the interfering signals. A more serious form of i.f. breakthrough is direct pick-up of signals in the i.f. stages, and this tends to be more likely with receivers having i.f. of 1600 kc/s or above. More efficient screening of the i.f. stages may be needed, with care taken to minimize stray coupling to external wires such as power leads, aerial, etc. In some cases, where particularly strong local signals are involved, it may be necessary to change the i.f. slightly. Breakthrough of signals can sometimes be difficult to eradicate where a converter is used ahead of a main receiver which picks up signals in the r.f. stages or in the interconnection leads. (Additional advice on preventing breakthrough when using a converter is given in Chapter 5—*V.H.F. Receivers*.) It should be noted that signals breaking through need not always be at the i.f.; very strong signals at half the i.f. can occasionally prove troublesome, the non-linearity of the mixer causing spurious signals to appear.

The most common form of type (b) interference is second channel or image interference. The susceptibility of a receiver to this form of interference is a function of the intermediate frequency used and the degree of pre-mixer selectivity. Where less than two r.f. stages are used, an important consideration is to maintain the Q of the input circuits. Image response is, however, only one of the many causes of this type of interference. The harmonics of the h.f. oscillator may beat against strong incoming signals which may themselves act as the h.f. oscillator and beat against other strong signals. When considering extremely strong signals such as those from a nearby transmitter, there are many possible spurious responses arising from combinations of the fundamental and harmonics of both the local station and the receiver's h.f. oscillator, many with associated image responses. In practice however, except for the station transmitter, spurious responses of these types can be reduced to insignificance by good pre-mixer selectivity and a high i.f., provided that in so doing cross-modulation effects are not introduced.

Interference of type (c), generally termed *birdies*, is of great importance in double superhets where, unless extreme care is taken, harmonics of the second oscillator—or image responses to these—are likely to appear in one or more wave-ranges of the receiver. For general coverage receivers, the only solution is most careful design and screening of the second oscillator to reduce harmonic generation. With amateur bands only receivers, particularly those using a tunable first i.f., careful analysis of possible harmonic interference can be useful in determining a suitable first i.f.

Harmonics of the b.f.o./c.i.o. can cause similar spurious responses, but because of the lower frequencies involved, these can usually be reduced to an insignificant level by screening.

Because of the difficulty of eliminating all spurious responses in double and triple-conversion receivers, some designers believe that the ideal amateur receiver would be a single-conversion superhet with a high intermediate frequency—but this would depend upon the development of high i.f. filters with a degree of selectivity at present available only in receivers having a low i.f. Crystal filters meeting such specifications are becoming available.

Smooth Regeneration

To obtain maximum benefit from regenerative detectors both for superhets and for " straight " receivers, it is essential to obtain smooth control as the stage goes in and out of oscillation. In some cases it may be found that the valve tends to plop in and out of oscillation, and the control cannot be advanced right up to the threshold of oscillation. The following are among the main causes of " ploppy " regeneration: too many turns on the feedback coil (this should have the minimum possible number of turns coupled tightly to the earthy end of the grid coil); the value of grid leak resistor may be incorrect (try the effect of different values, usually in a higher value direction); the anode and/or screen voltages may be too high, or derived from too high an impedance source (since the valve current conditions change during oscillation, screen voltages should be derived from a potentiometer network and high series feed resistors avoided); with straight receivers having the detector stage coupled directly to the aerial, it may be necessary to reduce aerial coupling by means of a series trimmer to avoid heavy loading at aerial resonances.

Another effect, known as *squegging*, can cause poor control of regeneration, this is due to too-violent oscillation and attention to the above points will usually cure this condition also. Poor decoupling of r.f. circuits can produce what is known as *threshold howl* denoted by an a.f. howl just as the valve goes into oscillation, the remedy here is to improve the decoupling. This effect is unlikely to occur with resistance-coupled stages.

In all regenerative circuits depending on positive feedback by induction, the correct " sense " of connection of the grid and feedback windings must be observed. The sense of the windings should be the same for " grid to earth " as " h.t. to anode."

OPERATING THE RECEIVER

To obtain the optimum performance from a communications receiver whether simple of complex, requires practice and a regard to fundamental technical principles. No two receivers are exactly the same, but some operating hints are generally applicable. Typical controls are shown in **Fig. 4.62.**

Unless the receiver has an extremely slow tuning rate it will be necessary to develop a delicate touch when tuning weak signals on the higher frequency bands. This will be greatly aided by a smooth-acting slow motion drive having a large reduction ratio. Particularly on c.w., it is important to tune signals carefully, when first heard, for maximum strength and minimum interference, to forestall the sudden appearance of an interfering signal. Maximum use should be made of any crystal filter, as described later, and the selectivity should be the optimum for the particular mode since this will improve the signal-to-noise ratio.

If the receiver has separate controls for a.f., i.f. and r.f. gain (often a single control—sometimes with differing tapers —is used for i.f. and r.f. gain), the relative settings of these controls influence the apparent signal-to-noise ratio of the weaker signals. Clearly, if the r.f. gain control is near minimum setting while the other controls are well advanced, most of the gain of the receiver will be after the first frequency changer and there will be almost maximum amplification of the unavoidable noise produced in this stage. On the other hand, if the r.f. gain is always kept at a maximum, the signal-to-noise ratio will be optimum, but there is a greater risk of cross-modulation effects on the stronger signals. Generally therefore the a.f. gain control should be set at about half-travel, the r.f. gain fairly well advanced, and the level of signal or inter-station noise when searching kept under constant control by means of the i.f. gain. In the neighbourhood of strong signals the r.f. gain should be reduced. On c.w. and s.s.b. signals it may be desirable to keep r.f./i.f. gain fairly low to increase the effectiveness of the b.f.o. (unless the receiver is designed for s.s.b.) but when listening to telephony stations with the a.g.c. switched on, the r.f. and i.f. gains should be well advanced, and the a.f. control suitably adjusted.

The receiver noise should always be peaked with the aerial trimmer when first switching to a band. It should not be necessary to readjust this trimmer on individual stations except perhaps on the 1·8–2 Mc/s band.

Consideration should always be given to ensuring a reasonably good match between the aerial input circuit and the aerial or transmission line. Most factory-built receivers are designed for use with an input impedance of about 400 ohms or alternatively low-impedance 75 ohm coaxial cable. If the actual aerial impedance differs appreciably from the design figure, a worth-while improvement in signal-to-noise ratio can be obtained by the use of an aerial matching unit (see **Fig. 4.63**), using the principles described in the section on h.f. aerials. It should always be remembered that any additional gain which can be achieved in the aerial input circuit is free of valve noise.

On many receivers, the a.g.c. (a.v.c.) switch must be turned off for c.w. and s.s.b. reception to prevent the sensitivity being reduced by the action of the b.f.o.

C.W. Reception with a Single Crystal Filter

Although amateurs played an important role in the development and application of crystal filters their use is not always popular, especially among operators who have not grasped fully the technical principles involved. Such operators complain that when interference occurs, they switch in the crystal filter only to lose the required station because of a considerable reduction of strength. A correctly adjusted filter should not reduce the strength of the signal by any large amount but rather should result in considerable improvement of the signal-to-noise ratio. The reason why some users complain of serious

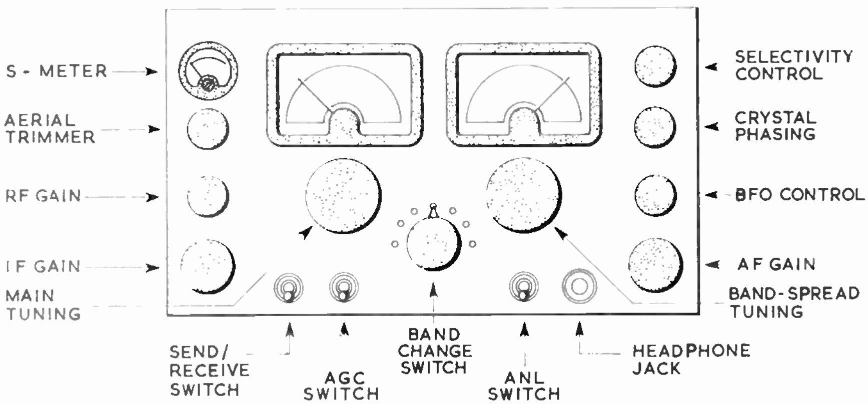


Fig. 4.62. Typical arrangement of controls for an amateur communications receiver.

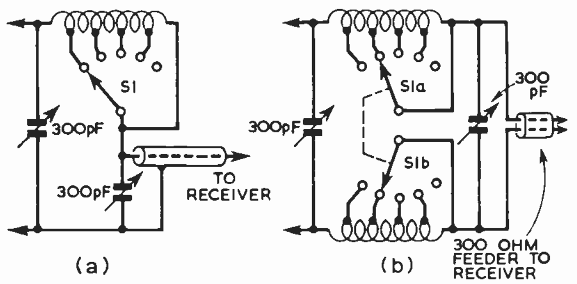


Fig. 4.63. Receiver aerial matching circuits based on the pi-network. (a) Unbalanced; (b) balanced. Since the impedance transformation is not usually as great as in transmitter output networks, capacitor values can generally be 150-300pF. Suggested winding details for the coils: 25 turns 28 s.w.g. enamelled copper wire wound on a 1in diameter former to occupy a space of approximately 1 in. and tapped at 2, 5, 8, 12 and 19 turns.

loss of strength is almost always because the b.f.o. control has not been correctly set in advance; this has meant that the station is not accurately tuned in when the crystal is switched into circuit; the loss of strength in such circumstances is really an indication that the crystal is doing its intended job of narrowing the pass band of the receiver.

It is essential with a filter using a single crystal that the b.f.o. pitch control should not be used as a form of fine tuning control, but should be set in advance to a frequency some 500 or 1000 c/s higher or lower than the crystal frequency. It should then be varied only when a change in the output note is required.

The importance of this stems from the following: suppose the b.f.o. has been set, say, 5000 c/s from the crystal or intermediate frequency but a signal is then tuned in to provide a convenient listening beat note of 1000 c/s. This means that the incoming carrier is either 4000 or 6000 c/s off the frequency to which the set is actually tuned. This may not be of great consequence with the receiver operating in a position of minimum selectivity, but immediately the passband is narrowed by switching in the crystal, the signal will be greatly attenuated; in order to tune it in correctly it would be necessary to readjust *both* the tuning and the pitch control, in practice the signal is often lost altogether.

Compare this with what happens when the b.f.o. has been correctly set in advance 1000 c/s off the nominal i.f. When the set is tuned, even in a position of broad selectivity, to provide a beat note of 1000 c/s, the signal will either be precisely tuned to the centre of the passband, or to a position 2000 c/s away. Switching in the crystal will thus have much less effect on its strength, either it remains exactly tuned, or at worst it will be on the wrong side of the beat note and can be retuned by adjusting only the tuning control.

To set the b.f.o. control, the phasing control should first be positioned at about mid-travel, and with the b.f.o. turned off, a steady carrier carefully tuned in to give maximum S-meter reading (or maximum noise output if there is no meter). The b.f.o. is then turned on and set first to zero beat with the carrier to ensure that this coincides approximately with mid-travel of the pitch control (if this is not possible, the b.f.o. trimmer may need adjustment), and then further adjusted to provide the beat note most acceptable to the particular operator—usually between 500 and 1000 c/s though a few operators prefer a much lower pitched note. This setting of the b.f.o. should be carefully noted; normally the control is not touched again, all signals being carefully tuned by means of the tuning control for maximum output. An interfering heterodyne note appearing in the passband can be eliminated by adjustment of the phasing control, the effect of which is to move slightly the very sharp rejection notch produced by the parallel resonance of the crystal.

A sharply peaked crystal filter may cause all stable c.w. signals to " ring " slightly and produce a hollow bell-like note which, if too pronounced, can make a signal difficult to read. If a variable selectivity control is fitted, the effect can usually be overcome by reducing the selectivity slightly.

RECEPTION OF S.S.B.

The reception of s.s.b. is not difficult but it does necessitate the mastery of a technique which may be unfamiliar to many. As in so many other fields, an ounce of practice is worth a ton of theory. The most likely place to find a sideband station is between 3·75 and 3·8 Mc/s or around 14·3 Mc/s. As the receiver is tuned through the signal the sound issuing from the speaker will reach a peak and will then start to fall off, but it will at no time become intelligible. Until a datum frequency is introduced in place of the missing carrier, intelligibility is impossible. For a first essay into s.s.b. reception, employing a receiver without a product or heterodyne detector the b.f.o. may be used.

With the noise peaked at maximum and the a.g.c. off, the b.f.o. should be switched on and the pitch control rotated slowly until the speech begins to sound recognizable. More likely than not, the output will sound heavily overmodulated, in which case an improvement may be made by backing off the r.f. and i.f. gain controls. A final touch to the pitch control, and the signal should become perfectly readable. If it sounds unnatural and overmodulated or like a gramophone record on a turntable which is not running evenly, the trouble is almost certain to stem from deficiencies in the receiver.

As has already been said, the insertion of a signal to take the place of the carrier is essential to the demodulation of a sideband signal. With a t.r.f. receiver, the only way of doing this would be to couple into the front-end the output of a local oscillator adjusted as closely as possible to the nominal frequency of the transmitting station. A similar system may of course be employed with a superhet receiver. The outstanding advantage of signal-frequency carrier insertion is that the intelligibility and quality of the audio output depends solely upon the stability of the carrier oscillator. As injection is done at the aerial terminals, nothing more than a small r.f. signal is required, so the oscillator may be designed for quality of output rather than for quantity. A BC221 or one of the LM series frequency meters will be found ideal. Drift in the heterodyne oscillator of a superhet will have no more effect on s.s.b. signals received by front-end carrier injection than it will on conventional a.m. It may be irritating, but nothing more. With a poor and unstable receiver, this is the only worthwhile method to use. The drawback is that the receiver tuning must be adjusted in synchronism with the carrier oscillator when changing frequency. A minor point to watch is that the carrier oscillator must not overload the front-end of the receiver. If its amplitude is great enough to drive the r.f. stage into non-linearity, the receiver would become susceptible to cross modulation by any strong signal.

When using a superhet, it is more convenient to reintroduce the carrier at an appropriate frequency after the sideband has been converted to the intermediate frequency. A

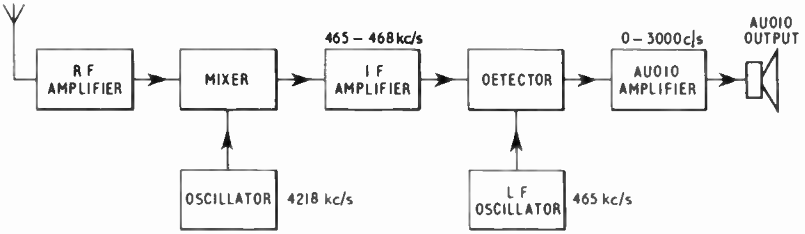


Fig. 4.64. Block diagram showing the method of carrier re-insertion at the intermediate frequency of a receiver.

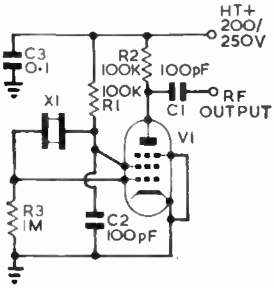
block diagram of this system is shown in **Fig. 4.64**. The b.f.o. is often employed as the carrier oscillator and produces regrettable results because it is asked to fill a role for which it was not designed. The average b.f.o. is a well-built unit which has a slow tuning rate and is quite stable enough to take the place of the carrier. In the majority of receivers its output is via a small capacitor to the anode of a diode detector, and its amplitude is restricted to the lowest level at which an acceptable beat note is obtained for the reception of c.w. telegraphy. When an attempt is made to mix a sideband signal with that from the b.f.o., considerable distortion will occur if the amplitude of the simulated carrier is exceeded by that of the sideband. To reduce distortion to negligible proportions, the maximum amplitude of the s.s.b. signal must be held well below that of the b.f.o. signal. This can be done by reducing the r.f. and i.f. gain but this solution can never be more than a compromise. When the gain is advanced to read a weak signal, splatter becomes unavoidable from any stronger signal which may happen to be adjacent. Increasing the output of the b.f.o. is rarely profitable, because stability is likely to suffer. In some receivers a considerable improvement may be effected by disconnecting the lead from the b.f.o. to the diode detector and connecting it to the grid of the first or second i.f. amplifier valves via a capacitor of 1 or 2 pF. This modification will not, however, work well if the i.f. stage is followed by a highly selective device such as a crystal or mechanical filter.

One answer to the problem is to replace the diode detector by a mixer-type demodulator. A representative circuit of this type is shown in **Fig. 4.65**. With this circuit, which is commonly called the *product detector*, the applied b.f.o. voltage is less critical than with a diode detector, yet the audio output is much cleaner. It is possible to add the double triode and the few associated components into most receivers, but the detector will work equally well if constructed as an outboard accessory. The product detector is also useful for c.w. reception but it will not demodulate a.m. or n.b.f.m. telephony. A switching arrangement such as that indicated in Fig. 4.65 must therefore be included to enable the receiver to handle all forms of signal. Conventional a.m. may be demodulated by the so-called exalted carrier method. This is the technique in which the original carrier is filtered out, as explained earlier in this chapter, and then replaced by a locally generated, but appreciably stronger, carrier.

For successful i.f. carrier reinsertion, it is essential that both the local oscillator and the b.f.o. should be as free from drift as it is possible to make them. A consistent slow drift is not much of a handicap but erratic, short-term frequency jumps are particularly disconcerting. In receivers which are prone to this fault, it is usually more satisfactory to install a crystal oscillator at an appropriate frequency within the i.f. passband than to attempt to cure the offending b.f.o. A typical circuit is given in **Fig. 4.66**, for which suitable crystals are readily available. If the i.f. passband is wide enough to accept two sidebands, it will suffice to obtain one crystal of which the frequency should be centred in the middle of the passband. If however, the passband approaches the more desirable selectivity characteristic of 3 kc/s, two crystals will have to be installed in a switchable circuit so that the carrier may be inserted to one side or the other of the passband, thereby permitting either upper or lower sideband reception at will. **Fig. 4.67** will give the idea much more quickly than words.

There are a few possible modifications which despite their simplicity may make a

Fig. 4.66. Crystal-controlled carrier insertion oscillator.



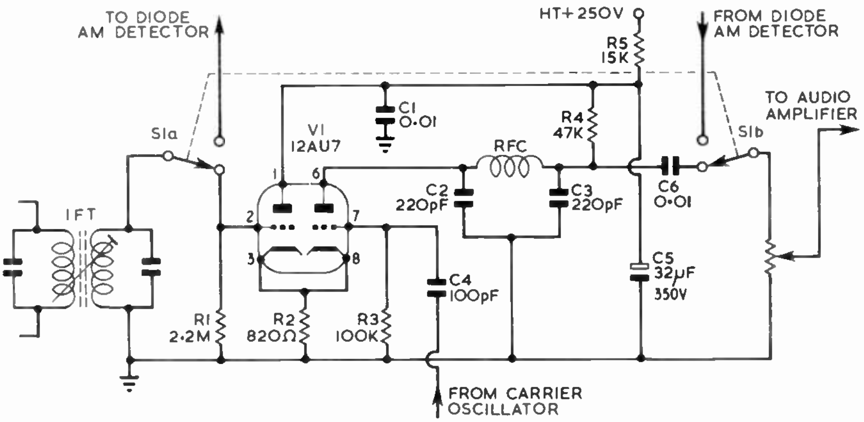


Fig. 4.65. Typical twin-triode product detector circuit. RFC—1·5mH to 2·5mH, VR1 500K to 1 Megohm.

considerable improvement to many local oscillators. As drift is frequently caused by changes in the potentials applied to the various electrodes of the oscillator and mixer valves, stabilization of the h.t. supply by a regulator voltage tube such as the OC3 or VR105/30 is invariably beneficial. If the mixer valve is included in either the manual or the automatic gain control loops, it should be disconnected therefrom and the cathode and grid returns solidly earthed. An alteration to the send/receive switching is highly recommended in many receivers. A common function of this switch is to disconnect the h.t. from all valves completely when it is thrown to SEND. A minor rearrangement of the wiring can be made to permit the local oscillator and the b.f.o. to continue running and so will minimize erratic drift caused by alternate heating and cooling. The resulting improvement in stability will vary widely from one receiver to another, but will almost always be out of proportion to the simplicity of the modification.

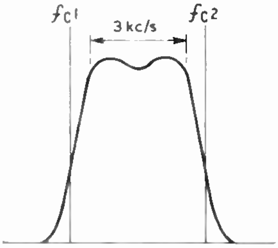


Fig. 4.67. Alternative carrier frequencies for selectable sideband reception with a highly selectable receiver.

Finally, if the receiver is not particularly selective, the advisability of reducing its bandwidth to 3 kc/s may be worth considering. The advantage to be gained from this modification is not confined solely to s.s.b. work, but a single sideband transmission 3 kc/s wide tuned on a receiver with a 3 kc/s passband represents the highest efficiency and best signal-to-interference ratio at present attainable in telephony communication. The finest way to increase selectivity at reasonable cost is to install two sections of half-lattice crystal filter in the i.f. amplifier or to add a compact mechanical filter.

Reception of D.S.B.

This section would not be complete without a mention of double-sideband, suppressed-carrier system of communication. Theoretically, this system has several advantages over s.s.b. but in amateur practice these cannot usually be achieved. It has the major disadvantage, compared with s.s.b., of occupying twice the bandwidth.

The least complicated way to receive d.s.b. is to turn it into s.s.b. by means of an extremely selective i.f. filter such as that in **Fig. 4.68**. With a filter of this kind, the signal is virtually indistinguishable from genuine s.s.b., except that tuning is more critical, and slight misadjustment will lead to a " growl " superimposed on the audio. This may be eliminated by reducing the low frequency response of the receiver audio stages in the manner recommended for transmitter speech amplifiers.

To make the most of d.s.b., both sidebands have to be accepted on a receiver with a passband of 6 kc/s, and the local carrier must be locked in phase with the nominal carrier by means of information derived from the sidebands themselves. The block diagram of **Fig. 4.69** shows one possible system. After conversion to the intermediate frequency, the signal is applied to two product detectors *A* and *B*. The local oscillator voltage is passed through a 90° r.f. phase shift network before reaching the detectors. If the injection voltage at *A* is in the correct phase relationship, maximum audio will appear in the output, but no signal at all will emerge from detector B because the carrier is introduced 90° out of phase. When the oscillator drifts slightly, the *A* channel will be little affected, but some output will start to appear from B. This will be in phase with the *A* channel for one direction of drift, and 180° out of phase for drift in the opposite sense. By amplifying the audio from both demodulators and combining it in a phase discriminator, a d.c. voltage is obtained which may be applied to a reactance valve to provide automatic phase control for the oscillator. The obvious complexity of this arrangement shows fairly clearly why it is simpler to generate a s.s.b. signal in the first place than to attempt reception of both sidebands of a d.s b signal.

RECEPTION OF N.B.F.M. SIGNALS

Conventional frequency modulation of the type used for v.h.f. radio and television (in the UK only on the 625-line system) stations is not directly suitable for h.f. communications because of the broad bandwidth required. These systems have frequency deviations of the order of ±30-75 kc/s with sidebands extending well beyond these limits. For

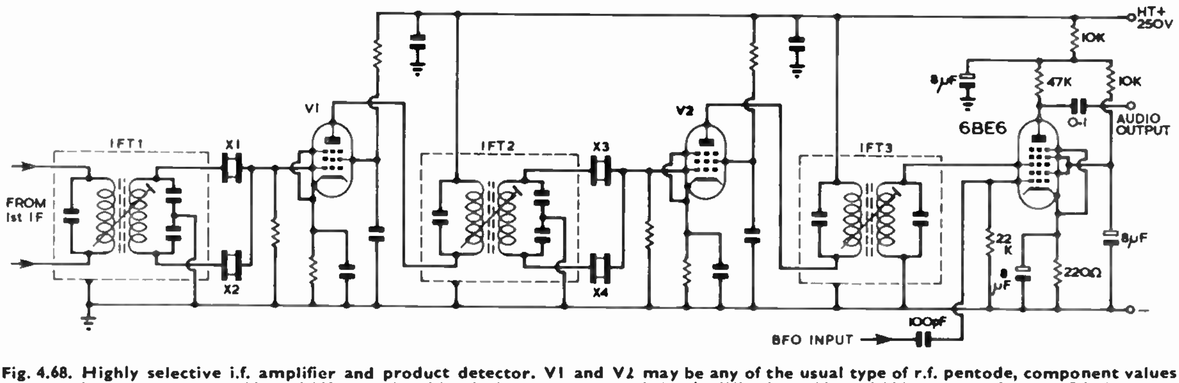


Fig. 4.68. Highly selective i.f. amplifier and product detector. V1 and V2 may be any of the usual type of r.f. pentode, component values being chosen to suit. X1 and X3 should be identical in frequency and should differ from X2 and X4 by approximately 2 kc/s.

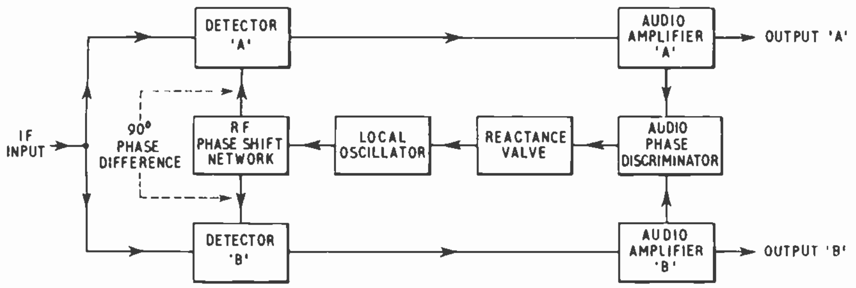


Fig. 4.69. Simplified block diagram of the synchronous detection system for d.s.b. reception. Audio output may be taken from either amplifier A or amplifier B but not from both in parallel.

communications applications much narrower deviations may be employed, though this means that the signal/noise improvement of wideband f.m. with pre-emphasis is lost.

Narrow band frequency modulation (n.b.f.m.) is permitted on all amateur bands with a maximum deviation of $\pm 2{\cdot}5$ kc/s and a maximum modulating frequency of 4000 c/s. Although this deviation may appear low in comparison with broadcast systems, the percentage deviation when received on a set having a final i.f. of 85 or 465 kc/s may in fact be much higher than that of an f.m. broadcasting station received with the standard i.f. of 10·7 Mc/s.

The communications efficiency of n.b.f.m. can therefore be good, and with the related *phase-modulation* technique offers the advantages of simple low-level modulation, without requiring the subsequent use of linear amplification; an n.b.f.m. or ph.m. transmitter may thus be significantly cheaper to built than either an a.m. or s.s.b. transmitter of equivalent power. Military tests have shown that in some circumstances phase modulation offers appreciable advantages over a.m. for portable h.f. operation.

With phase modulation the deviation, unlike that of true f.m., varies with the modulation frequency but can be converted to f.m. form by introducing a 6db/octave bass boosting network either in the transmitter or receiver audio circuits.

A major advantage of the n.b.f.m. system is that it is less likely to cause TVI than a.m., it is also attractive for portable work where the total power available is limited. It should be recognized that n.b.f.m. is rather more complex in setting up than a.m., and the need exists to adjust deviation when changing bands. Furthermore the system can cause appreciable sideband splatter unless the higher audio frequencies are attenuated (in theory an f.m. signal may have an infinite series of sidebands).

In practice, despite its potential value, the n.b.f.m. system is used by amateurs far less than either a.m. or s.s.b. except in the form of frequency shift keying for radio teleprinters. For this reason few h.f. communications receivers are fitted with a suitable *discriminator* for the demodulation of n.b.f.m. or ph.m. signals, and operators often rely on the technique of detuning the receiver to one side of the nominal carrier frequency, with the a.g.c. turned off. While this system can be reasonably effective when the overall response curve of the receiver is suitable, it is recommended that for serious f.m. operation a discriminator should be used.

To demodulate an f.m. signal a two-step operation is required: an f.m. signal from which all amplitude variations have first been eliminated by means of a *limiter* is applied to a frequency/amplitude conversion circuit, this changes the signal into one of conventional a.m. form with amplitude dependent upon the frequency deviation of the incoming f.m. signal and can be demodulated by means of a normal envelope detector. These two processes (and sometimes the amplitude limiting) are carried out in a single stage by means of an f.m. discriminator.

To understand why f.m. can be received by detuning a receiver, reference should be made to Fig. 4.70 which shows the selectivity curve of a typical i.f. amplifier. For a.m. signals, the receiver is normally tuned so that the wanted signal is in the middle of the passband or at the peak, i.e. point A. If the receiver is detuned to one side or the other, say point B, it will respond to a signal of varying frequency as though its amplitude was changing, since it will receive the signal with varying effectiveness according to the point on the slope represented by the particular incoming frequency. If the frequency variation of the signal is at an audio frequency (as it will be for n.b.f.m.), then the set provides an audio output proportional to the magnitude of the change in frequency. Any attempt to receive the signal with the a.g.c. circuits in operation will degrade the response. However, the frequency/amplitude conversion is generally far from ideal since the i.f. response curve is unlikely to be linear.

Many special discriminator circuits have been evolved some of which require the signal to be pre-limited, although in others the limiting is carried out within the discriminator stage. Some discriminators require the use of special multigrid valves such as the nonode (EQ80) or gated-beam (6BN6) valves; others need two diodes which may be either thermionic diodes or crystal (semiconductor) diodes. The most common forms of discriminators using diodes are the Travis (double tuned circuit), the Foster-Seeley and the self-limiting *ratio-detector*.

A discriminator for n.b.f.m. should be designed so that

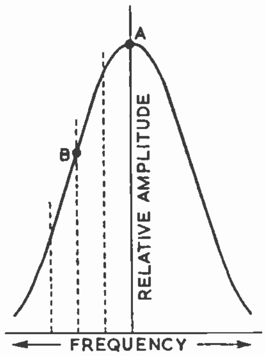


Fig. 4.70. Reception of an n.b.f.m. signal on the skirt of the i.f. selectivity curve of an a.m. communications receiver. A signal of varying frequency when tuned to a point such as B on the skirt will cause a corresponding variation in the current through the following detector.

its linear response just covers the basic deviation of $\pm 2\cdot5$ kc/s plus the more significant sidebands, or say roughly 7·5 kc/s bandwidth. It should be noted that the use of a wideband discriminator using a broadcast i.f. of 10·7 Mc/s is not suitable for n.b.f.m., and this means that the special ratio-detector discriminator i.f. transformers with a tertiary winding cannot be used. However, 465 kc/s i.f. transformers with centre-tapped secondary windings are marketed (intended primarily for use with crystal filter circuits) and this means that a Foster-Seeley discriminator can be constructed. Since this type of discriminator initially requires the removal of amplitude variations of the incoming signal, it is necessary to precede the discriminator with one, or preferably two, stages of limiting. This is usually achieved by the use of pentode amplifiers operated with a suitable time-constant network in the grid circuit and/or in a saturated condition by the use of very low screen voltages.

Fig. 4.71 shows a 465 kc/s n.b.f.m. adaptor using two limiting stages followed by a Foster-Seeley discriminator suggested by G2HCG; although originally intended for use on 144 Mc/s it would be equally suitable for h.f. operation. The 9001 type valve was chosen because of its small size, low power requirements, short grid base and relatively low gain; although other types (such as octals 6J7, 6SJ7, 6SS7) could be used, modern high-gain types should be avoided.

Travis or gated-beam discriminators can be designed and adjusted to give the required results, and to provide a high-volts-per-kc/s-deviation output; but few entirely suitable designs based on these circuits have been published in recent years.

A four-diode phase discriminator intended for use with 465 kc/s i.f. stages is shown in **Fig. 4.72**, offering the advantage of not requiring a centre-tapped i.f. transformer. This uses a capacitive divider to obtain the centre point in conjunction with two double diode integrator detectors. Such an arrangement, as with the Foster-Seeley discriminator, should be preceded by some form of amplitude limitation.

There would appear to be excellent scope for further development of n.b.f.m. for amateur h.f. and v.h.f. applications; but the relatively few amateurs equipped for really effective reception of this mode tends to discourage its wider use. An n.b.f.m. signal received on an a.m. receiver by off-tuning tends to sound under-modulated and does not give really good communications effectiveness.

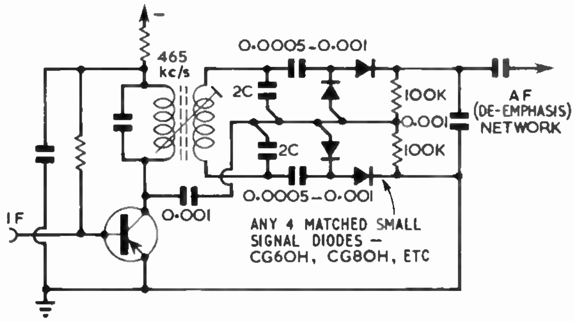


Fig. 4.72. A four-diode phase discriminator for narrow-band frequency modulation reception.

MAINTENANCE AND FAULT-FINDING

A difficult problem for an amateur operator is knowing whether his receiver is providing the optimum performance of which it is capable. Valve or component deterioration or the gradual drifting out of alignment of the tuned circuits may degrade performance to a marked extent before the user becomes aware of it, poor results being attributed to the vagarities of h.f. propagation conditions. It should be appreciated that, no matter how well constructed, receivers containing up to 20 valves are certain to require occasional servicing attention. The use of semiconductors instead of valves can bring about a considerable improvement in reliability—provided that care is taken in design and use to avoid subjecting the devices to over voltages.

In a multi-valve receiver, the failure of one or two stages to contribute their full quota of gain may remain unnoticed because of the considerable reserve of gain usually available in such sets yet, especially where the fault lies in an early stage, this may result in serious degrading of the signal-to-noise ratio. There is no simple answer to this problem. Few amateurs have available the laboratory-type instruments necessary to measure h.f. receiver performance to the degree of accuracy required to show up only slight deterioration. In practice, it is by no means unusual for a long spell of dis-

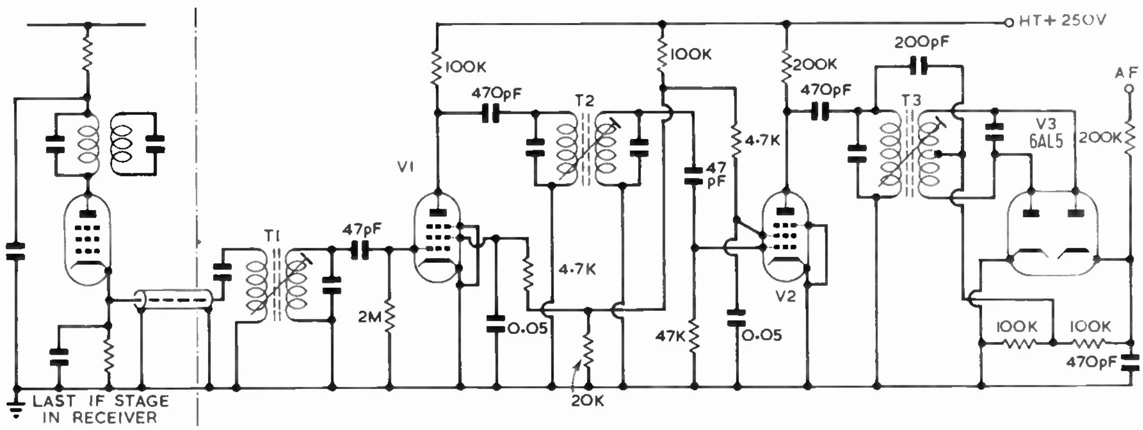


Fig. 4.71. A 465 kc/s n.b.f.m. adaptor.

appointing activity on the higher frequency bands to be traced finally to an unsuspected receiver or aerial fault. There are almost certainly many receivers in use which could be greatly improved by careful re-alignment, by replacement of leaky decoupling capacitors or by the substitution of new valves either the same type or one of the newer low-noise types, though the latter should only be undertaken by those with the requisite knowledge.

An operator should always endeavour to memorize the true " feel " of a receiver when it is working well, so that any later deterioration can be more readily noticed. The S meter readings of the crystal calibrator or an external oscillator such as a BC221 are useful for checking performance. It is useful also to keep under observation the results being achieved by other amateurs (noting the RST reports they give is not always a good basis for this—better to observe how much difficulty they have in copying weak signals). The selectivity characteristics of a receiver can be kept fairly accurately under review by noting its single-signal capabilities on c.w. signals or the degree of ringing on a single crystal filter.

Valve deterioration can often be readily checked by direct substitution of known good valves, though this can sometimes be an uncertain guide because of the need in some stages for slight retrimming to compensate for differences in inter-electrode capacitances when changing valves.

An occasional routine check of the voltages applied to the valve sockets of each stage is most useful, not only for locating actual faults but also for bringing to light gradual changes in valve currents and component values. For maximum usefulness such checks should always be carried out on the same testmeter (preferably of at least 1000 ohms per volt sensitivity or better still 10,000 or 20,000 ohms per volt) to eliminate the effect on the readings of the current drawn by the meter. It is good practice to note the readings each time into an equipment record book. Most receiver manufacturers issue service information listing typical valve voltages to be expected with meters of defined sensitivity, but it is better to prepare a table of measured voltages for the individual receiver, and to check these from time to time. As the valves age, there will inevitably be some slight changes (of the order of 10 per cent) which can be safely disregarded. Watch should be kept for the fairly pronounced changes which would indicate that a valve is becoming "soft," its emission failing, or it is otherwise reaching the end of its useful life. Measurement of valve voltages will also help pinpoint such common faults as the gradual increase in value of feed resistors carrying d.c., leakage of decoupling capacitors and high-resistance joints.

To trace the more obscure faults—or those of an intermittent nature—may require considerable patience and some knowledge of radio servicing techniques—many of which are basically applied common sense and a knowledge of Ohm's Law. Because an amateur is seldom so concerned with the time factor as a professional service engineer, it is usually possible for him to trace eventually even the most difficult faults, with the aid of only a testmeter and a small supply of substitute components: the amateur with experience of constructional work soon develops the necessary flair for fault-tracing.

Because it is easy to accidentally damage or destroy semiconductors during servicing, tracing faults in transistorized receivers calls for special care and additional knowedge. This question is discussed in the RSGB publication *Technical Topics for the Radio Amateur* and in the many radio servicing books.

Re-alignment

The performance of a receiver is governed to a great extent by the accuracy with which the various tuned circuits in the i.f. and r.f. stages are aligned. In all equipment there is the tendency for such circuits to drift gradually off-tune so that in time re-alignment becomes necessary. The first indication of poor i.f. alignment of a communications receiver is usually a marked reduction in the effectiveness of a crystal filter. Fortunately, modern first-grade components—particularly i.f. transformers and air-spaced timmers—are extremely stable, and in normal circumstances only minor re-alignment is likely to be needed. This is usually only to take account of variations in valve inter-electrode capacitances when replacing valves. The performance of older receivers however, can often be substantially improved by regular re-alignment, particularly where the receiver is subjected to vibration or excessive humidity.

There are two approaches to the problem of re-alignment (or three if one includes taking the set to a good service engineer with experience of communications receivers).

Most manufacturers issue detailed instructions on the procedure recommended for the model concerned, listing such important matters as the correct i.f. and r.f. alignment frequencies, order of adjustment and the layout of the various trimmers and cores. To carry out these procedures generally requires access to at least a reasonably good signal generator, output meter and a supply of non-metallic tools for adjustment of trimmers and cores.

The other approach needs only a keen ear, a few trimming tools, and—vitally important—a clear knowledge of what should and should not be attempted. Remarkably good results can be achieved by following this empirical method. But injudicious, haphazard or excessive adjustments can quickly and completely desensitize a receiver so that it can only be made to work again by a full re-alignment carried out with proper servicing equipment. The amateur who is capable of benefiting from this technique should need no guidance, those who do should never attempt to use it.

A useful accessory when checking the accuracy of the r.f. alignment of receivers having accessible air-cored coils, without requiring the alteration of trimmers, consists of an insulated non-metallic rod about 6 in. long with a small piece of dust-iron or ferrite core material fastened at one end of the rod and a similar small piece of brass at the other. When the brass end is inserted into an h.f. coil it will lower the effective inductance; whereas inserting the dust-iron end will increase the inductance. This device (often called a *tuning wand*) can be used to check alignment at various points on a waveband by noting the effect on a signal when the rod is gradually inserted. If the r.f. circuit is correctly adjusted, inserting either end of the rod will reduce the signal. On the other hand if the circuit is slightly off-tune, the signal can be peaked when inserting one or other end of the rod, depending on whether the circuit is tuned to the high or low frequency side of the signal. This device, unfortunately, cannot always be used on modern receivers, as these often have coils with variable cores which prevent the rod from being inserted.

PRACTICAL DESIGNS

For a number of years amateurs have tended to use factory-built h.f. communications receivers, either those designed specifically for amateur or commercial work, or modified ex-Government receivers, some manufactured as long ago as World War 2. Recently, however, there has been a marked revival of interest in home-construction, particularly of receivers of advanced, modern design. One reason for this is economic: the factory production of high-performance receivers to the most stringent modern requirements (and with the wide flexibility needed to cater for the differing interests of a large number of users) is inevitably costly, and the price of such models is often some hundreds of pounds. On the other hand, the experienced constructor who is prepared to spend considerable time and effort can often build extremely good receivers for a much more modest sum.

The newcomer to receiver construction is advised to gain experience before tackling an advanced, full-specification receiver, unless skilled assistance is readily available. There is no better way of gaining a sound knowledge of receiver theory and practice than by tackling several designs of a relatively simple nature. A straight (t.r.f.) valve or transistor receiver will teach the builder much about components, wiring and layout, careful adjustment of regeneration and other important points. A simple superhet is invaluable for making clear the problems of i.f. and r.f. alignment and the reduction of spurious images. Once basic experience had been gained, much more ambitious receivers can be tackled with every confidence, even by amateurs having available only a minimum of test equipment; this is especially true if the r.f. coil unit is obtained as a complete assembly or alternatively if plug-in coils are used.

Even if a factory-built receiver is used, there is usually no reason why it should not be improved, or made more suitable for the particular interests of the user: for example, ex-Government receivers such as the R1155, R107, R1475, CR100, BC342, BC348 and the Command series (BC453/4/5) or the more-esteemed HRO, Super Pro, SX28 and AR88 models. The performance of the simpler models can be greatly improved by fitting an external pre-amplifier, *Q* multiplier or crystal filter, S meter and calibration oscillator, while even the best receiver can be improved by an efficient aerial matching unit.

It is possible to extend the tuning range and improve performance on the higher frequency bands by means of a modern converter. High-performance converters can be built in a few evenings and yet can show the constructor almost as much about the principles and problems of superhet design as a complete receiver.

Many of the older receivers use r.f. and mixer valves which are relatively noisy by modern standards. The sensitivity of such receivers can be greatly improved by the use of a low-noise converter or by modifying the front-end circuits.

A set which is deficient in selectivity may be improved by adding an external unit, comprising a second frequency changer followed by an 85-100 kc/s i.f. section (a device often referred to as a Q5'er) using, if desired, the BC453 Command receiver for the purpose of alternatively an outboard i.f. amplifier with a half-lattice crystal filter, or a mechanical filter. A *Q* multiplier can also be used to improve selectivity.

For c.w. reception it may be possible to improve results by fitting an audio filter or selectoject type of arrangement.

The keen amateur never takes the design of even an expensive receiver too much for granted. All receivers turned out on an assembly line are essentially a matter of compromise, whereas the individual amateur tends to specialize in his interests. The c.w. operator will have very different ideas on what constitutes an ideal receiver to the operator who generally uses telephony. Those who work mainly on 1·8 and 3·5 Mc/s (where site noise levels are likely to be high) will not worry so much about extreme sensitivity as the 21 or 28 Mc/s enthusiast.

This is not to suggest that one should rush into tampering haphazardly with a complex receiver of modern design; but there is often no reason why new ideas should not be tried—particularly where they take the form of an external unit. Useful devices are regularly described in Amateur Radio magazines. It is important that the amateur should not be content to regard any receiver as a " black box " labelled " NOT TO BE OPENED."

For those who want to build a receiver, but do not wish to spend too much time on shopping for components, complete kits with step-by-step construction manuals are available.

A SIMPLE T.R.F. RECEIVER

The straight 1-V-1 (a traditional way of indicating a receiver having one stage of r.f. amplification, detector, and one stage of a.f. amplification) can still form an effective receiver for the listener and amateur provided that its fundamental limitations are recognized. These include an inherent lack of selectivity which becomes progressively worse with increasing frequency and the inability to resolve weak modulated signals. Apart from these limitations, it can compete with all but the best superhets in the realm of c.w.

A rear view of the t.r.f. receiver. The components may be identified by reference to Fig. 4.74 (b).

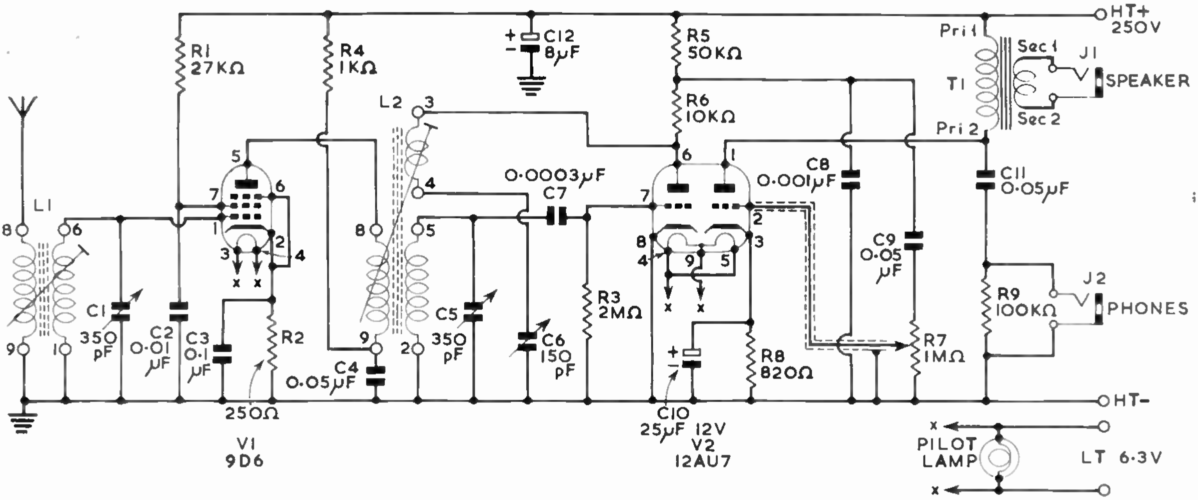


Fig. 4.73. Circuit diagram of simple t.r.f. receiver using miniature valves.

reception so long as extreme selectivity is not a first requirement, particularly on the lower frequency bands such as 1·8 and 3·5 Mc/s.

The circuit of a simple receiver of this type, built by G3NGS to a design by the late G8TL, is shown in Fig. 4.73. The circuit has three stages: a variable-mu pentode as r.f. amplifier, and two triodes contained in the same valve envelope, operating as regenerative detector and a.f. amplifier.

The signal picked up by the aerial is transformer-coupled to the first tuned circuit consisting of the secondary of L1 and the tuning capacitor C1 and applied to the control grid of V1, the r.f. amplifier valve. After amplification the signal is fed via the untuned primary of L2 and its associated tuned secondary to the coupling capacitor C7 and so to the grid of the detector V2a. It will be observed that a third winding (connections 3 and 4) is included in the L2 assembly and wired from the anode of V2a via the variable capacitor C6 to earth. This third winding is closely coupled to the grid coil (connections 5 and 2) to provide regeneration or reaction. Regeneration increases the sensitivity of V2a as a detector and if increased far enough causes the valve to oscillate. The amount of regeneration or reaction is controlled by the variable capacitor C6. For phone work it is best to set the control just below the point where oscillation begins but for the reception of c.w. (telegraphy) signals C6 should be advanced to the point where V2a just begins to oscillate. This point will be marked by a faint " rushing " sound in the headphones or loudspeaker. No advantage will be gained by increasing the amount of regeneration beyond this point. R6 acts as an r.f. load in the anode circuit of V2a for regeneration.

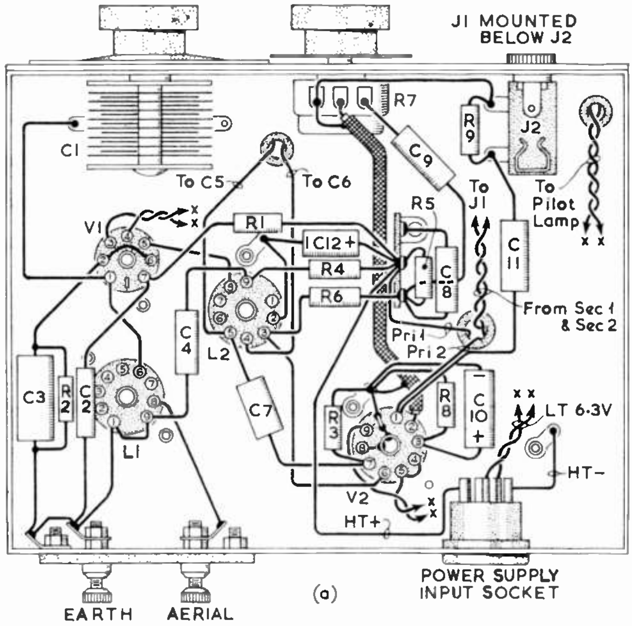
The audio signal from the detector is built up across the load resistor R5 and fed via C9 to the volume (or gain) control R7. The slider of this control (the centre terminal) is connected to the grid of V2b which functions as an audio amplifier to raise the level of the signal sufficiently to drive a small loudspeaker or headphones. Bias for this valve is developed by the flow of current through R8 in a similar manner to the bias for V1. The amplified audio signal is fed to the loudspeaker by transformer T1 which provides the necessary match for a low impedance (3 ohms) speech coil. The use of headphones is made possible by feeding the audio signal through C11 to the jack socket J2.

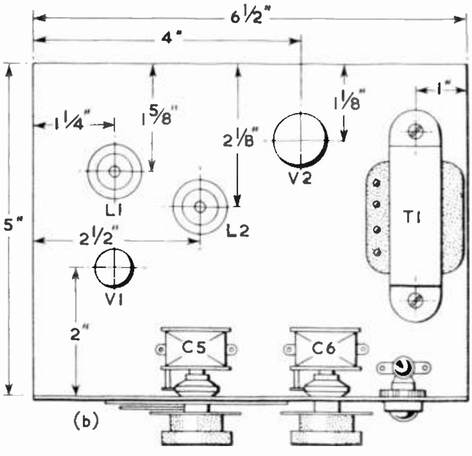
The complete receiver, with the exception of the power supply, can be built on an aluminium chassis measuring

TABLE 4.3

COMPONENTS TABLE FOR FIG. 4.73

C1, 5	350 pF air-spaced variable capacitor (Jackson Bros. (London) Ltd., U-type miniature)
C2	0·01 μF 350 volt working capacitor (Dubilier)
C3	0·1 μF 200 volt working capacitor (Dubilier)
C4, 9, 11	0·05 μF 350 volt working capacitor (Dubilier)
C6	160 pF air-spaced variable capacitor (Jackson Bros. (London) Ltd.)
C7	0·0003 μF silver mica capacitor (Dubilier)
C8	0·001 μF 350 volt working capacitor (Dubilier)
C10	25 μF 12 volt electrolytic capacitor (Dubilier)
C12	8 μF 350 volt electrolytic capacitor (Dubilier)
L1	Maxi-Q miniature coil (blue) for range desired (Denco (Clacton) Ltd.). Screening can supplied with coils.
L2	Maxi-Q miniature coil (green) for range desired (Denco (Clacton) Ltd.). Screening can supplied with coils.
R1	27 K ohms ½ watt (Dubilier)
R2	250 ohms ½ watt (Dubilier)
R3	2 Megohms ½ watt (Dubilier)
R4	1 K ohm ½ watt (Dubilier)
R5	50 K ohms 2 watt (Dubilier)
R6	10 K ohms 2 watt (Dubilier)
R7	1 Megohm potentiometer with switch (Dubilier)
R8	820 ohms 1 watt (Dubilier)
R9	100 K ohms ½ watt (Dubilier)
	Miscellaneous
T1	output transformer for speaker
2	Noval valve bases for coils (McMurdo XM9/UC1)
1	dial lamp (Bulgin D170 or similar)
2	phone jacks (Bulgin type J2)
1	9D6 valve (Brimar)
1	12AU7 valve (Brimar)
2	terminals (aerial and earth) (Bulgin)
2	knobs with skirts (Bulgin K401 and K405)
2	knobs (Bulgin K400 and K410)
1	six-way connector (Bulgin P149)
6	solder tags (Bulgin T17)
1	Noval valve base for V2 with screening can (McMurdo XM9/UC1)
1	B7G valve base for V1 with screening can (McMurdo XM7/UC1)
2	slow motion epicyclic 6 : 1 drives for C5 and C6 (Jackson Cat. No. 4511)





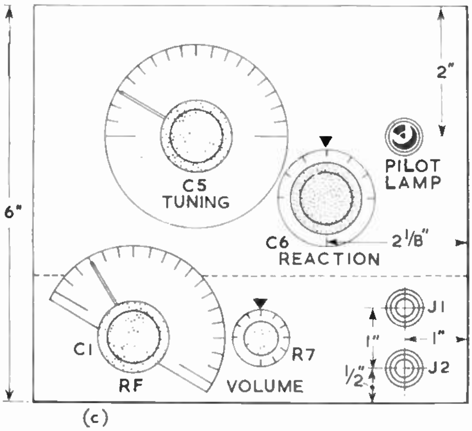


Fig. 4.74. Wiring and layout diagrams of the simple t.r.f. receiver. (a) Underchassis layout of the components showing point-to-point wiring. (b) Above-chassis layout of components. (c) Arrangement of the controls on the front panel.

6½ in. × 5 in. × 2 in. fitted with a front panel 6 in. high, though the dimensions are by no means critical.

Supplies to the valve heaters should be connected with twisted heavy gauge wire to reduce hum and voltage drop. One side of the heater line may be earthed at the power input socket or, if hum is prevalent, a " humdinger " should be fitted. A humdinger is a low-value potentiometer (about 50 ohms) with the outer two contacts connected across the heater line and the centre contact (slider) earthed. The slider should be moved around the centre of its travel until a minimum of hum is found. The potentiometer should then be left at this setting.

Component layout is not critical so long as interstage wiring is short and direct. The radio frequency amplifier (V1 and its associated components) should be well screened from the following stages to lessen the chance of feedback. For the same reason C1 should be mounted sub-chassis and C5 above chassis level. The only other components above the chassis are the speaker transformer T1 and the reaction control C6. Miniature slow-motion drives should be fitted to C5 and C6 to ease the tuning in of weak signals.

Fig. 4.74(a) shows the underchassis layout and provides a point-to-point wiring diagram. The placement of parts above chassis is shown in **Fig. 4.74(b)** and the front panel layout in **Fig. 4.74(c).**

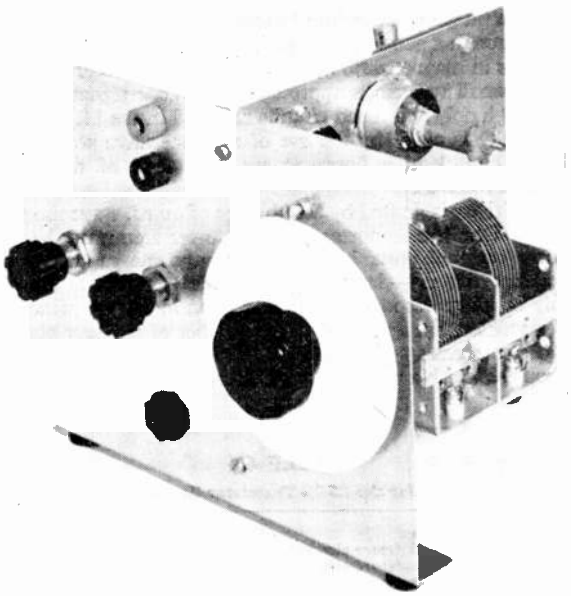
The power requirements of the receiver are very small and a power pack giving 6 volts at 1 amp and 250 volts at 30 milliamps will be found to be quite adequate. The live side of the mains may be brought into the set via the six-pin plug, taken through the switch on the volume control (R7), and taken out again to the primary of the mains transformer.

Before switching on for the first time, it is a good practice to check for any short across the h.t. terminals as this can cause serious damage to the power pack. If, on switching on, R5 and R6 heat up, check that the vanes of C6 are not bent or shorted by metal filings. If the set does not function after ample time for warming up, check that the coils have the same number and are inserted correctly.

A list of components is shown in the accompanying table together with the names of manufacturers of suitable items. There is, however, no reason why electrically equivalent parts of other makes should not be used.

Simple Superhet Receivers

Two and three valve " straight " receivers can still provide acceptable results for c.w. reception on the lower frequency bands, but require most careful adjustment and operation if tolerable selectivity is to be obtained on the higher frequency bands. This is especially true if the receiver is required for two-way working as opposed to general listening where it normally matters less if stations are at times blocked out by adjacent channel interference. For serious operation throughout the h.f. spectrum (or for use in conjunction with a v.h.f. converter), it is practically essential to use a superhet



The RSGB Transistor Four designed by G3HBW as a low-cost superhet.

design in which both high gain and good selectivity are obtained at a relatively low radio frequency.

For the home constructor, a first h.f. superhet represents an important milestone. This is not so much because of any inherently greater complexity in its construction; indeed a simple superhet may contain appreciably less components and require less careful screening than would a good t.r.f. receiver. Rather it is because in the construction of a superhet, the basic assembling and wiring represents only the first parts of the complete work, almost as much effort and as much skill may be required in the adjustment of the tuned circuits and the calibration of the receiver.

With a superhet one may complete the set absolutely correctly and yet be unable to receive any signals at all until a certain stage in the alignment is reached. This presents considerable difficulties to the constructor who does not have access to an adjustable calibrated oscillator of the signal generator or grid dip type and who has therefore to rely on " blind " adjustment of the circuits. It must be stressed that the complete alignment of a complex single or double conversion superhet receiver without an adjustable oscillator can be a difficult operation calling for considerable patience and not a little luck. Where any form of calibrated oscillator covering the i.f. and/or h.f. ranges is available the difficulties are greatly reduced.

RSGB TRANSISTOR FOUR

A simple, low-cost transistorized superhet receiver, using commercially available plug-in coils can be built cheaply and can give many hours listening for each small 9-volt (PP3) battery, as well as giving the constructor a useful insight into small communication receivers. A receiver of this type can be built for appreciably less cost than a mains-operated valve receiver, and the complete absence of the mains power pack is an attractive safety feature when the set represents an initial constructional project. The receiver described below is based on a design prepared by A. L. Mynett, G3HBW, for the RSGB Educational Committee.

It is intended primarily for the headphone reception of amateur stations in the 1·8 and 3·5 Mc/s bands using a single pair of coils (1·8-5 Mc/s) but is capable of receiving amateur and broadcast signals throughout the range 515 kc/s (580 metres) to 31·5 Mc/s (9·5 metres) with the appropriate plug-in coils. The circuit is shown in Fig. 4.75.

The receiver uses an OC170 (or AF115) h.f. transistor as

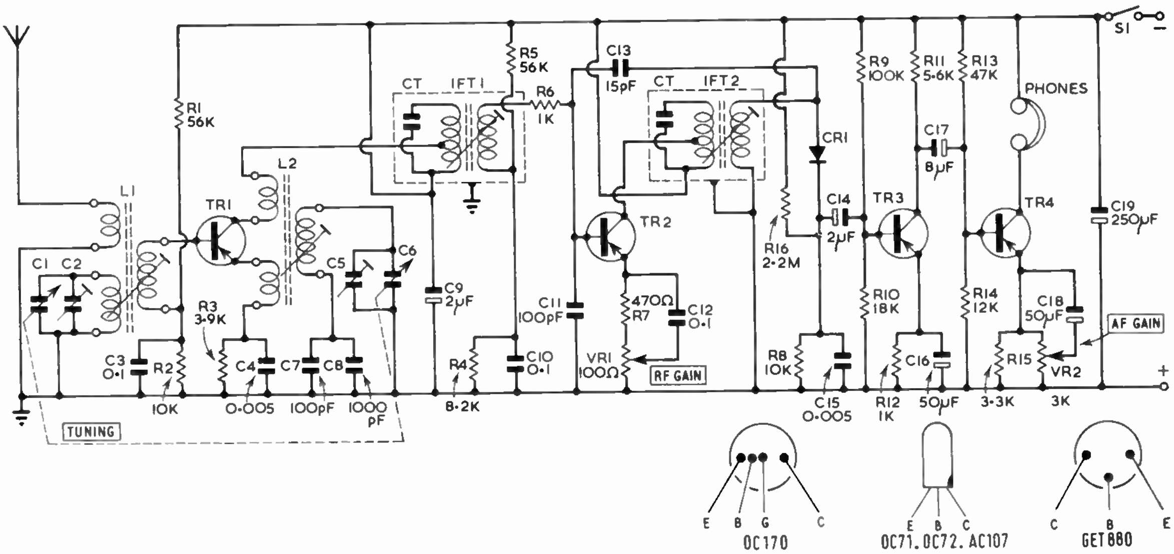


Fig. 4.75. Circuit diagram of the RSGB Transistor Four.

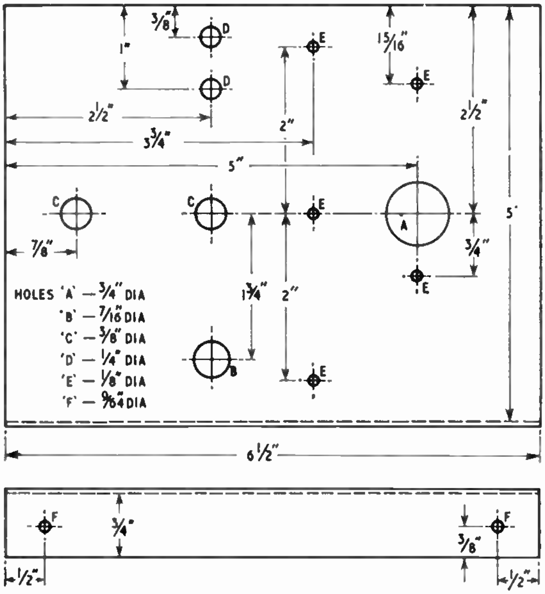


Fig. 4.76. Drilling diagram for the front panel.

a self-oscillating mixer (TR1) providing an i.f. output on about 465 kc/s. A second OC170 (TR2) is used as a regenerative i.f. amplifier, with the gain/regeneration controlled by VR1 which adjusts the amount of negative feedback in the stage to balance the positive feedback resulting from the effects of C13 and the internal capacitive feedback of TR2.

The amplified i.f. signal is demodulated by the diode envelope detector (CR1), followed by two stages of a.f. amplification (TR3, TR4) which can conveniently be OC71 or almost any similar a.f. transistor. Control of the a.f. gain is governed by adjustment of VR2 which varies the amount of negative feedback in the final stage.

Low-impedance headphones are connected directly in the collector circuit of TR4, and although this may appear bad practice (in general the direct current component should not flow through a good pair of headphones since this can cause changes in magnetization), in this case the collector current is too small to change appreciably the d.c. magnetic field. Although only two tuned circuits are used in the i.f. section, selectivity is aided by the use of a regenerative stage, and shows considerable improvement over that of a simple straight receiver.

Sensitivity is limited by the absence of an r.f. stage though this will be less important on the lower frequency bands where external noise is high; with only one tuned signal frequency circuit—damped by the aerial—there will inevitably be image responses, particularly on higher frequencies. Nevertheless, a surprisingly large number of amateur phone and c.w. signals can be received satisfactorily, particularly when used with a reasonably high aerial some 30 ft. to 60 ft.

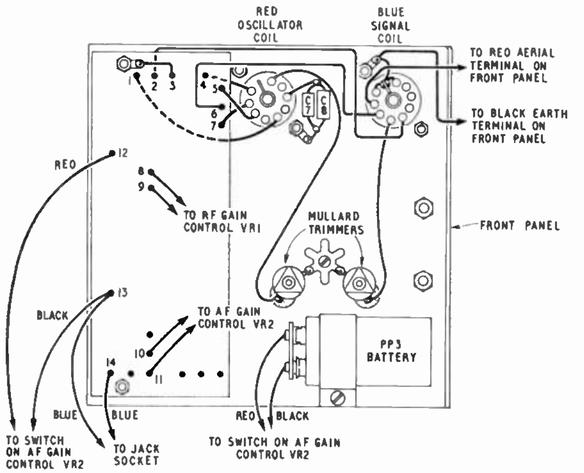


Fig. 4.77. Main components on the vertical chassis.

TABLE 4.4

Parts List for the RSGB Transistor Four—Fig. 4.75

Capacitors

Two	2 µF sub-miniature electrolytics (15V rating)
One	8 µF sub-miniature electrolytic (15 V rating)
Two	50 µF sub-miniature electrolytics (15V rating)
One	250 µF sub-miniature electrolytic (15V rating)
Two	0·005 µF paper (150V rating)
One	100 pF tubular ceramic
One	15 pF tubular ceramic
One	100 pF Suflex polystyrene (350V, 5 per cent rating)
One	1000 pF Suflex polystyrene (500V, 2 per cent rating)

Resistors

All Radiospares type R5, $\frac{1}{2}$-watt, 10 per cent rating.
One 470 ohms; two 1K; one 3·3K; one 3·9K; one 5·6K; one 8·2K; two 10K; one 12K; one 18K; one 47K; two 56K; one 100K.

Miscellaneous

The following parts can be obtained from Home Radio of Mitcham

One		12 yard reel Z8A connecting wire, any colour.
One		6 yard reel Z8B connecting wire, any colour (miniature).
One	VC12	310 pF plus 310 pF, two gang condenser.
One	DL50A	epicyclic drive, with flange.
One	—	Maxi-Q Coil, transistor tuning, range 3T, red. (Denco (Clacton) Ltd.).
One	—	Maxi-Q Coil, transistor tuning, range 3T, blue. (Denco (Clacton) Ltd.).
One	VR25	100 ohm wirewound potentiometer, 1 watt.
One	VR82	3 K ohm log potentiometer, with d.p. switch.
One	JH6	Igranic P72 c/c jack socket.
One	PK4	Clix Socket, red.
One	PK4	Clix Socket, black.
Two	VC62	Mullard trimmers, 3 — 30 pF.
Two	VH3	B9A valveholders, nylon (used as coil holders).
One	TRC72	Weyrad I.F. transformer, P50/3CC.
One	TRC73	Weyrad I.F. transformer, P50/2CC.
Two	—	$\frac{3}{4}$ in. dia. black fluted knobs.
One	—	$1\frac{3}{8}$ in. dia. black fluted knobs.
One	—	PP3 battery.
One	LK141	Chassis plate No. 4, Lektrokit.
One	LK2311	Bracket No. 1. Lektrokit.
50	LK3011	Pins soldering, Lektrokit.
Three	Z40	Grommets to fit $\frac{1}{4}$ in. holes.
One	—	Battery connector for PP3.

Nuts and bolts

Three	$\frac{3}{8}$ in. long	4 BA ch. hd cad steel screws.
Three	$\frac{1}{4}$ in.	4 BA ch. hd cad steel screws.
Two	$\frac{3}{8}$ in.	6 BA ch. hd cad steel screws.
Twelve	$\frac{1}{4}$ in.	6 BA ch. hd cad steel screws.
Two	$\frac{3}{16}$ in.	8 BA csk. cad steel screws.
Three		2 BA cad steel half-nuts.
Six		4 BA cad steel nuts.
Twenty-four		6 BA cad steel nuts.
Six		6 BA soldering tags.
One		4 BA star tag.
Six		6 BA cad. steel washers.

Transistors

Two	OC170	(TR1 and TR2)
Two	OC71 or OC72 or AC107 or GET880	(TR3 and TR4).

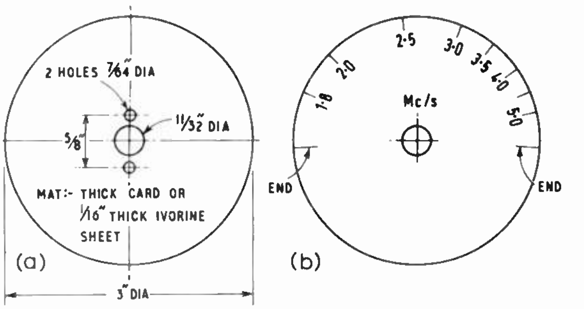


Fig. 4.78. Tuning scale and approximate calibration for the 1·8 to 5 Mc/s range.

long; an earth is preferable, particularly for reception on lower frequencies, but it is not essential.

The incorporation of a regenerative i.f. stage means that when this is set just above the threshold of oscillation c.w. stations and (with very careful tuning) s.s.b. stations can be received.

The receiver is constructed in the form of a front panel, consisting of a bent sheet of 18 s.w.g. aluminium, with a similar piece of aluminium to form a vertical chassis. Three

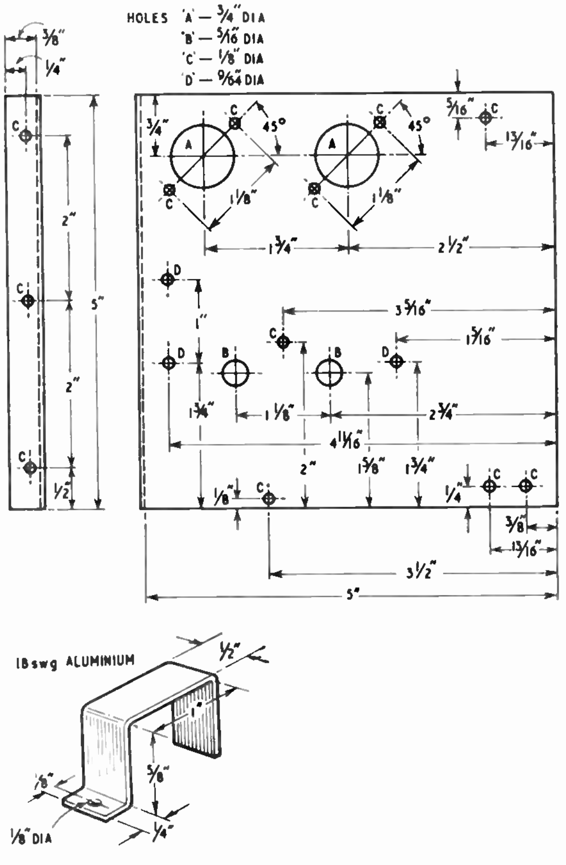


Fig. 4.79. Drilling diagram for the vertical chassis. (Left) the battery clip.

rubber grommets are used as " feet." Most of the detailed wiring is carried out on half of a Letrokit (LK141) perforated wiring board.

Since transistors can be damaged or destroyed by excessive heat or by incorrect polarity of the supplies, it is necessary to be most careful not to leave a hot soldering iron in contact with the transistor leads longer than absolutely necessary, and to check all wiring including the battery clip connector and switch. Do not clip the battery connector on to the battery until all wiring has been completed and carefully checked against the diagrams. Care should also be taken not to allow the hot soldering iron to come into contact with the casing of the polystyrene capacitors.

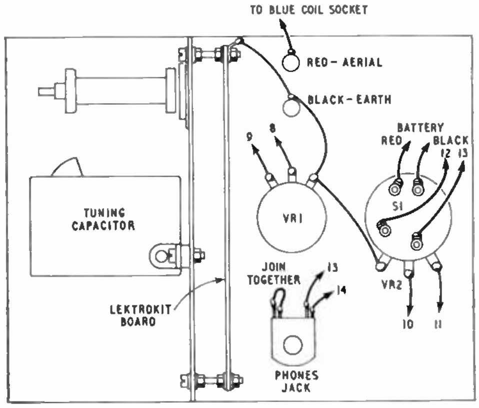


Fig. 4.80. Main components on front panel.

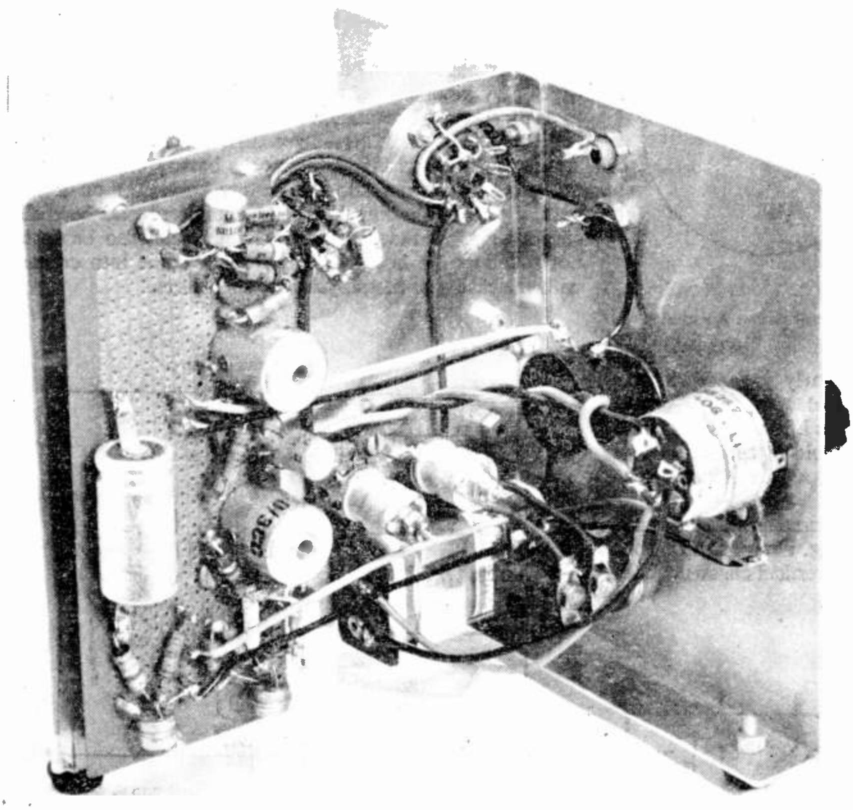
Adjustment and Alignment

When the receiver has been completed, and all wiring most carefully checked, plug-in the Maxi-Q coils (3T red and 3T blue), the blue coil being inserted in the socket nearer the front panel. Turn both coil core-adjusting screws until about ⅛ in. protrudes, and set both the Mullard trimmers to half-mesh. Plug in a pair of low resistance headphones (for example type DLR), connect the battery clip, turn the a.f. gain control fully clockwise, the r.f. gain fully anti-clockwise and switch on.

A faint hiss should be heard in the 'phones. Advance the r.f. gain and the i.f. stage should go into oscillation with a louder hiss. If this does not happen, advance the r.f. gain fully clockwise and adjust the core in one (either) i.f. transformer, first in one direction and then the other using a nylon trimming tool or chisel-pointed matchstick for this operation; do not use a metal screwdriver since this may crack the core. When the i.f. stage does oscillate, alternately adjust the i.f. transformer core and reduce the r.f. gain until finally the i.f. stage oscillates at the lowest possible setting of the r.f. gain control.

Should a signal generator be available, the receiver input circuits may be aligned and the dial calibrated, using the following procedure:

Turn the tuning knob clockwise until the gang capacitor vanes are completely meshed; set the signal generator to provide about 50 microvolts output (i.e. fairly weak signal) at a frequency of 1·62 Mc/s, feeding this into the aerial and earth terminals of the receiver. Turn the receiver's r.f. gain



Main assembly of Transistor Four.

until the i.f. just oscillates and then adjust the red coil core until the signal is received.

Then turn the tuning control anti-clockwise until the ganged capacitor vanes are completely out of mesh, tune the signal generator to 5·25 Mc/s and adjust the Mullard trimmer on the red (oscillator) coil side until the signal is again heard. If you now retune the set and generator to 1·62 Mc/s the setting may no longer be quite correct. Repeat the two adjustments at 1·62 Mc/s and 5·25 Mc/s until both ends are right at the same time. Then tune in the 1·62 Mc/s signal and adjust the blue coil core for optimum signal strength; and adjust the Mullard trimmer associated with the blue coil for maximum signal strength at 5·25 Mc/s. At this stage the receiver will be aligned and accurate calibration marks can be made on the dial by tuning the signal generator to 5·0, 4·5, 4·0 Mc/s, etc. Indian ink or a ball-point pen may be used to mark the dial.

Where (as will often be the case) no signal generator is available, the set can still be aligned quite easily without one. If the coil cores and trimming capacitors have been set as suggested earlier in this section, the receiver will probably not be far out of alignment. The scale of Fig. 4.78 can be copied on to the dial and should prove reasonably accurate. Then with an aerial plugged in, turn the tuning control clockwise until the vanes are meshed and then adjust the core of the blue coil for loudest signals at this end of the wave-range; then turn the tuning knob until the capacitor vanes are almost completely out of mesh, and adjust the Mullard trimmer associated with the blue coil unit signals are loudest; repeat these adjustments until optimum results over the band are obtained.

Frequency calibration may be accurately checked by listening for the frequency standard station MSF on 2·5 Mc/s and 5·0 Mc/s (this usually transmits a carrier modulated with distinctive one second " ticks ") another useful identification signal is the strong " Loran " navigational signals which can be heard at night transmitting a characteristic " buzzing " signal occupying a broad channel on about 1·95 Mc/s. Amateur stations should, of course, be heard between 1·8 and 2·0 Mc/s and also between 3·5 and 3·8 Mc/s, as well as many small ships transmitting phone in the range 2·0 to about 3·0 Mc/s.

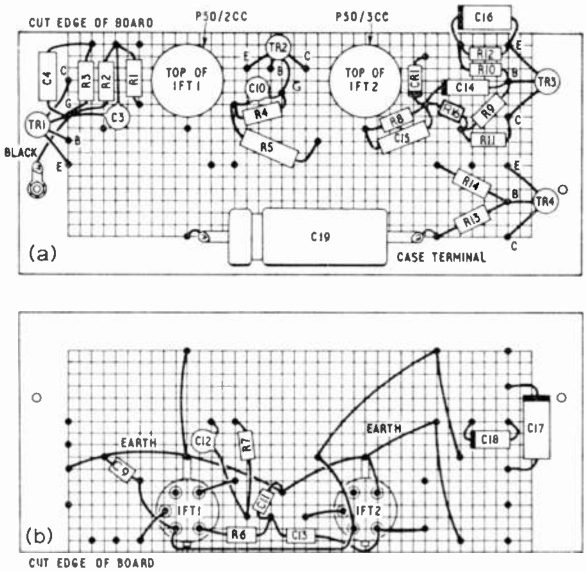


Fig. 4.81. Layout of perforated wiring board (a) top; (b) below.

LOW-COST HIGH STABILITY CONVERTER

Many older communications receivers provide adequate performance on the lower frequency bands but lack stability and/or sensitivity and good bandspread tuning on the higher frequency bands. One satisfactory answer to this problem, often providing an additional useful lease of life to be given to receivers acquired second-hand or as government surplus, is to use the basic receiver as a tunable i.f. system on, say, 3·5 Mc/s in conjunction with a sensitive crystal-controlled front-end converter for bands above 3·5 Mc/s.

Many suitable designs can be evolved from the circuits presented earlier in this chapter: basically such a converter will usually comprise an r.f. amplifier (possibly omitted if a low-noise mixer is used) mixer, crystal-controlled oscillator and possibly a cathode-follower to facilitate low-impedance coupling to the main receiver. Most published designs have used valves, but there is clearly scope for the development of transistor units for this application, provided care is taken to minimize cross-modulation problems.

Most circuits require a separate crystal for each band, or alternatively reverse the receiver tuning direction on one or more bands (for example by using a 10·5 Mc/s crystal to provide 3·5 Mc/s output on 7 Mc/s and on 14 Mc/s, or a 17·5 Mc/s crystal frequency for 14 and 21 Mc/s). However an economical and simple converter capable of good performance using only a single 3500 kc/s crystal has been described by F. Johnson, ZL2AMJ—see Fig. 4.82.

This two-valve unit functions as a crystal-controlled converter on the 7, 14 and 21 Mc/s amateur bands and as a straight pre-amplifier on 3·5 Mc/s. The converter provides a broad-band output on the 3·5 Mc/s band, and all tuning is carried out on the main receiver. It should prove particularly effective in conjunction with an older receiver having good bandspread tuning and stability on the 3·5 Mc/s band. The selectivity will depend entirely upon that of the main receiver when operating on 3·5 Mc/s. One disadvantage is that the unit produces a strong spurious marker signal on 7000 kc/s which can block the first few kilocycles of the 7 Mc/s band, this can be minimised or eliminated if the crystal is slightly lower in frequency than 3500 kc/s as described later.

The converter comprises the triode section of a 6U8 (ECF82) triode-pentode working as a grounded-grid amplifier. The pentode section of the functions as mixer with a low noise contribution. One half of a 12AT7 (ECC81) functions as crystal oscillator and the other section as a cathode follower output stage following the mixer. There are no tuned circuits at the broadband i.f. of about 3·5–4 Mc/s, consequently no ganged tuning circuits are used.

For 3·5 Mc/s reception the oscillator is switched off and the remaining stages act as a pre-amplifier. On other bands the crystal oscillator frequency is always 3·5 Mc/s lower than the band being received, so that the signal frequency rises as the main receiver is tuned higher across the 3·5 Mc/s band. For 7 Mc/s the crystal oscillates on its fundamental frequency of 3500 kc/s, for 14 Mc/s on its third "overtone" of about 10·5 Mc/s and for 21 Mc/s on its fifth "overtone" of 17·5 Mc/s. Note that for overtone operation the output is not precisely at the exact harmonic of the crystal fundamental, although close to it, so that a small correction will be needed to the original receiver calibration. Otherwise the 3·5 Mc/s calibration applies in terms of tens and hundreds on all bands, and the tuning rate will be the same on all bands. Thus 7·1, 14·1 or 21·1 Mc/s signals should be received on about 3·6 Mc/s and 7·2, 14·2 or 21·2 Mc/s on 3·7 Mc/s, etc.

In order to maintain this approximate 3·5 Mc/s calibration on all the other bands, the crystal needs to be fairly accurate on 3500 kc/s. However if this is not regarded as vital, it is often possible to use a crystal some kc/s lower in frequency. For example a 3490 kc/s crystal would tune 7000–7300 kc/s as 3510–3810 kc/s, 14,000–14,350 kc/s as 3530–3880 kc/s and 21,000–21,450 kc/s as approximately 3550–4000 kc/s. This tolerance would allow the use of a cheap surplus crystal provided that it is sufficiently active to oscillate readily on its fifth overtone.

Note that only two adjustable coils are used to tune the

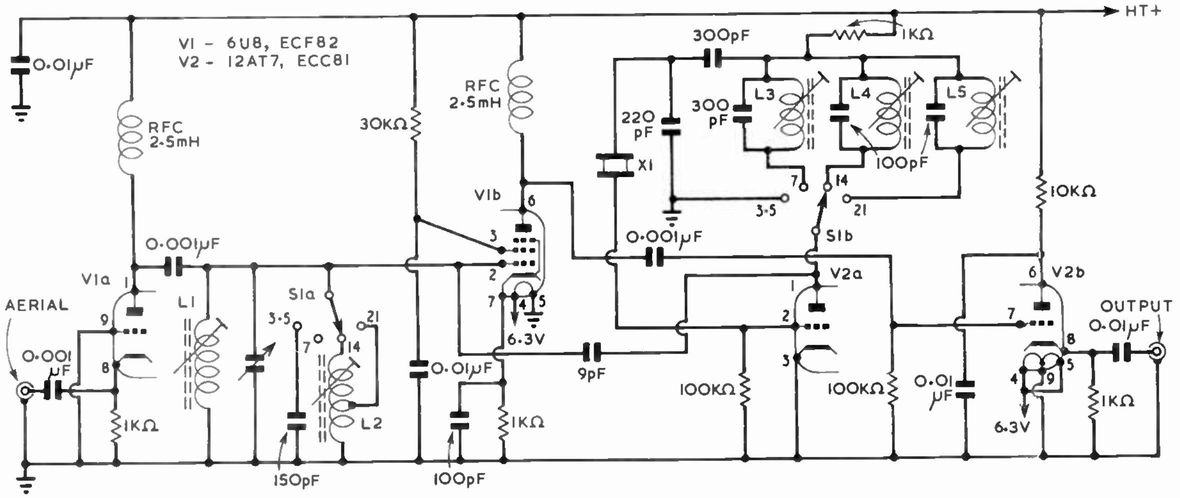


Fig. 4.82. Circuit diagram of the high stability converter.

four bands with L1 left in parallel with L2 on 14 and 21 Mc/s bands.

Construction and layout are not particularly critical provided that r.f. leads are kept short, and output circuits placed well away from input circuits.

All coils are wound on ¼ in. diameter slug-tuned unshielded formers using 30 s.w.g. enamelled wire: L1 42 turns; L2 26 turns, tapped at 17 turns from the "earthy" end; L3 35 turns; L4 13 turns; L5 8 turns, spaced over ⅜ in. Formers could be 0·3 in. diameter with a slight reduction in numbers of turns—this can easily be determined if a calibrated grid dip oscillator is available. The formers can be mounted around the Yaxley-type bandswitch.

Alignment

After completing and checking all wiring, the oscillator circuits must be adjusted so that the crystal oscillates on the correct overtone frequencies. This can be done most readily with a grid-dip oscillator used as an absorption wavemeter, though it should be possible to achieve satisfactory results by checking the output on the main receiver if this will tune to 10·5 and 17·5 Mc/s and care is taken not to be misled by harmonic output. With the absorption device check and adjust the cores of L3, L4 and L5 until the oscillator is operating on the required frequencies.

Then set the converter to 7 Mc/s and connect it to the main receiver which is set to 3·5 Mc/s. Without switching on adjust L1 for coverage of 7–7·5 Mc/s in conjunction with the 35 pF tuning capacitor by using the grid dip oscillator in the usual way (this can be done with incoming signals without a g.d.o. but the process is more difficult). Switch to 3·5 Mc/s and check that the r.f. tuning range now covers 3·5–4·0 Mc/s (or just the European section of the band—3·5-3·8 Mc/s—if this is all that is required). If the tuning range is incorrect, it may be necessary to replace or parallel the 150pF tuning capacitor with other values. Once the 3·5 Mc/s tuning range has been established, switch to 14 Mc/s and adjust L2 so that the tuning range covers 14–14·5 Mc/s. It should then be possible to tune the 21 Mc/s band without any further adjustment (the full tuning range will probably exceed the amateur band but this is of no consequence). In practice it will probably prove necessary to change the converter tuning only when tuning right across the band; when searching around only a small section of the band the control can be left unaltered.

Once the tuning ranges have been adjusted, the unit should be ready for use. Signals can be peaked with the converter tuning control when necessary, otherwise all tuning is done on the main receiver. If the gain of the h.f. converter should be excessive on some signals a gain control can be fitted on the cathode follower stage.

One possible difficulty is that there may be i.f. breakthrough from strong local 3·5 Mc/s signals when receiving on the other bands. Screening of the converter and the use of coaxial cable with screened connectors between converter and receiver may be sufficient to overcome any difficulties. Should this not be the case, it may in some cases be necessary to fit a further tuned circuit in the form of an aerial tuning unit between the aerial and converter. Provided that the main receiver is efficient on 3·5 Mc/s this combination should provide excellent sensitivity and stability on all the bands concerned.

Details of a rather more advanced crystal-controlled converter (6BZ6 tuned r.f. amplifier; ECC85 low-noise triode mixer; and ECC85 Butler crystal oscillator with separate crystals for each band) are given in the RSGB publication *Communication Receivers*.

THE Q MULTIPLIER

A useful method of improving the selectivity of receivers is by using a *Q multiplier*. As already explained this is a circuit usually constructed as an outboard unit by which the selectivity characteristics of a conventional i.f. section may be appreciably improved without modification of the receiver itself. When used for peaking, the device is most useful during the reception of c.w. signals when it can be tuned to select any one of several signals on slightly different frequencies within the normal passband. When used for rejection the tunable notch can remove a troublesome carrier without any noticeable effect on the desired signal.

Fig. 4.83 shows a practical circuit developed by G2BVN. When S1 is turned to the "PEAK" position the tuned circuit (L2, C2, C3 and C4) is in parallel with the receiver i.f. coil to which the unit is connected and the *Q* of the combination is then raised by the positive feedback around V1b. The 5K variable resistor in the cathode of V1b controls the amount of regeneration and therefore the selectivity. In the "REJECT" position of S1 the tuned circuit is connected from the grid of V1a to earth and at resonance, because of the high impedance of the tuned circuit, a voltage at the anode of V1a is fed also to the grid. This causes the apparent anode impedance of V1a to be very low and severely damps the i.f. coil across which it is connected. At frequencies off resonance the tuned circuit impedance is comparatively low so that very little of the voltage at V1a anode is fed to the grid. In the absence of this negative feedback the anode impedance of V1a is quite high and very little damping occurs. The width of the rejection notch is controlled by the effective *Q* of the tuned circuit. This is governed by the degree of regeneration and is adjustable using the other 5K variable resistor in V1b cathode circuit.

An advantage of this form of *Q multiplier* is that the peak or notch can be tuned across the passband of the receiver allowing signals to be selected without critical adjustment of the main receiver tuning. The unit is simple to construct and align and can be installed without altering the receiver wiring.

The heart of the unit is the coil L2 which uses a ferrite pot core construction and has an inductance between 120 and 150 microhenries for intermediate frequencies in the range 450 to 470 kc/s. Suitable pot type coils are available from Electroniques (Standard Telephones and Cables Ltd., proprietors), Harlow, Essex. To retain the high *Q* of this coil it should be mounted at least 1½ inches away from any ferrous metal. The capacitors C3 and C4 should be close tolerance, good quality, silver mica types. Coil and capacitor values for other intermediate frequencies are shown in **Table 4.5.**

The coil L1 is used to tune out the capacitive reactance of

TABLE 4.5

I.F.	L1	L2	C3	C4
85 kc/s	15-60 mH	0·5-2·5 mH	2500 pF	7500 pF
475 kc/s	1·5-3·0 mH	120-150 μH	1000 pF	3000 pF
735 kc/s	750-1000 μH	70-100 μH	750 pF	2250 pF
915 kc/s	250-500 μH	60-90 μH	500 pF	1500 pF
1600 kc/s	50-120 μH	40-60 μH	250 pF	750 pF

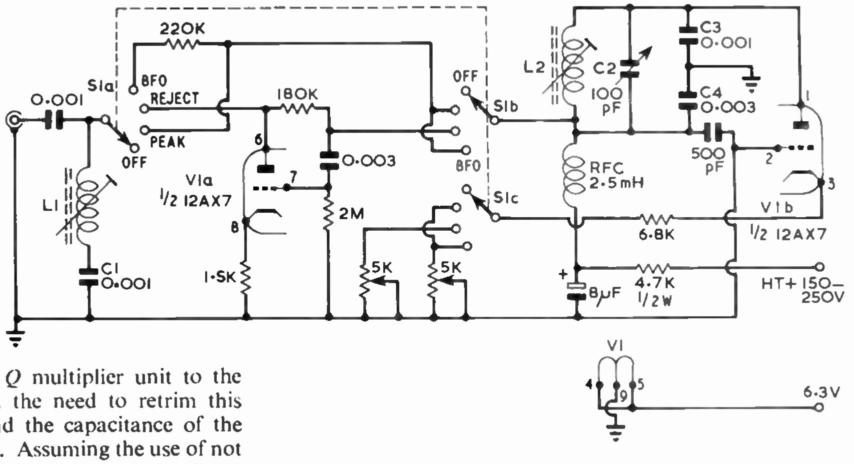


Fig. 4.83. Circuit diagram of the Q multiplier. The fixed capacitors, with the exception of C3 and C4, can be of the disc ceramic type. C3 and C4 should be close tolerance silvered mica. Resistors, unless otherwise indicated, can be of ¼ watt rating. The values of C3 and C4 are for an intermediate frequency of 450-470 kc/s. Values for other frequencies are given in Table 4.5. The function switch is a three pole four way Yaxley type.

the coaxial cable, connecting the *Q* multiplier unit to the receiver i.f. transformer, to avoid the need to retrim this circuit. The inductance of L1 and the capacitance of the cable should tune to the receiver i.f. Assuming the use of not more than 30 in. of good quality semi-airspaced or cellular coaxial cable, L1 will need to have an inductance of 1·5 to 3 millihenrys. L1, C1 may be omitted if there is no objection to retrimming the receiver i.f. transformer. Although the coaxial cable may be connected to any of the i.f. transformers in the receiver, it is advantageous to connect it to the first i.f. transformer as this will reduce the chances of blocking or cross modulation.

With the receiver on and the *Q* multiplier connected, turn S1 to OFF and tune in a steady signal, making sure it is correctly centred in the i.f. bandpass. Adjust the core of L1 for the highest S meter reading or maximum a.f. output. If the coil does not peak then the connecting cable will have to be shortened or lengthened. If the highest signal strength is obtained with the coil fully out this would suggest the cable is too long. Once correctly set this adjustment will not have to be changed.

To adjust L2 turn S1 to PEAK, the selectivity control to maximum resistance and C2 to mid-travel. Adjust L2 slug until the steady signal is peaked, then slowly rotate the selectivity control, re-peaking the slug when necessary. As the control is advanced the signal level should rise and the peak will become sharper until a point is reached where the circuit will break into oscillation. The point of maximum selectivity will be just below this point. With the *Q* multiplier peaked, the tuning may be varied to boost any signal within the i.f. bandpass and attenuate other frequencies. At maximum selectivity telephony stations should be unintelligible and it will be necessary to back off the control slightly. Should the circuit not go into oscillation, the cathode resistor (6·8K ohms) should be reduced, but not more than absolutely necessary. Too low a value may indicate that the *Q* of L2 is low. The cathode connection of the resistor should be made directly to pin 3 on the valve base.

To obtain best results from the REJECT position, some practice is necessary as the adjustment is critical. A steady carrier should be tuned in and the b.f.o. adjusted to give a beat note of about 100 c/s. C2 and the reject potentiometer should then be alternately adjusted until the best null is obtained, at which point the tuning should be found to be extremely sharp.

SIMPLE Q MULTIPLIER

A simpler circuit for a *Q* multiplier for selectivity peaking only (that is, with no rejection notch facilities) and not requiring any additional high-*Q* coils has been described by W3FYG. This arrangement can be incorporated in most receivers where it is desired to increase selectivity, particularly for c.w. reception: see **Fig. 4.84**.

The circuit uses a single additional double triode (e.g. 12AT7 or 12AX7) as a multi-vibrator type oscillator using the existing i.f. transformer in the receiver as the tuned circuit. The device is most effective with single conversion receivers of only moderate selectivity, such as those depending entirely upon i.f. transformers on 465 kc/s or 1600 kc/s or above to shape their response.

Coarse (R2) and fine (R1) controls are fitted for the control of cathode coupling and thus govern regeneration. Initially R1 is set to minimum resistance and R2 adjusted until the valve just goes into oscillation; then R1 is backed off and used as the panel control. Remember when fitting the device that the additional capacitance across the i.f. transformer will require slight adjustment of the core or trimmer of this tuned circuit after installation.

A possible difficulty is that there may be a tendency for the subsequent i.f. stage to go into oscillation. This can be overcome, as shown in **Fig. 4.84**, by the removal of the bypass

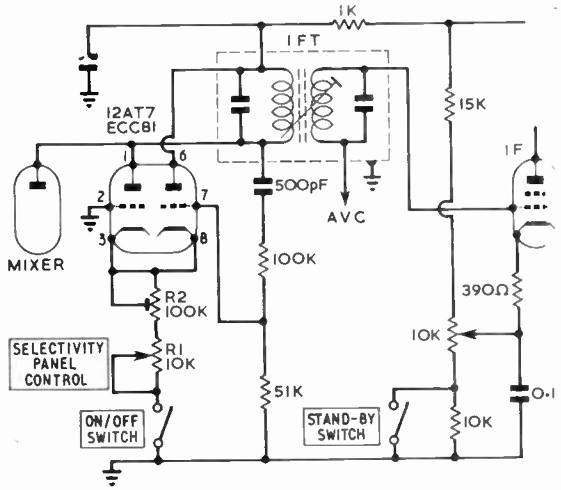


Fig. 4.84. Q-Multiplier without additional coils. The double triode valve is a 12AT7.

capacitor from across this valve's cathode-bias resistor; the slight loss of gain is more than compensated for by the beneficial effect of the Q multiplication.

STENODE RECEPTION FILTER

The single-crystal filter, when functioning in maximum selectivity conditions, is generally regarded as providing a degree of selectivity which is too sharp for satisfactory a.m. or s.s.b. phone reception, since the speech appears to become woolly and often almost completely unintelligible due to the loss of the high-frequency speech components.

This concept is in contradiction to the original use of this form of filter for the "stenode radiostat" receiver of the early 'thirties where a crystal filter of the same basic design

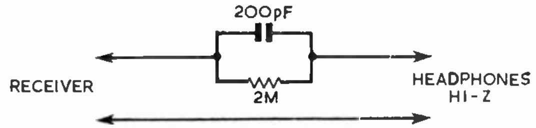


Fig. 4.85 Stenode tone-correction filter for phone reception through a crystal or Q-multiplier filter.

as that shown earlier in this chapter was combined with *tone correction* to permit the reception of speech and music. The stenode principle is to provide suitable treble-rise tone correction in the a.f. circuits to compensate for the treble-cut introduced by the sharp selectivity characteristics of a crystal filter. The bandwidth of signals passing through this type of filter—often only about 200 c/s for 6db down unless the filter is damped by a high impedance output—will normally render speech difficult to follow; but since the slope of the filter is by no means vertical, a high degree of tone correction (about 6db/octave) will restore the speech. This feature of the original Robinson stenode principle has been revived by L. A. Moxon, G6XN, who has further shown that it has the benefit of reducing selective fading of a.m. signals by reducing modulation percentages at higher audio frequencies (providing a form of exalted carrier reception) and by partial suppression of one sideband.

The simple tone-correction filter of Fig. 4.85 can be inserted between the normal envelope detector and first a.f. amplifier (or even directly in the headphone lead) of an existing receiver, and provides the necessary treble-rise needed to match reasonably closely the treble-loss of a crystal filter. This technique can provide a most economical system of obtaining good adjacent channel selectivity for phone reception.

AMATEUR-BANDS-ONLY DOUBLE SUPERHET

There are many advantages to be gained by restricting the coverage of a home-built receiver to the amateur bands only. The receiver described in the following pages was evolved and built by A. J. Shepherd, G3RKK in order to provide good performance at moderate cost and to meet the need for a receiver which can be built easily with very few tools and capable of alignment with a minimum of test equipment.

It is basically a high-sensitivity double-superhet of the conventional type (i.e. fixed first and second intermediate frequencies) capable of good reception of a.m. and s.s.b. phone signals and c.w. telegraphy. It is most suitable for construction by enthusiasts with reasonable experience of home-construction, but who do not feel entirely confident of tackling the more complex type of tunable first i.f. receiver such as that described later in this chapter. A list of the components is given in **Table 4.6.**

It should be noted that details of a later, modified version of this receiver have been published in the RSGB BULLETIN reflecting the designer's continuing search for improved results in several directions, including reduced oscillator drift on the higher frequency bands and a considerable number of circuit changes. These various modifications can

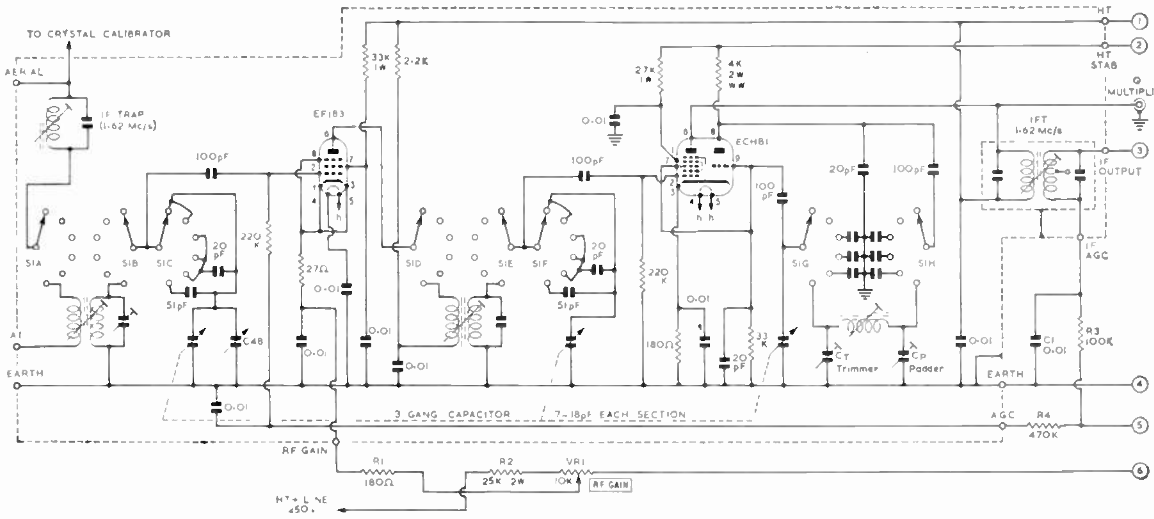


Fig. 4.86. The front-end of the receiver (Electroniques type QP166). The bandswitch is shown in the 1.8 Mc/s position. Only the coils and trimmers for 28 Mc/s are shown; the others occupy corresponding positions on the other bandswitch contacts. The oscillator switch section SIC is fitted with shorting contacts.

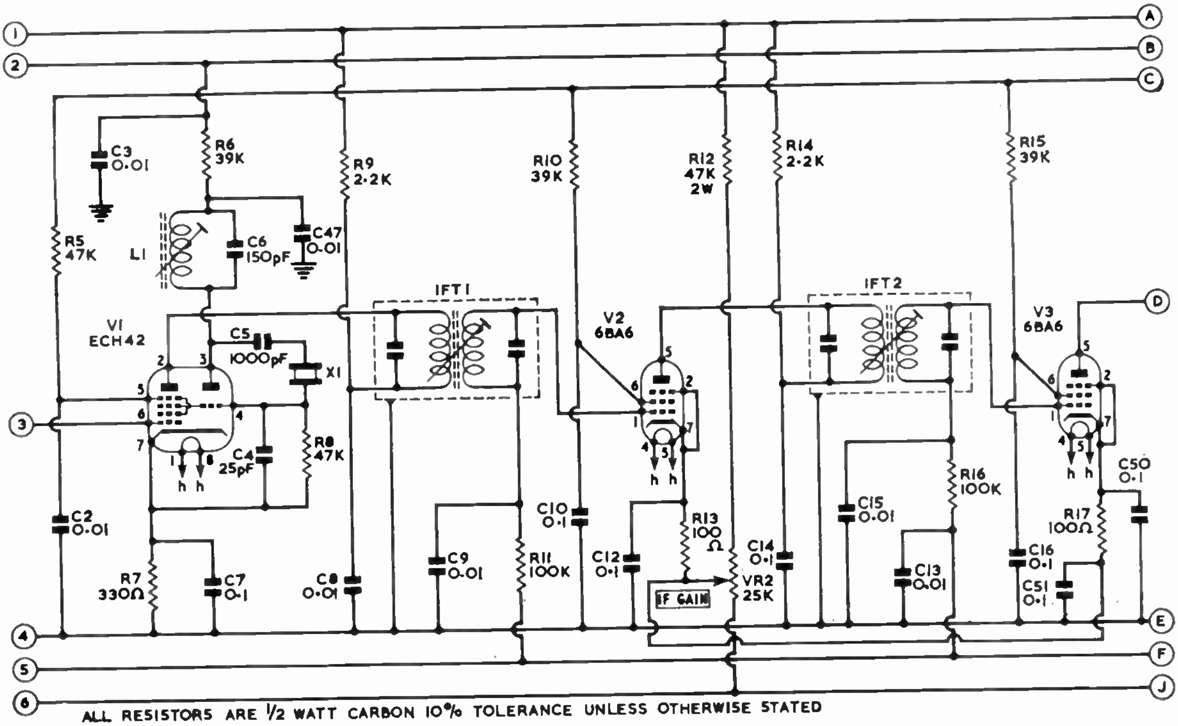


Fig. 4.87. The second frequency changer and 85 kc/s i.f. strip.

result in improvements but involve rather greater complication in construction; it is therefore proposed to base this section on the original design which has been successfully duplicated by many amateurs.

The key to the simplicity of the construction and alignment lies in the use of a commercial front-end supplied pre-aligned. A suitable unit is available from Electroniques (Standard Telephones and Cables Ltd., proprietors), Harlow, Essex. This provides about 170° of bandspread coverage on six amateur bands: Band 1 (160 metres) 1·8-2 Mc/s; Band 2 (80 metres) 3·4-4 Mc/s; Band 3 (40 metres) 7-7·5 Mc/s; Band 4 (20 metres) 14-14·4 Mc/s; Band 5 (15 metres) 21-21·5 Mc/s; Band 6 (10 metres) 28-30 Mc/s.

It will be noted that the width of the 3·5 Mc/s band covers the full allocation available to American amateurs, of which only 3·5-3·8 Mc/s are available to European amateurs. The availability of a 2 Mc/s wide band (Band 6) means that the receiver can be used with a 144 Mc/s crystal-controlled converter to give full coverage of the European 144-146 Mc/s (2 metres) band.

Ease of tuning is achieved by the use of a high grade tuning dial having a reduction ratio of 110:1. The tuning rate is better than 10 kc/s per knob revolution, except on Band 6 where it is about 40 kc/s per knob revolution.

The front-end unit comprising the signal frequency and local oscillator tuning circuits, r.f. amplifier, first frequency converter and associated circuits is supplied pre-aligned to provide an output signal on the first i.f. of 1620 kc/s with a high degree of image interference rejection and with a parallel-tuned i.f. trap in the aerial input circuit to reduce breakthrough of strong signals on or near the first i.f.

The second i.f. of 85 kc/s makes it possible to obtain a high degree of selectivity (about 3 kc/s bandwidth for −20db reduction on peak) without using crystals or a mechanical filter. Response of the a.f. circuits is restricted to about 400—3000 c/s, the low frequency cutting helping to offset the effect on a.m. signals of the highly selective i.f. response. A better shape factor could be achieved with a half-lattice crystal filter but this would tend to make the receiver more difficult for newcomers not having test equipment. For c.w. reception, an external 1620 kc/s *Q*-multiplier is a useful addition (and is built-in for the Mark 2 version); a suitable socket is provided for connecting a *Q*-multiplier across the first i.f. transformer to provide a controllable degree of extra selectivity or a sharp rejection notch.

A product detector is fitted for use on c.w. and s.s.b. with a diode integrator envelope detector for a.m. signals. The receiver has a peak noise limiter to remove audio peaks, a fast and slow acting a.g.c. system, an S-meter and a built-in 100 kc/s crystal calibrator for providing check points throughout its tuning range.

Frequency stability depends to a considerable extent upon the mechanical construction. With rigid bracing and suitable ventilation, drift can be kept low. Once the set has fully warmed up, frequency drift may be less than 500 c/s per hour, even on 30 Mc/s, and considerably less on the lower frequency bands. The use of silicon diode rectifiers with their low operating temperatures helps keep cabinet temperatures down. A voltage regulator tube provides a stabilized 150 volt line for supplying stages which could be affected by supply variations.

Basically the receiver has an EF183 (6EH7) high-slope tuned r.f. amplifier stage and ECH81 (6AJ8) first frequency converter in the factory-built front-end. The 1620 kc/s signal from this sub-unit goes to V1 (ECH42, 6CU7), the second frequency changer with crystal-controlled oscillator. V2 (6BA6) and V3 (6BA6) are both 85 kc/s i.f. amplifiers. V4 (6AT6) is a diode a.g.c. rectifier with delay, and a.f. amplifier. V5 (6AM5) is the audio output stage. V6 (6BA6) is the 100 kc/s calibrator and V7 (6AM6) the high-stability beat frequency oscillator covering about 84–86 kc/s. V8 (12AU7) is the product detector. The crystal diode CR1 (OA81) is used to protect the S-meter (M1) from flow of reverse current. CR2, 3 (OA79) form the voltage-doubling diode integrator detector; CR4, 5 (OA81) is an adjustable peak noise clipper.

Values shown for R25 and T2 are suitable for use with 6AM5 output valve which provides about 0·5 watt. For higher output power a 6BW6 or EL84 is suggested with modification of these components.

The crystal (X1) used in conjunction with V1 is 1537 kc/s, though since the 85 kc/s i.f. transformers can be tuned over the range 80–90 kc/s, the crystal can be anywhere in the range 1530–1540 kc/s without changing the first i.f. from the factory figure of 1620 kc/s. Suitable crystals are usually available in the "surplus" market. The tuned circuit in the anode circuit of this stage is resonated to the crystal fundamental frequency to reduce harmonic content. A standby sensitivity control VR5, allows the receiver to be used to monitor strong signals from the transmitter.

If a crystal calibrator is already available, V6 and its

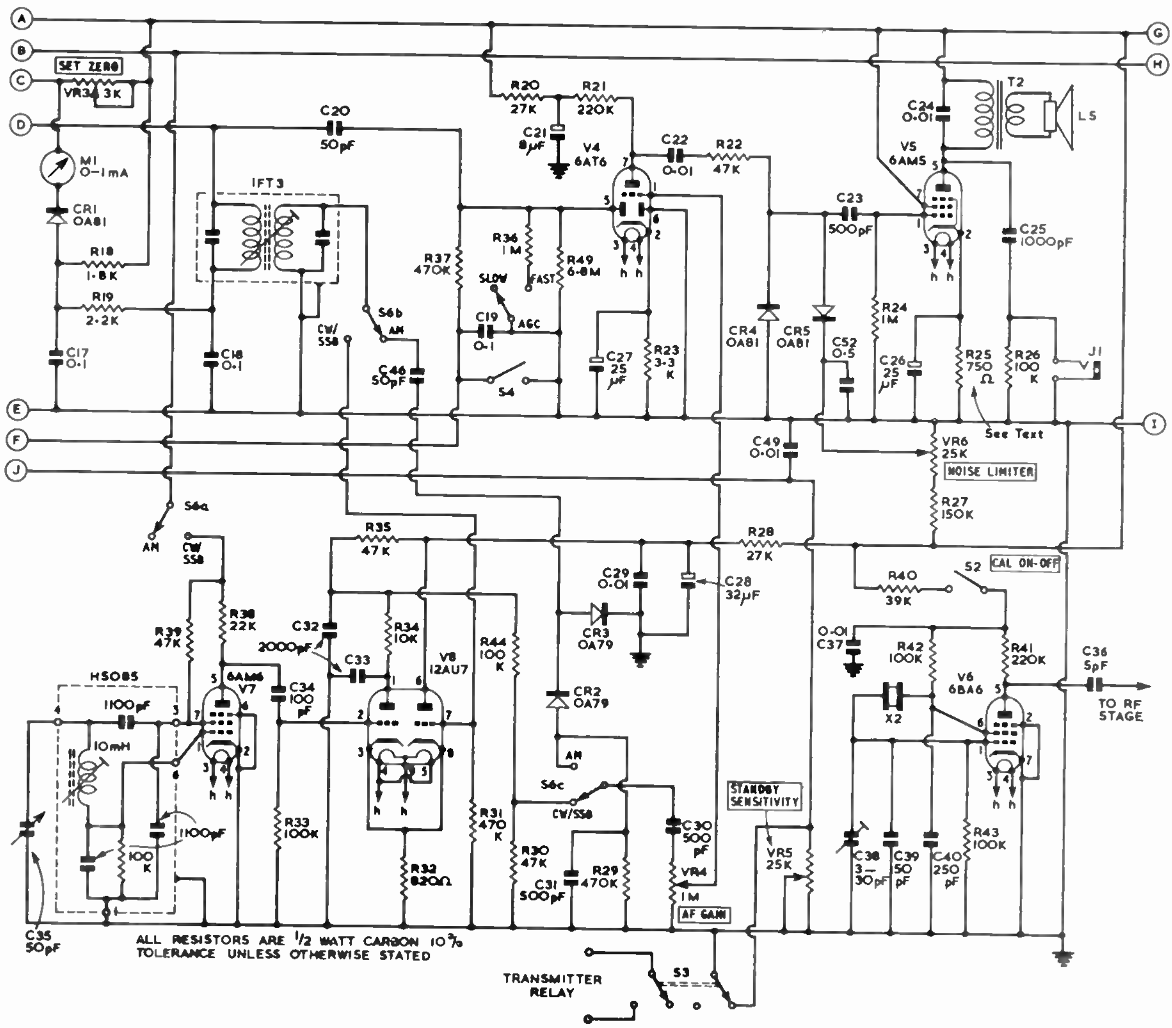


Fig. 4.88. The S meter, crystal calibrator, a.g.c., detectors and a.f. sections. C35 is the b.f.o. pitch control.

associated components could be omitted, but otherwise this will prove a most useful accessory. For highest accuracy the second harmonic of the 100 kc/s crystal may be zero beat by adjusting C38 against the 200 kc/s *BBC* Light Programme high stability transmitter while using a separate broadcast receiver.

Construction

The layout is shown in **Fig. 4.89**, this differs slightly from the photographs which were taken at an earlier stage of development. It is recommended that 16 s.w.g. aluminium is used for the chassis work, which should be well braced.

Care should be taken throughout construction to ensure that components and wiring are as rigid as possible, especially any affecting the frequency stability. Ventilation should be arranged to keep the operating temperature low, without erratic air currents passing through the oscillators. The front-end unit should be well braced in the $6\frac{1}{16}$ in. square cut-out. It may be found worth fitting a screen over the oscillator section of the front-end. The crystal-controlled oscillator of V1 should be well screened to minimize spurious responses. The b.f.o. should be constructed using 16 s.w.g. tinned copper wire for stability, and this also should preferably be screened.

Screening cans fitted to all valves should be painted matt black to assist heat radiation. Low loss skirted valveholders of either the ceramic or nylon loaded types should be used. A B7G valveholder is fitted to the rear of the chassis and

The receiver in its cabinet showing the arrangement of the controls

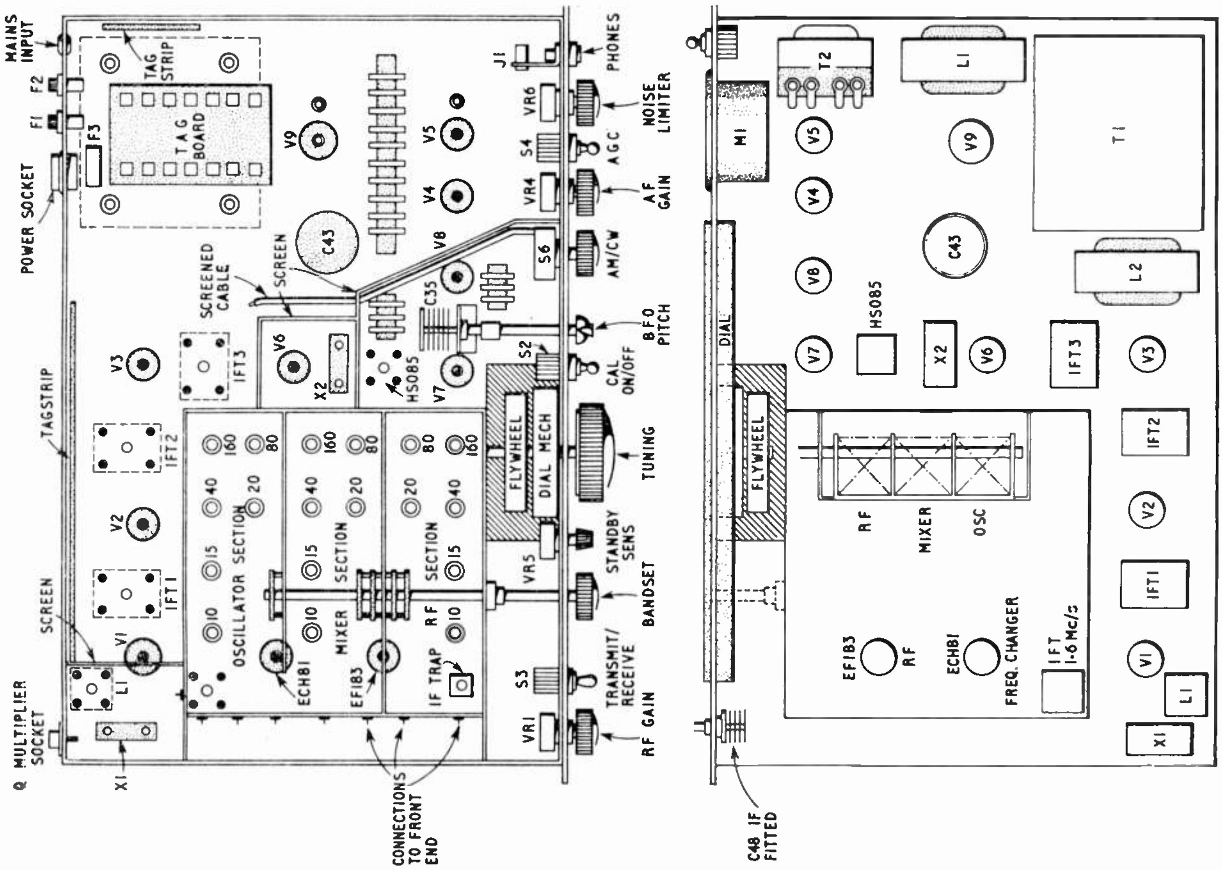
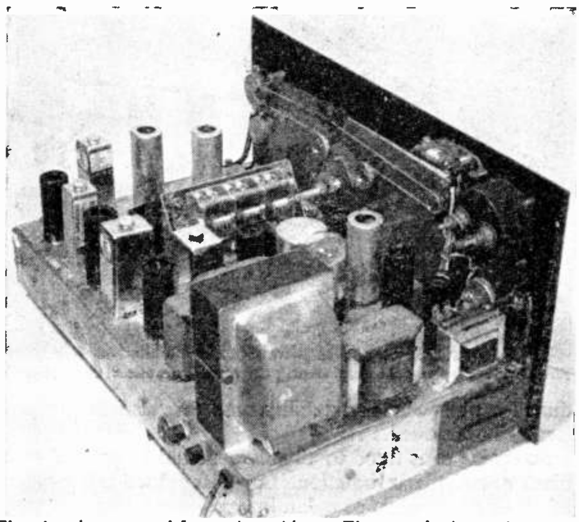


Fig. 4.89. The layout of the receiver showing the positions of the major components.



The chassis removed from the cabinet. The terminals on the rear drop of the chassis are for the loudspeaker and transmitter relay.

supplies power for a Q multiplier or other external units. Disc ceramic 0·01 μF capacitors are used extensively because of their small size, low inductance and relatively high working voltages. With paper capacitors the plastic case type is to be preferred to those with waxed paper containers. Any long wires carrying i.f. or a.f. signals should be screened. All a.g.c. and h.t. line connections are made via the long tagstrip mounted along the rear of the i.f. stages.

Alignment

Alignment is relatively simple, owing to the use of the pre-aligned front-end. Several methods are possible, but the original model was readily aligned by the following method.

For alignment a test signal within 5 kc/s of 1620 kc/s is required. If no accurate signal generator is available, the test signal is best obtained by replacing the 100 kc/s crystal of V6 by one of 1620 kc/s or a sub-harmonic of the frequency. A readily obtainable "surplus" crystal is the FT241A 54th harmonic type marked " 29·1 Mc/s (Channel 19)." This has a fundamental of 405·5 kc/s and will usually provide a fourth harmonic on about 1622 kc/s.

The output from V6 is temporarily removed from the aerial socket, and taken to the "I.F. OUT" lead-through on the front-end, via a 1000 pF capacitor. Calculate what will be

TABLE 4.6

COMPONENTS LIST

B.F.O. Unit, H5O85 (Electroniques (Harlow) Ltd.).
C1, 2, 3, 8, 9, 13, 15, 22, 24, 29, 37, 41, 42, 44, 47, 49, 0·01 μF, 500 volt wkg., disc ceramic.
C4, 25pF silver mica.
C5, 1000pF silver mica.
C6, 150pF ceramic.
C7, 10, 12, 19, 50, 0·1μF paper.
C11, 14, 16, 17, 18, 0·1μF 400 volt wkg., paper.
C20, 39, 46, 50pF silver mica.
C21, 8μF, 400 volt wkg., electrolytic.
C23, 30, 31, 500pF ceramic.
C25, 1000pF ceramic.
C26, 27, 25μF 25 volt wkg., electrolytic.
C28, 32μF 400 volt wkg., electrolytic.
C32, 33, 2000pF ceramic.
C34, 100 pF silver mica.
C35, 50pF variable (J.B. type C804).
C36, 5pF silver mica.
C38, 30pF trimmer.
C40, 250pF ceramic.
C43, 24 + 16μF 450 volt wkg, electrolytic.
C48, 4-13pF (Eddystone type 588), aerial trimmer if required.
C52, 0·5μF.
CR1, 4, 5, OA81.
CR2, 3, OA79.
Cabinet, 16 in. x. 10½ in. x 8 in. (Philpotts' Metalworks).
Chassis, 15 in. x 10 in. x 2½ in., 16 s.w.g. aluminium.
Dial and Drive, Eddystone type 898.
Front End, QP166 Amateur Bands Bandspread Qoilpax (Electroniques (Harlow) Ltd.).
F1, F2, 1 amp fuses.
F3, 250mA fuse.
IFT1, 2, 85 kc/s type DIF1 Series II (Electroniques (Harlow) Ltd.).
IFT3, 85 kc/s type DIF1/D Series II (Electroniques (Harlow) Ltd.).
L1, 50-75μH type DLM14 (Electroniques (Harlow) Ltd.).
L2, 3, 10H 150 mA.
M1, 0-1mA moving coil meter.
MR1, 2, 1000 p.i.v., 500 mA silicon diodes.
R1, 180 ohms.
R2, 25 K ohms 2 watts.
R3, 11, 16, 26, 33, 42, 43, 44, 100 K ohms.
R4, 29, 31, 37, 470 K ohms.
R5, 8, 22, 30, 35, 39, 47 K ohms.
R6, 10, 15, 40, 39 K ohms.
R7, 330 ohms.
R9, 14, 19, 2·2 K ohms.
R12, 47 K ohms 2 watts.
R13, 100 ohms.
R17, 100 ohms.
R18, 1·8 K ohms.
R20, 28, 27 K ohms.
R21, 41, 220 K ohms.
R23, 3·3 K ohms.
R24, 36, 1 Megohm.
R25, 750 ohms (see text).
R27, 150 K ohms.
R32, 820 ohms.
R34, 10 K ohms.
R38, 22 K ohms.
R45, 3·9 K ohms 4 watts.
R46, 100 K ohms 2 watts.
R49, 6·8 Megohms.

All resistors are ½ watt carbon unless otherwise stated.

RFC, 2·5mH 50mA.
S1, bandswitch incorporated in coil pack.
S2, s.p.s.t. toggle (calibrator on/off).
S3, d.p.d.t. toggle (transmit/receive).
S4, s.p.s.t. toggle (a.g.c. on/off).
S5, d.p.d.t. heavy duty toggle (mains on/off).
S6, 3 way 3 pole Yaxley, only 2 positions used (a.m./s.s.b./c.w.).
S7, s.p.s.t. toggle (a.g.c. time constant—slow/fast).
T1, 250-0-250 volts, 150mA; 6·3 volts, 3·5 amps.
T2, output transformer to suit valve (For EL91, 6000 ohms/3 ohms, 2 watt type).
V1, ECH42.
V2, 3, 6, 6BA6 (EF93).
V4, 6AT6, EBC90.
V5, 6AM5, EL91 (see text).
V7, 6AM6, EF91.
V8, 12AU7, ECC82.
V9, VR150/30 (OD3).
VR1, 10 K ohms (r.f. gain).
VR2, 25 K ohms (i.f. gain).
VR3, 3 K ohms (Radiospares pre-set for S meter set zero).
VR4, 1 Megohm (a.f. gain).
VR5, 25 K ohms (Radiospares pre-set for adjusting standby sensitivity).
VR6, 25 K ohms (noise limiter).
X1 (see text).
X2, 100 kc/s crystal.
Other components required are fuseholders, knobs, pilot lamp, plug, sockets, tag board, tag-strips, valveholders, screening cans, flexible coupler, and grommets.

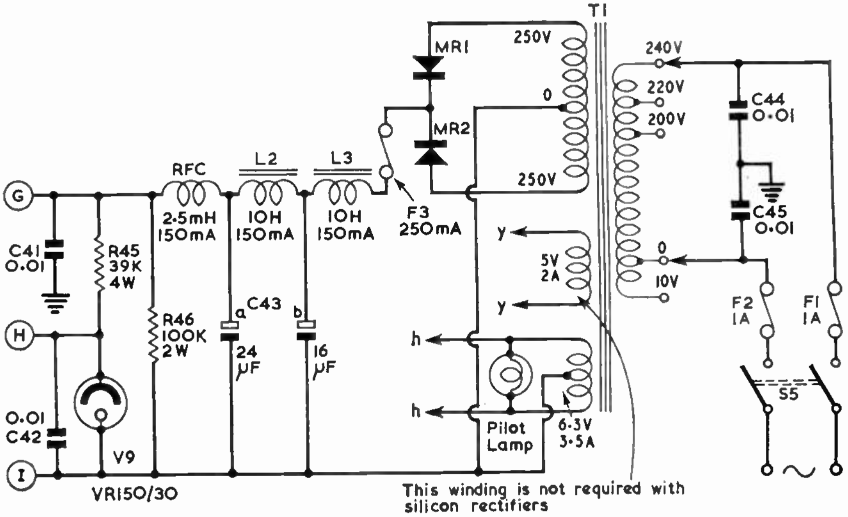


Fig. 4.90. The power supply.

the second i.f. (i.e. 1622 kc/s minus the conversion crystal frequency).

Now switch the receiver on, and allow it to warm up thoroughly. Switch on the crystal calibrator. Turn the i.f. gain to maximum; a.g.c. switched on and the S-meter zero-control turned to give quarter scale deflection on S-meter. Should a small a.f. oscillator be available it can be used to modulate V6 crystal output so that an audible check is available, but this is not essential. Remove the two front-end valves.

The i.f. transformers are preset to 85 kc/s on the outer resonance. If the calculated second i.f. is less than 85 kc/s tune cores slightly inwards; if greater than 85 kc/s tune the cores slightly outwards. Adjust each core only by half a turn at a time, starting with IFT 1 and work towards the detector. Repeat the entire sequence with a further half-turn adjustment and so on until S-meter reading begins to rise. Then peak each core for maximum reading, readjusting the i.f. gain control and set zero control as the alignment proceeds so that the S-meter reading is kept reasonably low.

Reconnect the crystal calibrator to its normal position and re-insert 100 kc/s crystal. Set to Band 1, replace the front-end valves and allow them to warm up thoroughly, with aerial still disconnected. With gain controls at maximum, a kick in the S-meter reading should be observed as the receiver is tuned over the centre of the band; this should be 1900 kc/s and should be carefully tuned for peak S-meter reading. At this stage all the i.f. circuits should be finally peaked up, the b.f.o. switched on and, with the pitch control set at mid-scale, adjusted for zero beat.

The aerial may then be connected and the trimmers on the r.f. coils (front compartment of the front-end) carefully peaked at the centre of the band, with the aerial trimmer set to mid-scale. This can be done using external signals, or for example a signal from a transmitter v.f.o. Work carefully, and avoid making large haphazard changes.

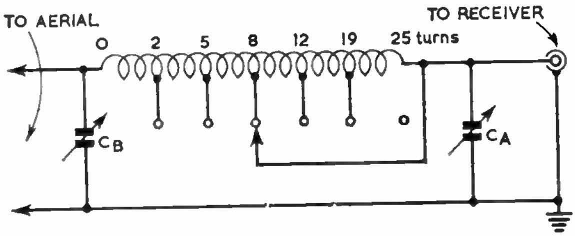


Fig. 4.91. An aerial tuning unit. C_A should be 500pF, C_B will depend upon the type of aerial in use, but 300pF is a good basis for experiment. The coil consists of 25 turns on a 1 in. diameter former, tapped at 2, 5, 8, 12 and 19 turns and occupying a space of 1 in. 28 s.w.g. wire is suitable.

The aerial connection will depend upon the type of feeder to be used:

75 ohms balanced input: terminals A and A1;

75 ohms unbalanced (e.g. coaxial feeder): terminals A and E with A1 strapped to E.

Single-ended: terminal A with A1 strapped to E.

To provide an accurate impedance match, it is possible to use an aerial tuning unit of the type shown in Fig. 4.63.

Stability

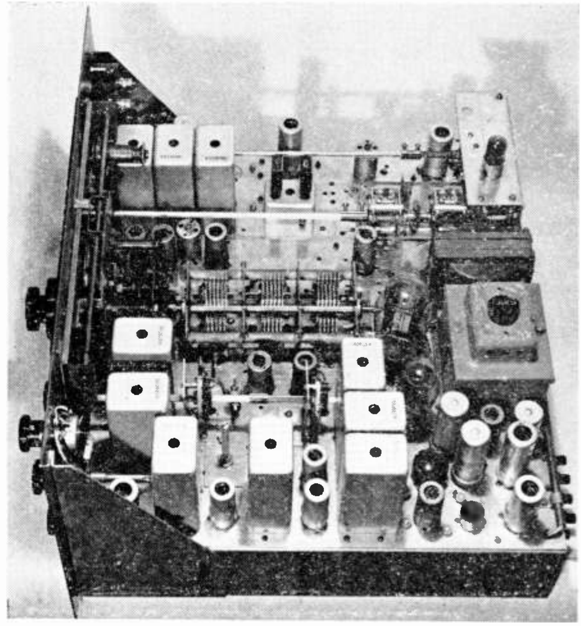
Probably the greatest problem facing the constructor will be that of achieving maximum frequency stability. If there are any sudden changes in frequency, the receiver oscillators should be carefully inspected. If the same amount of shift occurs on all bands the trouble is probably in the crystal conversion oscillator or the b.f.o. If the shift is greater on the higher frequency bands suspect the front-end.

Note that the front-end is fitted with negative temperature coefficient capacitors in the oscillator circuit to provide thermal compensation. For optimum stability this thermal compensation must be in accord with the normal operating temperature, and this will be affected by the ventilation. It may therefore be necessary to adjust the ventilation, after carefully checking stability on a stable signal in the 21 to 28 Mc/s band (allow receiver to warm up for at least 30 minutes). Adjust the b.f.o. for zero beat.

If the tuning control capacitance has to be consistently increased (plates further in mesh) the oscillator is drifting higher in frequency, suggesting that there may be excessive ventilation, and vice versa. In the prototype model optimum stability was achieved with the lid of the cabinet open and a few suitably placed ventilation holes drilled in the chassis.

When operating, the setting of the gain control knobs will affect performance For best signal-to-noise ratio, the r.f. gain should be maximum; a.f. gain about half-travel and overall gain kept under control by altering i.f. gain control. If very strong local stations cause cross-modulation or blocking, the r.f. gain should be reduced, and i.f. and a.f. gain increased.

For reception of s.s.b., the b.f.o. pitch control should be set to a point near the extremity of its travel, corresponding to upper or lower sideband reception. The main control is then rotated until the speech becomes intelligible, possibly with a slight final adjustment of the b.f.o. It should not be necessary to switch off the a.g.c. except on very weak signals but a.g.c. should be switched to the longer time constant. Ample b.f.o. injection voltage should be available, but on very strong s.s.b. signals it may be necessary to reduce the i.f. gain slightly to avoid distortion.



The advanced amateur bands receiver. The components may be identified by reference to Fig. 4.94.

ADVANCED RECEIVER WITH TUNABLE FIRST I.F.

The increasing popularity of s.s.b. transmissions has focused attention on the shortcomings of communications receivers which previously would have been considered highly satisfactory. The reception techniques developed for s.s.b. can also be applied with advantage to normal a.m. and c.w. signals. For example, where the receiver is made sufficiently selective it can be tuned to one sideband of a conventional double sideband a.m. signal and reject both the carrier and the second set of sidebands; the signals can then be treated as though they were s.s.b. transmissions. This system can be arranged to provide immediate selection of the upper or lower set of sidebands, permits a significant increase in selectivity to be used without affecting a.f. response as well as reducing phase distortion and selective fading. A receiver having good s.s.b. performance is also likely to provide excellent results on c.w., giving true single signal reception combined with high stability and a low tuning rate.

The requirements of optimum stability and a low, linear tuning rate on all bands call for the use of a crystal-controlled h.f. oscillator in conjunction with a tunable first i.f. section. For a receiver of this type to be free of spurious responses careful selection must be given to the frequency coverage of the tunable section, to screening and to construction generally.

The block diagram of an advanced amateur receiver meeting the above needs, developed by G. R. B. Thornley, G2DAF, is shown in **Fig. 4.92**. The design considerations will not be dealt with here, but the circuit and performance specifications are given to show in practice the main features of a high-performance receiver of modern design.

The use of a crystal-controlled h.f. oscillator eliminates the problems of tracking front-end oscillator circuits and simplifies construction and final alignment. Nevertheless, it should be appreciated that the construction and alignment of a receiver of this type cannot be undertaken lightly; for the average constructor the work involved is likely to take up to six months or more of leisure time.

The receiver has 20 valves (including rectifiers and voltage regulator tube) and covers seven bands, each 500 kc/s wide: *1*, 1·5–2·0 Mc/s; *2*, 3·5–4·0 Mc/s; *3*, 7·0–7·5 Mc/s; *4*, 14·0–14·5 Mc/s; *5*, 21·0–21·5 Mc/s; *6*, 28·0–28·5 Mc/s; *7*, 28·5–29·0 Mc/s. Provision is made for two extra 500 kc/s bands to cover completely the 28·0–29·7 Mc/s band.

Altogether 16 quartz crystals are incorporated: seven for the h.f. oscillator, to provide switched bands; six in three half-lattice bandpass crystal filters centred on 460·8 kc/s with pairs of crystals spaced 2·2 kc/s apart, to provide the main selectivity; two for the carrier insertion oscillator, 1·5 kc/s above and below the centre i.f., to provide selectable sideband reception; and a 100 kc/s calibration oscillator.

For all bands, the tunable i.f. covers 5·0–5·5 Mc/s and the v.f.o., which is the second conversion oscillator, operates on a single, non-switched band of 5460–5960 kc/s; at this

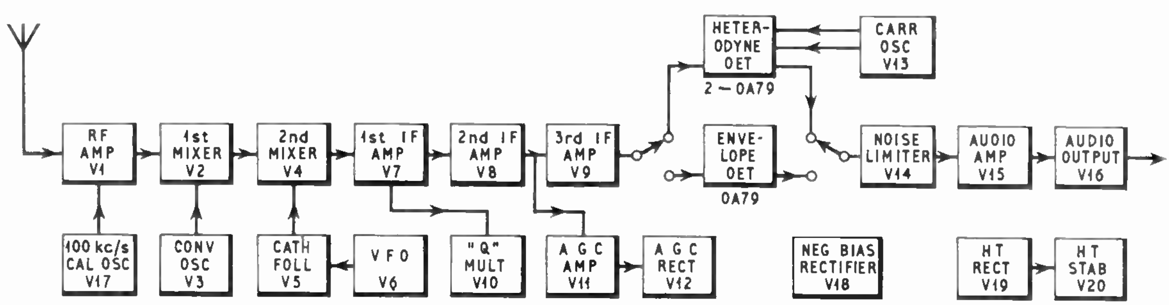


Fig. 4.92. Block diagram of an advanced amateur communications receiver with tunable first i.f.

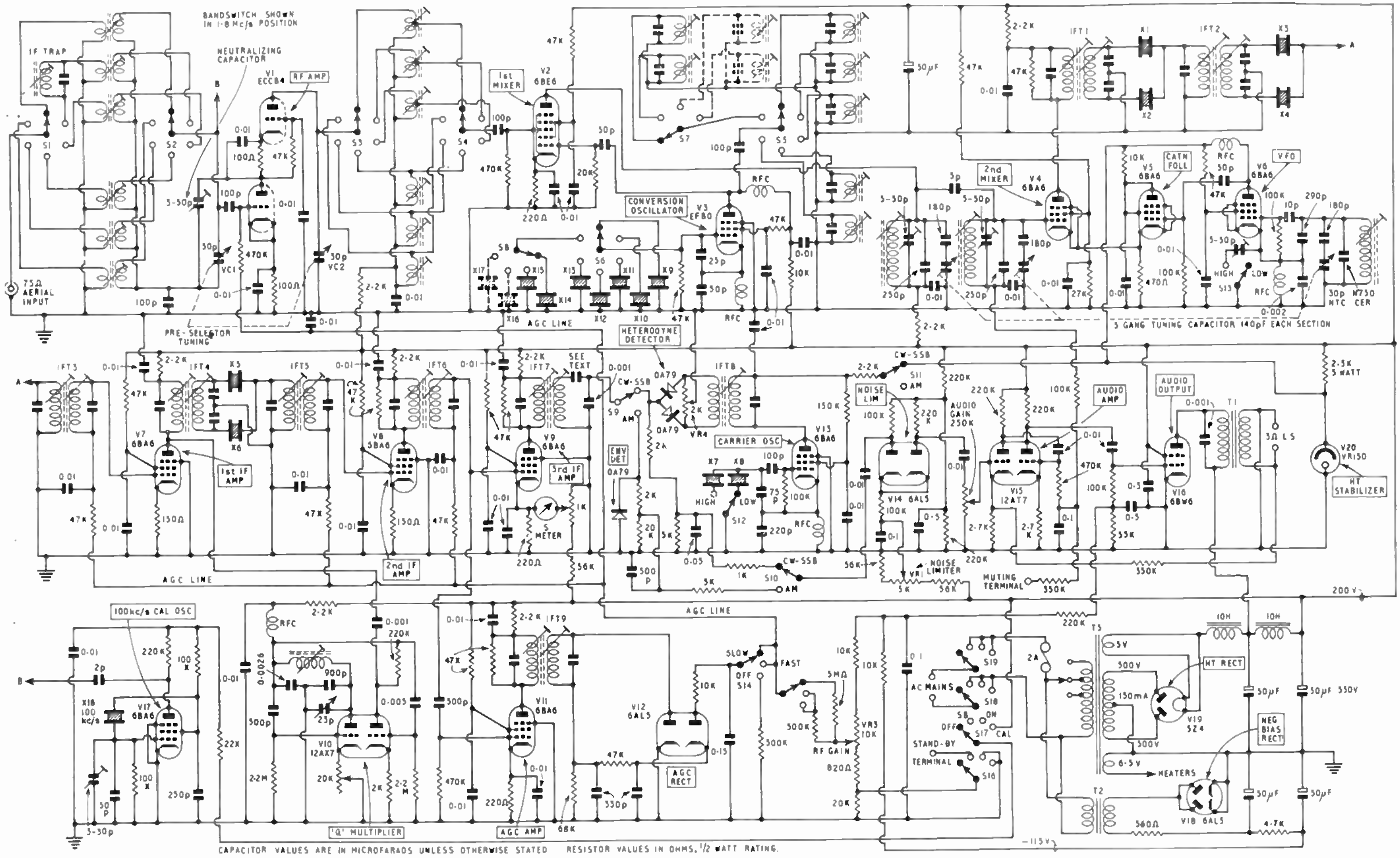


Fig. 4.93, Circuit diagram of the advanced double conversion amateur bands receiver with tunable first i.f.

S1, 2, 3, 5, 6, Yaxley type ceramic bandswitch, 6 bank 1 pole 6 way (shown fully anti-clockwise in the 1·5-2 Mc/s position).
S7, 8, Yaxley ceramic bandswitch (10m band selector), 2 bank 1 pole 4 way.
S9, 10, 11, Yaxley type paxolin (demodulator selector), 1 bank 3 pole 2 way.
S12, 13, Yaxley type ceramic switch (sideband selector), 2 bank 1 pole 2 way.
S14, S15, Yaxley type paxolin (a.g.c. switch), 1 bank 2 pole 3 way.
S16, S17, Yaxley type paxolin, 1 bank 2 pole 4 way.
S18, S19, Yaxley type paxolin, 1 bank 1 pole 4 way.
(S16, S17, S18 and S19 are the operational switch).
X1, X3, X5, 461·9 kc/s (series resonant frequency).
X2, X4, X6, 459·7 kc/s (series resonant frequency).
X7, 459·3 kc/s (parallel resonant frequency in situ).
X8, 462·3 kc/s (parallel resonant frequency in situ).
X9, 7000 kc/s.
X10, 9000 kc/s.
X11, 6250 kc/s.
X12, 6500 kc/s.
X13, 6625 kc/s.
X14, 8375 kc/s.
X15, 8500 kc/s.
RFC, all 2·5mH
T1, audio output transformer (36 : 1).
T3, 300-0-300V 150mA, 5V 2A, 6·3V 5A.
T2m 13 volt heater transformer with 200-250V primary, 13V winding connected across 6·3V line.
All 0·01 μF bypass capacitors are 400V d.c. working.

frequency high stability can be achieved, particularly as the v.f.o. is isolated from the second mixer by a cathode follower stage. Since the overall stability of the receiver will be determined largely by this oscillator, stability remains closely similar on all bands. A Q multiplier circuit is used to provide a tunable rejection notch.

The full circuit diagram is given in Fig. 4.93.

The aerial input coil on each band provides a match to 75 ohm feeder. As high Q r.f. circuits are required to minimize image response, the r.f. stage is neutralized. A cascode r.f. amplifier (V1, ECC84, 6CW7) is preferred to a conventional pentode circuit because of its superior cross-modulation characteristics. Signal frequency preselection is by the input coil in the grid circuit and the series fed coil in the anode circuit of V1. These are selected (together with the aerial input) by three banks of the main bandchange selector switch and pre-tuned for each band with the two ganged 50 pF tuning capacitors. The 160, 80 and 40m anode coils are tapped to provide a reasonably constant stage gain on all bands, with the fourth section of the bandswitch feeding the signal into the grid of the first mixer V2 (EK90, 6BE6). The necessary heterodyning frequency is obtained from a crystal controlled harmonic oscillator V3 (EF80, 6BX6). The required crystal and anode coil are selected by the remaining two banks of the six bank bandchange switch. As suitable types of switch wafers are normally six-way, the additional 500 kc/s sections of the 10m band are covered by an auxiliary two bank two pole four-way Yaxley type switch brought out to a separate panel control. This enables full coverage of the 10m band to be obtained if required.

The tunable i.f. section covering the 5·0 to 5·5 Mc/s range comprises two tuned circuits in the grid of the second mixer V4 (EF93, 6BA6) and in the grid of the v.f.o. V6 (EF93, 6BA6) arranged as a Colpitts ocsillator and tuned by a three gang variable capacitor of 140 pF each section. The v.f.o. output is via a cathode follower V5 (EF93, 6BA6) strapped as a triode, with direct injection into the cathode of the second mixer, V4. Automatic v.f.o. frequency correction when switching sidebands is obtained by a 3–30 pF trimmer capacitor (pre-set to the required value) and a selector switch in the v.f.o. cathode circuit.

The second i.f. output at 460 kc/s is fed into a two half-lattice crystal bandpass filter and to the first i.f. amplifier V7 (EF93, 6BA6) and then into a third half-lattice filter section and further amplified by the second and third i.f. stages V8, and V9 (both EF93, 6BA6). A Q multiplier notch filter V10 (ECC83, 12AX7) is fed from the anode of V7. The output of the third i.f. amplifier is switched to the OA79 balanced diode bridge c.w. and s.s.b. demodulator or to a single OA79 diode envelope detector for a.m. reception. The carrier insertion oscillator V13 (also type EF93, 6BA6) is crystal controlled, the required crystal being selected by S12 the single pole two-way SIDEBAND switch ganged to the v.f.o. correction switch S13.

The required audio output is selected by S10 and fed into the negative and positive peak clipping noise limiter V14 (EB91, 6AL5) and the level set by the potentiometer VR1 (NOISE LIMITER) control. Output is fed via VR2 the AUDIO GAIN control, to the audio amplifier V15 (ECC81, 12AT7) and then to the output valve V16 (6BW6). Negative voltage feedback is provided over all audio stages from the secondary of the output transformer. The necessary bias for the output valve is fed via a potential divider from the negative bias supply comprising the transformer T2 and the rectifier valve V18 (EB91, 6AL5). In the STAND BY position of the operational switch S16, 17, 18, 19 the bias is automatically increased and the output valve standing current reduced to approximately 20 mA to lower the loading on the mains transformer T3.

The i.f. input to the grid of the third i.f. amplifier (V9) also feeds the a.g.c. amplifier V11 (EF93, 6BA6), the output of which is fed to the shunt rectifier and gate diode V12 (EB91, 6AL5). Resultant current pulses at audio frequency charge the 0·15 μF reservoir capacitor. The charge time constant is fast—of the order of 0·01 second, but discharge can only take place via the AVC switch S14, 15, and either the 500K or 5 Megohm resistors. The discharge time constant is therefore slow—of the order of 0·1 and 1·0 second respectively.

The 10K ohm potentiometer VR3 is the RF GAIN control and applies bias from the negative supply simultaneously to the a.g.c. line and to the anode of the gate diode. This holds up the decay of the available a.g.c. voltage to a level pre-set by the gain control. Because of Miller effect, the varying a.g.c. voltage would affect the grid input capacity of the controlled valves and this would particularly affect the bandpass filter response characteristics, degrade the selectivity and affect the audio frequency response. This effect is overcome in the two associated i.f. amplifier valves V7 and V8, by providing negative r.f. feedback with unbypassed 150 ohm cathode resistors. The primaries of the two i.f. transformers in the anode circuits of V4 and V3 are damped with 47K ohm resistors to prevent filter ringing—sufficient damping of V7 anode circuit is provided by the Q multiplier input loading.

The main h.t. supply is from the mains transformer T3 and the rectifier valve V19 (GZ30, 5Z4), with choke input providing 220 volts to the anode of V16. The second stage smoothing provides an output of 200 volts to the main h.t. positive rail. A VR150/30, 150C3 voltage regulator (V20)

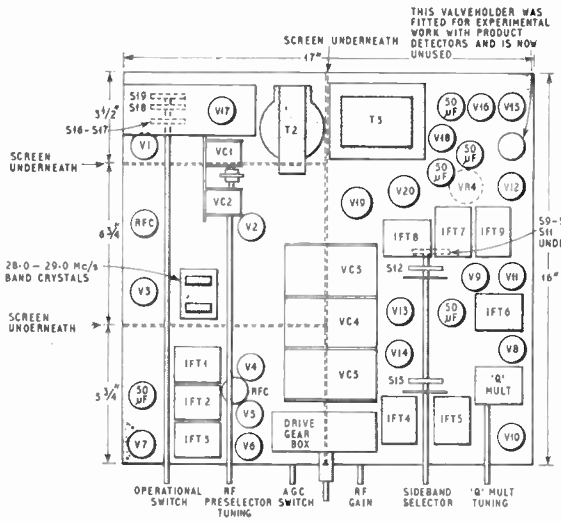


Fig. 4.94. Above chassis layout in the high performance double conversion amateur bands communication receiver. The dotted lines indicate the screening below the chassis.

provides 150 volts stabilized to feed the v.f.o. and cathode follower and the carrier insertion oscillator (V13).

The calibration oscillator V17 (EF91, 6AM6 or equivalent) provides harmonic output from a 100 kc/s crystal and gives accurate calibration pips every 100 kc/s throughout the receiver range. Operation is controlled by S17, part of the main control switch.

Construction

The receiver is built on a 16 s.w.g. aluminium chassis measuring 17 in. by 16 in. by 3½ in. deep. The 19 in. by 10½ in. panel is standard rack and panel size. The actual measurements are determined by the physical size of the components used, particularly the i.f. transformers and the three gang tuning capacitor. It is strongly recommended, however, that the chassis is not made smaller than suggested unless this is possible without reducing the signal frequency coil compartments. The *Q* of the coils and those in the tunable i.f. circuits can be reduced considerably by too close proximity to the cross screens and chassis sides. Close spacing of coils can also lead to undesirable mutual coupling and absorption effects. The above-chassis layout is shown in **Fig. 4.94.**

As the r.f. amplifier is neutralized by out of phase feedback from the grid input circuit, the frame of the grid section of the two gang preselection capacitor is above earth. Accordingly this component is made up with two separate 50 pF capacitors mounted on an insulated panel and ganged together with an Eddystone type 529 flexible shaft coupler.

The three i.f. transformers in the bandpass filter require capacitive centre taps. These are provided by removing the original capacitor across the winding and replacing it with two of double the value in series. The junction of the two capacitors is taken out to a separate lead-out wire as the centre tap of the winding. The secondary of IFT7, feeding the bridge diode c.w. and s.s.b. demodulator must be modified to low impedance series output by replacing the original capacitor with a 1000 pF unit and whatever value is required to resonate the coil in series, as shown in the circuit diagram. Low impedance output is required to feed the c.i.o. output to the OA79 bridge diode demodulator via IFT8. This is provided by removing one " pie " from a standard i.f. transformer and replacing it with a 50 turn scramble winding of 36 s.w.g. enamelled wire tightly coupled to the existing coil.

All variable potentiometers are standard types (wire wound below 100K ohms and carbon above 100K ohms) except for VR4 in the balanced diode bridge circuit; this is for r.f. balancing and must be non-inductive and therefore carbon.

If surplus FT241 crystals are used they are moved to the frequency required by grinding one edge. The crystals for the c.i.o. are positioned at the 20db points. It should be remembered that the actual oscillating frequency *in situ* will be between 100 and 200 c/s lower than the series resonant mode. A small amount of neutralizing capacity—2 pF—was used across two of the filter sections but this would not necessarily give the same response characteristics with some other type of i.f. transformers. Further advice on lattice filters of the type used in this receiver is given in Chapter 10 (*Single Sideband*).

The conversion oscillator valve (V3) is a type EF80, 6BX6 and is particularly suitable for this requirement. It

TABLE 4.8

Band Mc/s	Harmonic Conversion Oscillator Frequencies					
	Required	Crystal used	Second har. of crystal	Third har. of crystal	Fourth har. of crystal	Output
1·5-2·0	7000	7000	14,000	21,000	28,000	Fundamental
3·5-4·0	9000	9000	18,000	27,000	36,000	Fundamental
7·0-7·5	12,500	6250	12,500	18,750	25,000	Second Harmonic
14·0-14·5	19,500	6500	13,000	19,500	26,000	Third Harmonic
21·0-21·5	26,500	6625	13,250	19,875	26,500	Fourth Harmonic
28·0-28·5	33,500	8375	16,750	25,125	33,500	Fourth Harmonic
28·5-29·0	34,000	8500	17,000	25,000	34,000	Fourth Harmonic

TABLE 4.7

COIL WINDING DETAILS

Function	Freq. or band	Winding	Remarks
Signal Frequency V1 Grid	160m	100T sec. 7T prim 40 s.w.g. enam.	Paxolin former for 160, 80 and 40 metres. Polystyrene for 20, 15 and 10 metres. ⅜ in. diam. 2 in. long, ⅜ in. diam. with dust core. (Denco *Maxi-Q*)
	80m	55 T sec. 3T prim. 32 s.w.g. enam.	
	40m	30T sec. 2T prim. 24 s.w.g. enam.	
	20m	14T sec. 2T prim. 22 s.w.g. enam.	
	15m	9T sec. 1·5T prim. 22 s.w.g. enam.	
	10m	7T sec. 1T prim. 20 s.w.g. enam. (5paced 12T per inch). All primaries tightly coupled at cold end of secondary.	
Signal Frequency V1 Anode	160m	100T tap at 50T down, 40 s.w.g. enam.	Formers as above.
	80m	55T tap at 20T down, 32 s.w.g. enam.	
	40m	30T tap at 10T down, 24 s.w.g. enam.	
	20m	14T 22 s.w.g. enam.	
	15m	9T 22 s.w.g. enam.	
	10m	7T 20 s.w.g. enam. (Spaced 12T per inch.)	
Tunable I.F. V2 Anode & V4 Grid	5·0-5·5 Mc/s	14T 22 s.w.g. d.s.c. close wound paxolin former as sig. frequency coils.	Cement windings with Denfix polystyrene cement.
V.F.O. V6 Grid	5·46-5·96 Mc/s	13T 22 s.w.g. d.s.c. close wound paxolin former as sig. frequency coils.	
Conversion Oscillator V3 Anode	7·0 Mc/s	25T 32 s.w.g. enam. Shunt Cap = 50pF	Former: Aladdin ½ in. diam. 1 in. long, with 3/16 in. diam. dust core.
	9·0 Mc/s	16T 28 s.w.g. enam. Shunt Cap = 75pF	
	12·5 Mc/s	14T 24 s.w.g. enam. Shunt Cap = 50pF	
	19·5 Mc/s	6T 22 s.w.g. spaced to ½ in. long. Shunt Cap = 40pF	
	26·5 Mc/s	6T 22 s.w.g. spaced to ½ in. long. Shunt Cap = 25pF	
	33·5 Mc/s	5T 22 s.w.g. spaced to ½ in. long. Shunt Cap = 15pF	
	34·0 Mc/s	5T 22 s.w.g. spaced to ½ in. long. Shunt Cap = 10pF	
I.F. Trap	5·3 Mc/s	15T 24 s.w.g. enam. Shunt Cap = 350pF	Former: Aladdin as for conv. osc.
Q Multiplier	460 kc/s	60T approximately 9/42 Litz wire on *Maxi-Q* pot core ⅞ in. diam. with adjustable slug. (Adjust number of turns as necessary to obtain correct tuning.)	

must not be replaced with a EF91, 6AM6, EF93, 6BA6 or similar, because these valves will not give the required r.f. output voltage.

Any suitable moving coil meter of between 500μA and 1·5 mA full scale deflection is suitable for the S meter. The original scale can be covered with glazed drawing paper, hand calibrated and finally removed and marked in with Indian ink.

There is nothing worse than attempting to tune a selective receiver with a " lumpy " drive mechanism or with one that has backlash in the gearing. A really first class reduction drive—preferably with a reduction of at least 100 : 1—is essential. A readily available type is the Eddystone type 898 drive and dial assembly.

All the coils, i.f. transformers and switches for this receiver are available from Electroniques (Standard Telephones and Cables Ltd., proprietors, Harlow, Essex).

Alignment

An ambitious project such as the construction of a double superhet communication receiver will only be undertaken by an experienced amateur with past constructional knowledge. Detailed alignment instructions are not therefore given but the procedure, as in any receiver, is to start from the back and finish at the front. In this case, that means alignment of the carrier insertion oscillator, followed by the tunable i.f. and v.f.o. section, and finally the signal frequency circuits. The 100 kc/s pips from the calibration oscillator in conjunction with the S meter will be found very useful during this process.

The level of the conversion oscillator and v.f.o. inputs to the two mixers V2 and V4 is important. There should be sufficient input voltage to obtain correct mixing action but not so much that spurious product generation is increased. A simple way of determining the approximate peak drive voltage to the first mixer without needing a diode probe valve voltmeter is to break the 20 K ohm grid resistor connection at the earth end and use a 500μA meter to measure the grid current. This will vary over the six bands because of the different mode of operating the conversion crystals but should be between 100 and 300μA. The cathode drive to the second mixer will have to be measured with a diode probe valve voltmeter and should be approximately 2 volts r.m.s.

The output of the carrier insertion oscillator V13, also measured with a valve voltmeter, should be between 10 and 20 volts r.m.s. at the anode while the r.f. to the two OA79 diodes (either end of the balancing potentiometer) should be about 2·0 volts r.m.s.

Audio negative feedback is provided both as an aid to reduce distortion and also to give control of the total audio amplification. The 330 K ohm feedback resistor is adjusted in value so that at full audio gain setting with the a.g.c. operating, the drive to the output valve V16 is just less than the point of positive grid excursion.

With the suggested tunable i.f. range (5·0 to 5·5 Mc/s) the receiver can be tuned to MSF on 5·0 Mc/s by setting the bandswitch to the " 3·5–4·0 Mc/s " position and the preselection tuning capacitor to the minimum capacity position. This will peak the signal frequency circuits to 5·0 Mc/s and the first mixer will behave as an amplifier and feed the MSF signal into the tunable i.f. stages. This will provide a standard frequency (correct to five parts in 10^9) for receiver drift measurements and also for adjustment of the 100 kc/s calibration oscillator by means of the pre-set capacitor across the grid circuit of V17, so that the calibration pip on 5·0 Mc/s is zero beat with MSF.

The 56 K ohm resistor feeding h.t. to the noise limiter control VR1 is a nominal value only. It should be adjusted to obtain correct limiter action.

The signal frequency discrimination against i.f. breakthrough is at its worst point when the receiver is in use on those frequencies nearest to the tunable i.f. range. This is at the h.f. end of the 80m band. Accordingly, a 5·3 Mc/s i.f. trap is connected in series with the input to the primary of the 80m coil and this gives an additional 20–40db attenuation over the range 3·6 to 3·8 Mc/s. To tune the 5·3 Mc/s i.f. trap, set the main tuning control to 3·7 Mc/s (bandswitch in the 3·5–4·0 Mc/s position) and feed a 100 millivolt signal into the aerial terminal from a signal generator set to 5·3 Mc/s. Adjust the trap dust core for minimum S meter reading.

Muting

Modern practice is to control a transmitter and receiver by means of a relay with either voice (vox) or press-to-talk operation. Breaking h.t. or cathode connections usually causes clicks and thumps and the most satisfactory control method is a source of negative voltage applied to the grids of the controlled valves. In many cases there is already a source of negative voltage provided for the p.a. bias supply and the available potential is usually suitable for muting requirements.

The controlling voltage can be any value from about 30 to 100 volts and is connected to the receiver " muting " terminal. During transmission periods the applied voltage disables the second mixer and the second audio amplifier and so " kills " the receiver. The time constant is fast and the control is free from clicks and thumps. As the controlled r.f. valve (V4) is not connected to the a.g.c. line the action and the time constant of the receiver a.g.c. system remains unaffected. Obviously there must always be a return path for the controlled grid circuits and if muting is not required the " muting " terminal must be strapped to earth.

The alternative method is to control the receiver from the transmitter send-receive switch. If this is required the receiver can be linked back to the transmitter control switch circuits by means of the STAND BY (SB) position of the receiver operational switch S16, 17, 18, 19 and the stand-by terminal provided.

Performance

Automatic Gain Control

Audio rise is 6db for 80db change in signal input above 1μV, and 3db for 60db change in signal input above 10μV.

TABLE 4.9

SIGNAL-TO-NOISE RATIO

Band Mc/s	Aerial input in microvolts for 10db signal-to-noise ratio A3a Reception	Aerial input in microvolts for 20db signal-to-noise ratio A3a Reception
1·5-2·0	0·3	1·2
3·5-4·0	0·28	1·0
7·0-7·5	0·38	1·4
14·0-14·5	0·32	1·2
21·0-21·5	0·32	1·0
28·0-29·0	0·32	1·0

I.F. Breakthrough Rejection.

This is not less than 60db.

Second Channel Rejection

The image rejection to both mixing processes is not less than 60db on all bands.

Selectivity

The selectivity is 2·5 kc/s wide 6db down and 3·7 kc/s wide 60db down. The shape factor of the filter is 1·48.

Spurious Responses

On all bands self-generated spurious responses are below a level (throughout the 500 kc/s tuning range) equivalent to a 0·25μV aerial input signal.

Stability

The initial drift from switching on is approximately 500 c/s; the v.f.o. is stable within 10 to 15 minutes (depending on ambient temperature) from cold. Thereafter drift is less than 100 c/s over any one hour period. The measurements were made receiving MSF on 5·0 Mc/s with the open chassis on a bench under normal room temperature conditions. Self-generated heat from the chassis in a poorly ventilated cabinet would adversely affect these figures and require additional negative temperature coefficient compensation. The necessary value would have to be found experimentally.

Circuit Improvements

More modern valves such as the 6BZ6 and 6CD6 which have more linear characteristics are capable of better performance in an r.f. stage than the cascode circuit. Double triode mixers such as the ECC85 have lower noise and much better cross modulation and overload characteristics than the 6BE6. A logical improvement therefore is to substitute these in the G2DAF receiver. It is also assumed that, at the present very competitive prices, most constructors will be prepared to purchase new current production crystals correctly ground for optimum operation on the required frequency. Accordingly a second ECC85 valve is used as a fundamental or overtone oscillator. This ensures that there is less risk of harmonics from the second converter v.f.o. beating with the first conversion oscillator output to produce birdies. A circuit diagram of the improved front-end is shown in **Fig. 4.95(a)**. It may be used as a converter with a receiver tuning over the range 5·0 to 5·5 Mc/s using the 75 ohm output circuit shown. The second mixer may be changed to a double triode ECC85 with advantage and the easily obtainable mechanical filters offer improved selectivity.

Fig. 4.95b shows the necessary changes for the double triode second mixer V4 and the Kokusai mechanical filter. Both the tunable i.f. coils L1 and L2 may be wound on Neosid or Aladdin 0·3 diam. formers with 2¼ in. × $\frac{13}{16}$ in. × $\frac{13}{16}$ in. screening cans. Each will require a winding of 32 turns of 32 s.w.g. enamelled wire. The shunt capacitance across the coil is made up with a fixed mica capacitor and a Philips 3-30 pF trimmer, and this value shunt controls the frequency variation of the circuit caused by the rotation of the variable tuning capacitor. By adjustment to the dust core and the

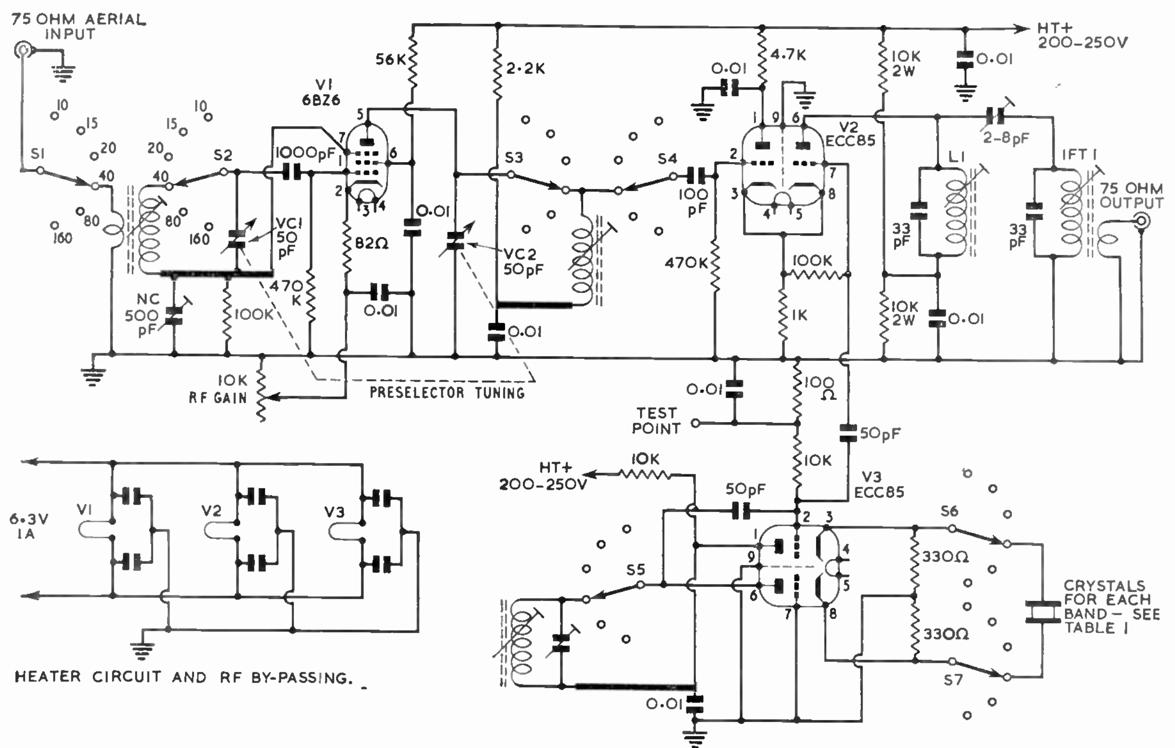


Fig. 4.95a. Circuit diagram of the improved front-end for the G2DAF receiver. For clarity only the 40m band coils are shown and the common connecting points for the other ranges are indicated by heavy lines.

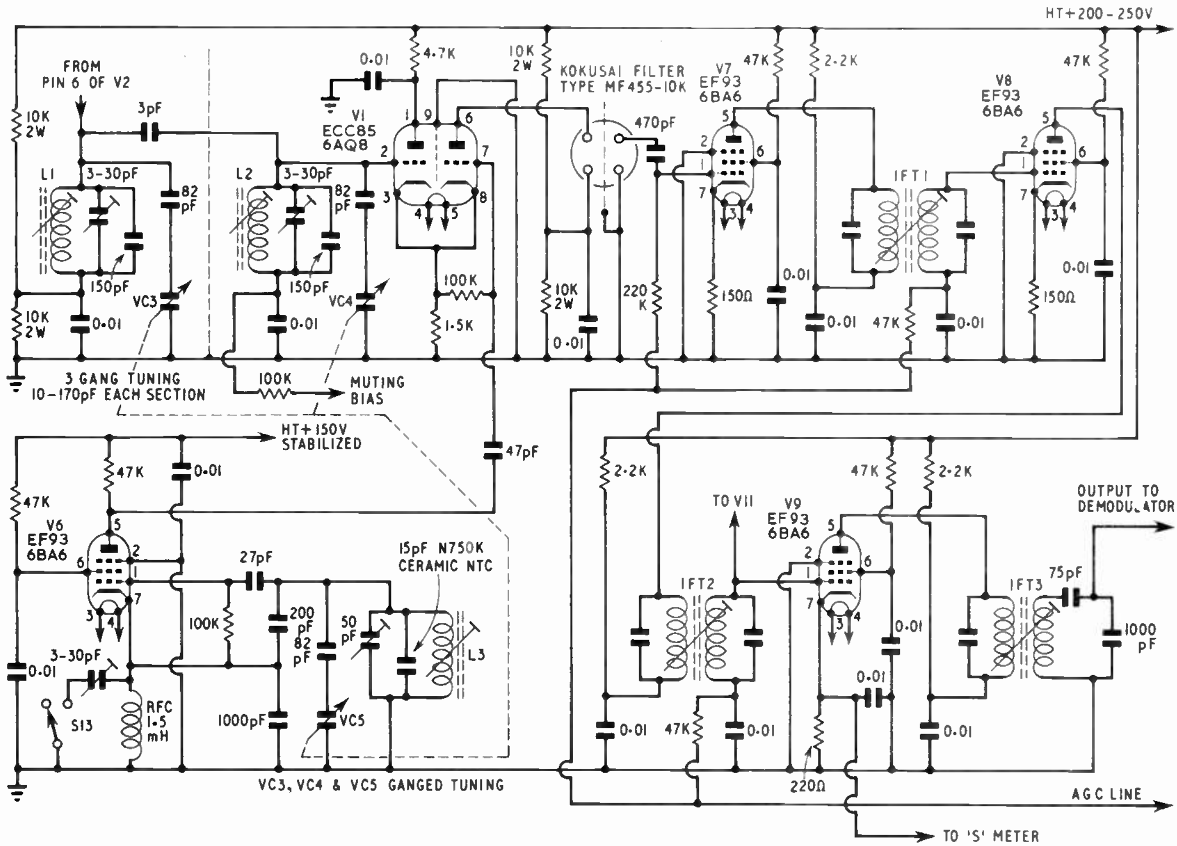


Fig. 4.95b. Circuit diagram of the modified tunable I.F. section for the G2DAF receiver.

Philips trimmer, each coil can be made to track correctly so that it covers the range 5·0 to 5·5 Mc/s, with the dust core at the l.f. end, and the trimmer at h.f. end. L1 and L2 may be the original coils supplied by Electroniques and as these are unscreened they must be mounted in the position indicated, but under the chassis. If this procedure is adopted it is most important that one coil cannot " see " the other coil and a small cross screen must be fitted between L1 and L2 so that the *only* coupling for r.f. is via the 3 pF top capacitor.

Major changes necessary in the i.f. strip are removal of the original filter crystals and associated coupling transformers: this enables considerable simplification and a worthwhile saving in chassis space. It also allows the three valves V7, V8 and V9, and the i.f. transformers to be positioned in a straight line along the right-hand side of the central chassis section.

As the load presented by the mixer valve, V4, to the v.f.o. is at high impedance and the grid input capacitance is very small due to the negative feedback across the 1·5 K ohm cathode resistor, the original v.f.o. cathode follower valve V5 can be omitted. A mixer injection of between 1 and 2 volts r.m.s. is ample for the ECC85 valve, and this low level output from the v.f.o. also reduces the amplitude of birdies that might be produced by higher order harmonics of the v.f.o. beating with the conversion crystal in use. It may be found that the mixer operates more quietly if the 47 pF capacitor to the oscillator anode is transferred to the cathode.

In regard to v.f.o. stability it should be clearly understood that, there is no such thing as a " driftless " L/C oscillator. A quartz crystal has a high degree of frequency stability because quartz is a material with a low temperature co-efficient. Replacing the crystal by building an equivalent series tuned circuit using L and C—as for instance in the so-called " Clapp " v.f.o.—does not give the same stability as a crystal, for it is not possible to manufacture coils and variable capacitors with the temperature co-efficient of natural quartz.

The v.f.o. coil former should be ceramic, with a dust core mounted on a screwed brass rod fitting into a pressure loaded clutch so that there is neither end or side float of the core within the winding. Highly recommended is the ½ in. diam. Cambion former Type 1538-2-2 with a 20063-K slug. The winding of 14·5 turns of 22 s.w.g. enamelled wire should be close wound under tension, and finally thoroughly doped with polystyrene cement.

The Kokusai Type MF 455-10K mechanical filter has a nominal bandwidth of 2 kc/s at the 6dB points. In practice the passbands are usually wider than this and KW Electronics Ltd. will be able to supply an MF 455-10K filter with a 6dB bandwidth of 2·5 kc/s if so requested. Filter bandwidth in a receiver is inherently a compromise between the conflicting requirements of the maximum possible selectivity and a reasonable audio bandwidth without objectional coloration.

CHAPTER 5

V.H.F./U.H.F. RECEIVERS

In a v.h.f. receiver it is possible to realize a performance superior in terms of sensitivity and signal-to-noise ratio to that normally obtained on the lower frequencies where, owing to various forms of interference both man-made and natural, a limit is imposed beyond which any attempt to recover signals is fruitless. In v.h.f. reception there is no atmospheric noise except that originating in lightning discharges in the immediate vicinity of the receiver, while the bands are sufficiently wide compared with their occupation that interference between stations is, as yet, not generally a serious problem.

The limiting factor is extra-terrestrial noise and at frequencies up to at least 250 Mc/s, receivers can be designed which will respond to signals only slightly above this level. Interference from the ignition systems of cars is more noticeable on the higher frequencies and in fact becomes objectionable from about 20 Mc/s upwards. Such forms of impulsive interference can be greatly reduced, however, by the use of a noise limiter.

Noise Factor

For these reasons the performance of v.h.f. receivers is usually (and more usefully) specified in terms of *noise factor* which may be defined as:

$$\frac{\textit{aerial noise} + \textit{receiver noise}}{\textit{aerial noise}}$$

and measured as the noise power present at the output of the receiver. With the theoretically perfect receiver, producing no noise in itself, this equation becomes 1/1, i.e. a noise factor of 1 or 0db. The noise factor of a practical receiver that does itself generate unwanted noise is a measure of the amount by which it falls short of perfection.

It is customary to regard each valve stage in the receiver as a noiseless amplifier and to consider the noise that it does in fact generate as originating in a fictitious resistor in its grid circuit. The value of this resistor is such that the noise voltage which would be developed across it is multiplied by the gain of the valve, the result being the real noise present in the valve output circuit. The subject of receiver noise is discussed more fully in Chapter 15 (*Noise*) and the measurement of noise in Chapter 19 (*Measurements*) to which the reader is referred for a more extensive explanation.

Front-end Stages

The requirements of the early stages in a v.h.f. receiver are (a) low noise content, (b) high power gain, and (c) ability to develop a signal voltage between grid and cathode.

The first of these requirements implies that noise due to the fictitious resistor in the grid circuit, known as the *equivalent noise resistance*, and that due to *induced grid noise* must be as low as possible: for definition of these terms see Chapter 2 (*Valves*). Since a pentode has an equivalent noise resistance which may be several times higher than that of a triode owing to the added effect of *partition noise*, triodes are used exclusively for low noise input stages. The second requirement implies that the stage must be capable of developing gain at the desired frequency with the load presented by the input of the following stage. This demands a valve with as high a slope as possible consistent with low capacitances and high input impedance.

In a multi-stage receiver with, say, eight amplifying stages the noise originating in the grid circuit of the first valve will be amplified by eight valves, that in the second by seven, and so on. If the effective noise voltages are denoted by V_1, V_2, $V_3 \ldots V_8$ and the stage gains by $G_1, G_2, G_3 \ldots G_8$, the total noise present at the output of the receiver will be $V_1 (G_1, G_2, G_3 \ldots G_8) + V_2 (G_2, G_3 \ldots G_8) + V_3 (G_3 \ldots G_8)$ and so on. It follows, therefore, that if G_1 is high, say 20 or more, only the noise due to the first stage is of importance, and provided that the remaining stages are reasonably efficient they will contribute little to the overall noise. In cases where the gain of the first stage is low because of an unsuitable type of valve for the frequency concerned or if a considerable bandwidth is required, or when both of these conditions exist, the noise contributed by the second or even the third stage may become important.

If all the noise is not due to the first stage, the noise factor is increased by the noise factor of the second stage. The formula for the overall noise factor $F_{1,2}$ of two stages of independent noise factors F_1 and F_2 is:

$$F_{1,2} = F_1 + \frac{F_2 - 1}{G_1}$$

where G_1 is the available power gain of the first stage. As an example, a mixer has a noise factor of 25 (14db). A triode amplifier preceding such a mixer with a noise factor of 5 (7db) and a power gain of 10 (10db) will cause the receiver to have an overall noise factor of:

$$F_{1,2} = 5 + \frac{25 - 1}{10} = 7{\cdot}4 \ (8{\cdot}7\text{db}).$$

From this example it is clear that the mixer noise must be low or the gain of the r.f. stage must be high unless a valve with extremely low noise is used in order to achieve a good overall result.

The third requirement is that the input resistance which appears in the grid circuit due to the transit time of the electrons must be high. Although the noise associated with this resistance (i.e., the induced grid noise) is high because its temperature can be taken as five times the room temperature, the signal voltage applied to the grid will be low unless a step-up in voltage between the aerial (or its feeder) and the grid can be obtained.

When the input circuit is matched to the input impedance of the valve, an optimum power gain is achieved and the valve is said to be *power-matched.* This condition does not, however, result in the lowest noise. Where the step-up ratio is adjusted for optimum noise factor in contrast to optimum power gain the valve is said to be *noise-matched.*

Fig. 5.1 shows the variation with frequency of the noise factor and optimum source resistance of a grounded grid stacked-ceramic triode of similar construction to that described on page 2.22, Chapter 2 (*Valves*). It will be seen that when the valve is power-matched the noise factor is constant at 5·5db up to about 400 Mc/s, whereas below this frequency a higher source resistance (over-coupling) produces considerable improvement in the noise factor. Above this frequency the noise factor does not diminish because the optimum source resistance is then comparable with or may be less than the cathode input resistance ($1/gm$ for this valve is $1000/9 = 110$ ohms).

The noise factor of a valve is affected by the cathode temperature, owing to the existence of the space charge in the grid-cathode region. The extent of the space charge is critically dependent on the temperature of the cathode and consequently there will be considerable increase in noise if the heater current is reduced by even a very small amount, e.g. only 5 per cent. It is most important, therefore, to ensure that the correct heater voltage is applied to valves, particularly in the early stages of the receiver. Adjustment of the value of the grid bias in relation to the contact potential of the grid is a means of improving the noise factor. If the grid bias of a valve is steadily reduced below the cut-off point the noise steadily diminishes until the point is reached where grid current begins to flow; beyond this point the noise increases rapidly. **Fig. 5.2** shows the relation between noise factor and grid bias at a constant anode current for two typical grounded grid triodes. The general shape of the curves is the same but the minimum noise factors occur at slightly different grid bias values. The optimum bias will vary with the type of valve and different specimens of the same type will also show some variation. It is often worth experimenting to obtain the optimum bias but care should always be taken to keep the anode current constant at the recommended value by means of an adjustable series h.t. dropping resistor.

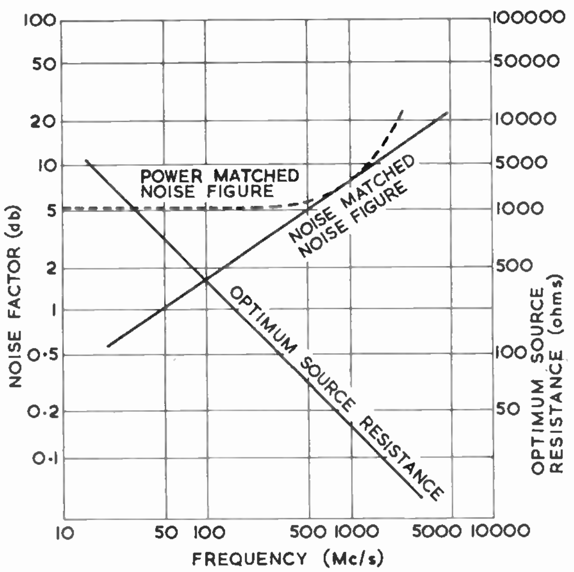


Fig. 5.1. Noise factor and optimum source resistance plotted against frequency for a typical grounded grid triode (type 7077, g_m 9 mA/V, I_a 6·4 mA).

Circuit Noise

Noise due to circuit components other than valves is produced solely by the resistive component; inductive or capacitive reactances do not produce noise. Coils have negligible resistance in the v.h.f. ranges but the leakage resistance of capacitors and insulators is important and it is therefore imperative to select good quality components, including such items as valveholders and switches. Circuit noise can be regarded as including noise caused by regeneration and although it is common practice to enhance the gain of low frequency circuits by applying regeneration this should be avoided at all costs in v.h.f. receivers because of the additional noise produced. Common causes of regeneration are:

(a) Insufficient decoupling of supply leads and, particularly, heater and cathode circuits. In the range 20–30 Mc/s 0·01 μF non-inductive capacitors are a *minimum* value for this purpose.

(b) Incomplete neutralizing of triode r.f. amplifiers. An adjustment that merely ensures that the circuit does not actually oscillate is not sufficient. Accurate neutralizing is essential and the appropriate methods are described later in this chapter.

(c) Circulating r.f. currents in the chassis, due to multipoint earthing.

(d) Poor earth contacts to the chassis. For low noise circuits a brass or copper chassis should be employed; aluminium is not recommended because it is extremely difficult to make a lasting effective connection to it.

Effect of Bandwidth

Before leaving the subject of noise, there is one further point which must be considered. If the noise factor of a receiver is measured with a noise generator it will be found to be independent of receiver bandwidth. This is because the noise from the generator is of the same nature as ordinary noise so that if the bandwidth is doubled the overall noise is also doubled. For the reception of a signal of finite bandwidth, however, the optimum signal-to-noise ratio is obtained when the bandwidth of the receiver is only just sufficient to accommodate the signal and any further increase in bandwidth results merely in additional noise. The signal-to-noise ratio at the receiver therefore depends on the power per unit bandwidth of the transmitted signal. To illustrate this point, suppose that the receiver produces 1μV of noise for each 10 kc/s of bandwidth and that an amplitude modulated transmitter radiates a signal 10 kc/s wide and produces 10μV at the receiver. The signal-to-noise ratio is therefore 10, provided that the receiver uses a bandwidth of 10 kc/s. If the bandwidth of the transmission is reduced to 5 kc/s and the radiated power remains the same, the input to the receiver will still be 10μV, but if the bandwidth of the receiver is also reduced to 5 kc/s, only 0·5 μV of noise will be accepted and

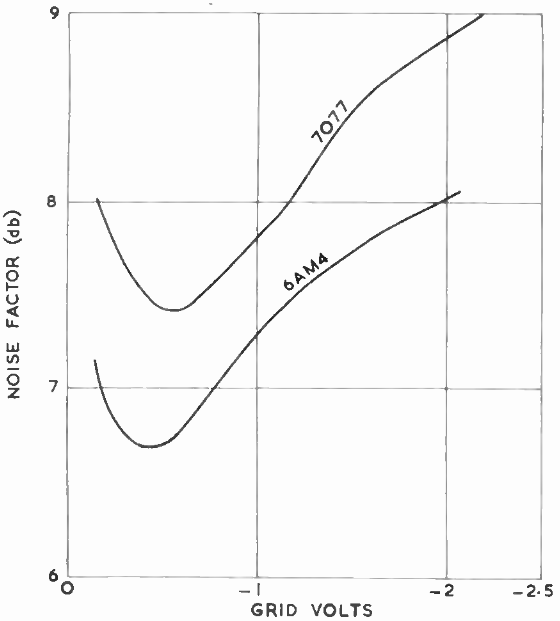


Fig. 5.2. Noise factor plotted against grid bias for grounded grid triodes, the anode current being held constant (7077 at I_a 6·4 mA at 450 Mc/s; 6AM4 at I_a 10 mA at 250 Mc/s).

the signal-to-noise ratio will increase to 20. An unstable transmission, requiring a large bandwidth at the receiver, results in a degraded signal-to-noise ratio, so there is clearly a great advantage to be gained by the employment of stable transmitters and receiving equipment having the smallest practicable bandwidth.

CHOICE OF THE FIRST I.F.

Double frequency conversion is usually employed in a receiver for v.h.f. reception and it is convenient to build the r.f. stages and first conversion circuits as a separate unit, known as a *converter*, the output of which is fed into a communications receiver with a suitable frequency coverage for the first i.f.

Two methods are possible for tuning over a band when double conversion is employed:

(*a*) The oscillator in the converter may be fixed in frequency and may, therefore, be crystal controlled, tuning being effected by variation of the first i.f. (i.e., by tuning the main receiver) over a band equal in width to that of the v.h.f. band to be covered.

(*b*) The oscillator in the converter may be variable and the main receiver tuning set at the frequency chosen for the first i.f. It has already been shown that the signal-to-noise ratio of a receiver depends upon the bandwidth, and this bandwidth being that of the most selective stage (normally the i.f. amplifier in the main receiver) and the amount of noise heard will thus be determined provided no significant noise is contributed by the later stages. The use of a bandwidth exceeding 5 kc/s for telephony and considerably less for c.w. reception is therefore undesirable when a good signal-to-noise ratio is the prime consideration.

In any superheterodyne receiver it is possible for two incoming frequencies to mix with the local oscillator to give the i.f.—the desired signal and the image frequency—and this has an influence on the choice of the i.f. To avoid confusion, the i.f. of the main receiver will in future be referred to in this chapter as the *second i.f.*

One purpose of r.f. amplification ahead of the mixer in a v.h.f. converter is to attenuate signals and noise present at the image frequency and the higher the first i.f. the greater such attenuation will be. Many of the r.f. amplifier circuits employed in v.h.f. receivers have an extremely large bandwidth and if they are so wide that the image ratio is unity (i.e., if the sensitivity of the " second channel " is the same as that for the wanted signal) the noise factor is degraded by 3db. Care should therefore be taken to ensure that the first i.f. is sufficiently high. As the first i.f. is raised, the performance of the main receiver tends to fall off and its contribution of noise may become significant if the overall gain of the converter is insufficient. The value of the first i.f. finally decided upon depends on the type of converter to be used and the frequency on which it is to operate.

Two other factors affect the choice of first i.f. First, it is desirable that no harmonic of the local oscillator in the main receiver should fall in the v.h.f. band in use, and second, there should be no breakthrough from stations operating at the frequency selected as the first i.f.

Many communications receivers produce quite strong harmonics in the 70 and 144 Mc/s bands and although these are high-order harmonics and are therefore tuned through quickly they are troublesome inasmuch as they are distracting when searching for signals in the band in question. The problem is more acute when the converter is crystal controlled than when a tunable first oscillator is employed. In the former case freedom from harmonics is required over a tuning range equal to that of the v.h.f. band to be covered whereas in the latter only harmonics from one spot frequency of the local oscillator in the main receiver have to be considered.

The avoidance of breakthrough signals at the frequency selected for the first i.f. is also more difficult when the converter is crystal controlled. Because it is practically impossible to find a band several megacycles wide which is completely unoccupied, other steps have to be taken to avoid breakthrough. Frequencies in the range 20–30 Mc/s are often chosen since fewer strong signals are normally found there than on the lower frequencies, but this condition may well be reversed during periods of high sunspot activity.

Particular attention should be paid to the screening of the main receiver which should be tested by itself with the i.f. stages operating at maximum gain. Signals may be picked up on the aerial terminals and if this occurs they should be replaced by a coaxial socket. Unwanted signals may also enter the receiver along the power supply cable or the loudspeaker leads and these may need to be decoupled or filtered. Other sources of such interference are long earth leads and unearthed control spindles.

When the main receiver is considered satisfactory from this point of view, a further test should be made with the converter connected, preferably by a short length of coaxial cable, but with its local oscillator inoperative. If any signals are heard they will be first i.f. signals and the decoupling or filtering of the power supply leads and any similar connections should be checked again. If difficulty is still experienced, a stage of i.f. amplification may have to be incorporated in the converter or, if one is already in use, its gain may need to be increased so that the main receiver can be run in a less sensitive condition.

After all abnormal sources of signal pick-up have been eliminated there may still exist a path for i.f. signals through stray capacities in the converter itself. This may be eliminated to a very large extent by connecting a piece of coaxial cable as a quarter wavelength (at signal frequency) shorted stub in parallel with the aerial feeder at the input of the converter. This provides a virtual short circuit to signals at the first i.f. while maintaining a high impedance to the wanted v.h.f. signals. The actual physical length of the stub will depend of course upon the velocity factor of the cable used in its construction.

THE LOCAL OSCILLATOR

It is clear that if the final bandwidth of the complete receiver is to be the same as that used on the lower frequencies the stability of the oscillator in the converter must be as good as the oscillator in the main receiver. For the 70 and 144 Mc/s bands it is possible to build tunable oscillators of good stability and " note " working at frequencies either above or below the signal frequency by the value of the first i.f. On the 420 Mc/s band tunable oscillators are normally designed to operate at one-half or one-third of the required frequency and are followed by suitable frequency multipliers. Crystal controlled local oscillators can be used satisfactorily in any v.h.f. or u.h.f. converter.

The advantages and disadvantages of the two types of oscillator may be summarized as follows.

Tunable Oscillators

(a) *Advantages*

(i) A directly calibrated dial is possible.
(ii) The cost of a crystal is eliminated.
(iii) Only one valve is required.
(iv) There is less likelihood of harmonic interference from the main receiver.
(v) Only one clear channel is required on the main receiver.

(b) *Disadvantages*

(i) Long-term oscillator drift makes calibration of the dial unreliable.
(ii) Warm-up drift can be troublesome.
(iii) It is difficult to obtain a better than T8 note (but not impossible).
(iv) Change of oscillator or mixer valve alters the calibration.
(v) Microphony occurs on strong signals due to vibration of the oscillator tuned circuit by acoustic feedback from the lousdpeaker.
(vi) They cannot readily be used remotely.

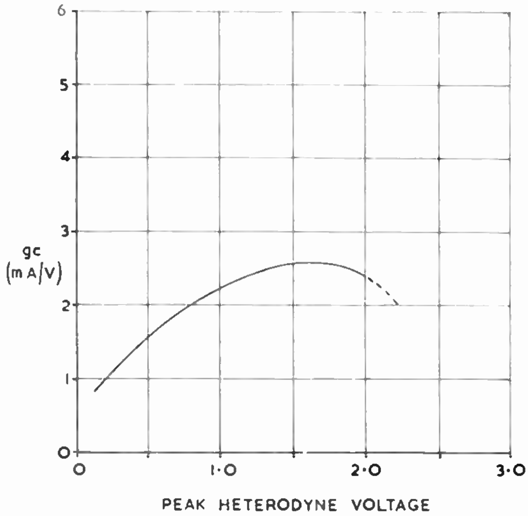


Fig. 5.3. The relation between conversion conductance and oscillator (heterodyne) voltage in a typical triode mixer (6AM4, V_a 125 volts, R_k 220 ohms).

Crystal-controlled Oscillators

(a) *Advantages*

(i) Accurate vernier logging of stations on the main receiver dial is possible.
(ii) A T9 note is assured.
(iii) Negligible short term or warm-up drift.
(iv) Absence of controls allows remote operation.

(b) *Disadvantages*

(i) Expensive both in valves (for 420 Mc/s and above) and crystal.
(ii) Possibility of additional self-generated whistles.
(iii) More adjustments and therefore more difficult to align in the first place.

The question of oscillator injection arises irrespective of the type of oscillator it is decided to use. As already stated, if the r.f. stages have high gain and low noise factor, the noise contributed by later stages tends to be less significant, but this assumes reasonable efficiency in the mixer stage. **Fig. 5.3** shows a typical conversion conductance curve for a mixer stage and indicates how the i.f. output varies as the injection voltage is changed, the signal voltage being kept constant: it will be noticed that a considerable loss of gain results if inadequate oscillator voltage is used. The noise level is at its highest with inadequate injection voltage, and although the noise increases when the injection voltage is too high it is still far less than when it is too low. In the designs presented in this chapter every effort has been made to ensure that adequate injection voltages are attained.

The injection voltage in a triode mixer—the type used by most amateurs at v.h.f.—is found by measuring the current flowing through the mixer grid resistor due to the oscillator injection, i.e., the difference in grid current with and without the oscillator operative. The maximum value of oscillator voltage normally to be expected is about 3 volts peak, and with a grid resistor of about 50 K ohms, this corresponds to a grid current of only 60 μA, so any lower injection voltage may become difficult to detect. It must not be supposed that a voltage of this magnitude is easy to achieve; to produce such a voltage at the mixer grid is in fact often quite difficult. **Fig. 5.4** shows a typical triode mixer circuit in which the coil L1 is tuned to resonance at, say, 70·3 Mc/s and the signal voltage developed across it fed through a 30 pF capacitor to the mixer grid. Suppose L2 is carrying oscillations at 50 Mc/s, coupled through a 10 pF capacitor to the grid of the mixer (V2). At 50 Mc/s, the inductance L1 is in effect a short-circuit to earth, and at this frequency the 50 Mc/s oscillations therefore " see " the mixer grid circuit as shown in **Fig. 5.5.**

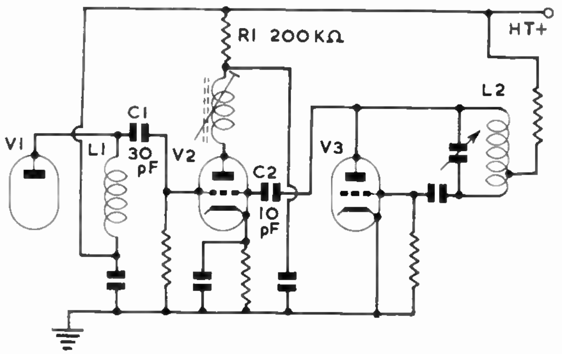


Fig. 5.4. Typical triode mixer (V2) and associated oscillator (V3).

The 10 pF and 30 pF capacitors (C2 and C1) in the mixer grid circuit become a potential divider such that only one quarter of the original 50 Mc/s voltage at the anode of V3 appears at the grid of V2.

An obvious course would be to increase the value of C2, but signal voltages at the grid of the mixer "see" this capacitor between grid and earth since L2, appears as a short circuit at signal frequency. Thus only three-quarters of the signal voltage at the anode of V1 appears at the grid of V2 with the existing capacitor ratio; if C2 was increased in value even less voltage would be available at the mixer grid.

Experimental adjustment of the values of C1 and C2 may be made (with appropriate readjustment of L1 and L2) to effect the best compromise. Capacities of 30 pF and 10 pF respectively are of the correct order of magnitude because an r.f. voltage of 10 to 20 is commonly obtained at the anode of V3 and one quarter of this voltage at the mixer grid is usually adequate.

The required value of oscillator injection voltage depends upon the anode potential of the mixer valve; the lower the anode voltage the lower the required injection voltage and it is this fact that has made the use of relatively high values for R1 (Fig. 5.4) commonplace.

If R1 is replaced temporarily by a variable resistor of about 250 K ohms which can be adjusted while listening to a weak signal, it will be found that the optimum value of mixer anode potential is readily obtained. The value of resistance thus determined may weaken strong signals slightly but this is of little consequence in relation to the marked improvement effected in the response to weak signals.

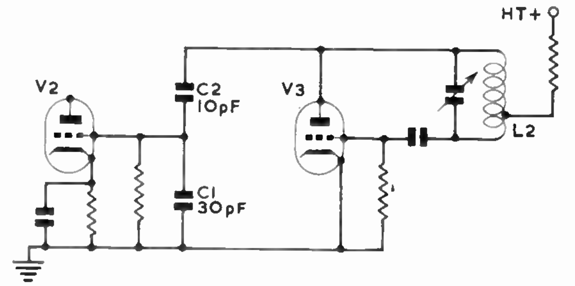


Fig. 5.5. The triode mixer circuit of Fig. 5.4 arranged to show the potential divider formed by the oscillator coupling and r.f. to mixer stage coupling capacitors at the injection frequency.

TUNABLE OSCILLATORS

The requirements for a tunable oscillator are similar to those for a variable frequency oscillator designed for transmitter control, i.e. robust construction to guard against vibration, the employment of only the best quality components, including valve holders and insulation, short and rigid wiring and self-supported coils of stout wire. The operating frequency of the oscillator may be either above or below the signal frequency but to ensure the best stability it is usual for it to be on the *low* frequency side of the signal. Where the choice of such a frequency leads to interference with a local television channel, operation on the *high* side of the signal may be necessary although a small change in intermediate frequency sometimes enables the oscillator frequency to be placed to avoid such trouble.

The valve employed will, of course, have considerable influence on the results obtained, and the type chosen should be of robust construction with short leads to the electrodes.

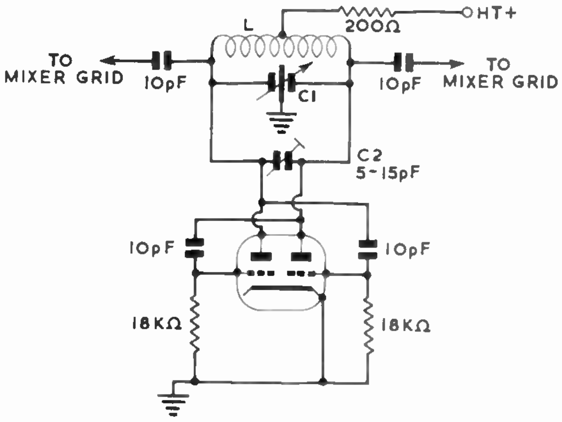


Fig. 5.6. The Kalitron oscillator. L is 2 turns of 16 s.w.g. $\frac{3}{16}$ in. diameter for 120 Mc/s. C1 may be 7·5 pF per section.

Modern all-glass types on either B7G or B9A bases are suitable and some advantage may be obtained by using the "special quality" or "trustworthy" (ruggedized) variety.

There are three main types of circuit for tunable v.h.f. local oscillators: (*a*) Kalitron push-pull, (*b*) v.h.f. Colpitts, (*c*) Hartley.

The Kalitron is particularly applicable to the requirements of the push-pull mixer, while the Colpitts and the Hartley circuits are primarily intended for use with single-ended mixer circuits. Each of the circuits is capable of producing good results if well constructed.

The Kalitron Oscillator

This circuit, shown in Fig. 5.6, is well suited to the 6J6 valve in which the two triodes have a common cathode connection and the cathode lead inductance is therefore cancelled out, but other types such as the 12AT7 or 12AU7, which are double triodes with separate cathodes, can be used successfully.

For either 70 or 144 Mc/s converters the oscillator tuning capacitor C1 may be a small split-stator type with one fixed and one moving plate per section. The wire used for the coil should not be thinner than 16 s.w.g. and the coils should be mounted directly on the soldering lugs fitted to the capacitor. The wires to the anode tags of a high quality (preferably p.t.f.e.) valveholder should be equally stout. Careful proportioning of the parallel capacitance C2 and the size of the coil will enable the required bandspread to be obtained.

With an h.t. supply of 100 volts the circuit produces about 18 volts peak-to-peak across the anode coil which gives an injection voltage of the correct order to the mixer with 10 pF coupling capacitors. The actual value of the injection voltage will depend upon the relative frequency to which the grid circuit of the mixer is tuned and the oscillator frequency. for optimum results it is worth trying slightly different values of coupling capacitor. It should be noted that the h.t. supply to the coil is through a low value resistor R; if a choke is used at this point the frequency of oscillation is very likely to be determined by the choke instead of by the tuning inductance.

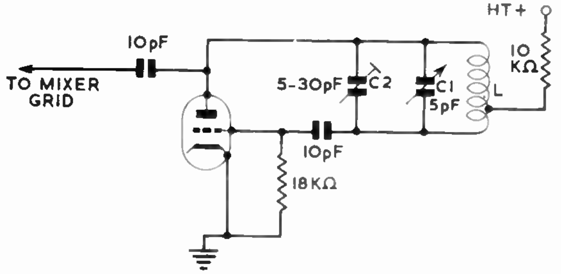


Fig. 5.7. The v.h.f. Colpitts oscillator circuit. For 120 Mc/s L should be two turns of 16 s.w.g. ½ in. diameter.

The V.H.F Colpitts Oscillator

This circuit, shown in Fig. 5.7, is a very popular one and is readily seen to be a true Colpitts oscillator by referring to Fig. 5.8 where the anode-to-cathode and grid-to-cathode capacitances are shown. At lower frequencies these capacitances are not large enough effectively to tap the tuned circuit and therefore external capacitors must be added.

The 6C4 valve is very satisfactory in the circuit as it is free from microphony and gives a fairly high output. A suitable value for the tuning capacitor C1 is 5 pF with a 5-30 pF preset capacitor C2 in parallel for bandsetting. The required bandspread may be obtained by suitably proportioning the coil and the bandsetting capacitor. As in the Kaliatron circuit, the h.t. feed to the tuning inductance is through a resistor and not through an r.f. choke in order to avoid the risk of spurious oscillations.

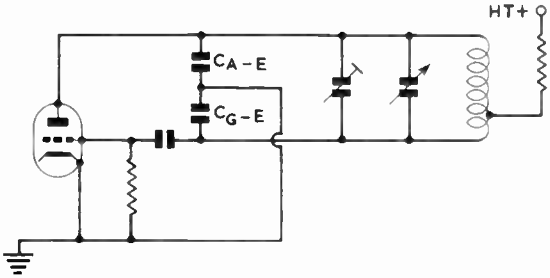


Fig. 5.8. The circuit of Fig. 5.7 with the addition of the valve capacities to show the normal Colpitts oscillator circuit.

The Hartley Oscillator

The Hartley oscillator circuit shown in Fig. 5.9 differs from the v.h.f. Colpitts oscillator arrangement (Fig. 5.7) only in the position of the h.t. feed to the tuning coil L but it generally gives rather more output.

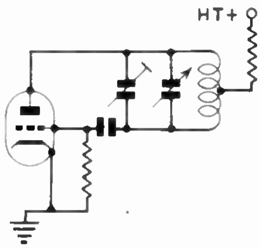


Fig. 5.9. The Hartley oscillator. Similar values to those shown in Fig. 5.7 apply.

Oscillator Construction

The stability of any of these circuits may be improved by adopting a trough-line construction for the inductance in place of a coil. The following description of a 144 Mc/s Hartley circuit illustrates the use of such an arrangement. Suitable values for the various components will be found under Fig. 5.10.

The oscillator is built in a copper trough measuring 7 in. × 1¾ in. × 1¾ in. (see Fig. 5.11). The tuned circuit consists of a hairpin-shaped loop of silver plated copper tube or wire of

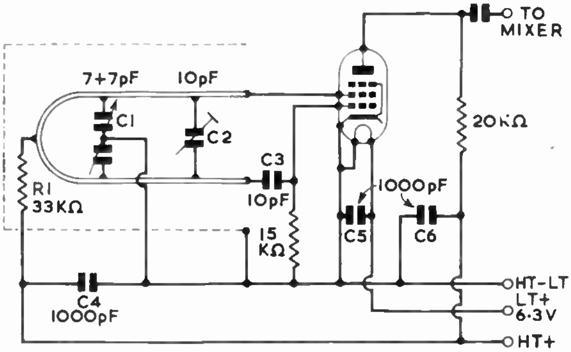


Fig. 5.10. Hartley oscillator for improved stability. C1 is a miniature split stator capacitor; C2, a miniature air spaced trimmer (Wingrove and Rogers); C3 disc ceramic or silver mica type; C4, C5, C6, hi-K ceramic feedthrough. The valve is a 6AK5. The inductance measures 4½ in. from the open ends to the point where R1 is connected. The dotted lines represent a copper box surrounding the inductance.

⅛ in. diameter and 4½ in. long held centrally in the box by two transverse strips of polystyrene ¼ in. thick secured to the sides by two short 6 B.A. bolts on each side tapped into the strips. The centre-to-centre spacing of the limbs of the hairpin is ¾ in. The valveholder is fitted in the side of the box 1¼ in. from one end so that the screen grid pin is immediately adjacent to one end of the hairpin line. The distance between the grid contact and the other side of the line is conveniently bridged by the grid capacitor (C3). All earthy connections on the valveholder and the earthed end of the grid-leak are taken to a tag under one of the valveholder fixing bolts. The most rigid construction is ensured if the heater and h.t. bypass capacitors C5 and C6 are of the feedthrough type but

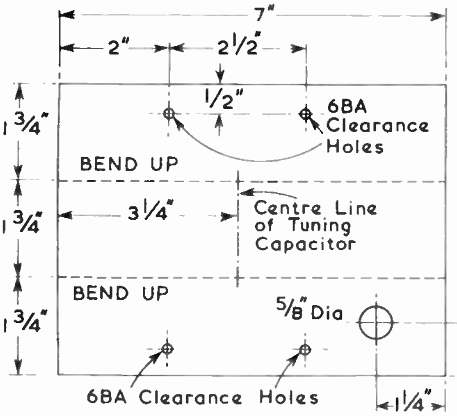


Fig. 5.11. Details of the 18 s.w.g. copper box for the oscillator of Fig. 5.10. Two pieces 1¾ in. square are soldered into place to form the ends of the box. No holes in the bottom of the box for outgoing leads are shown.

silvered mica capacitors may be used provided they are securely fixed.

The bandsetting capacitor (C2) is mounted on a small block of polystyrene ¼ in. thick bolted to the bottom of the box between the two sides of the hairpin loop to bring its contact lugs close to the ends of the line. The split-stator tuning capacitor (C1) should be of the metal frame type with one moving and two fixed plates in each of its sections. It is mounted across the side members of the box using 6 B.A. bolts, the width can be varied slightly to suit the length of the capacitor. The shaft of C1 is 3¾ in. from the valve end of the box and the connections between the stators and the line are soldered approximately 1½ in. from the open end of the line. The bandspread available with any given variable capacitor depends on this distance—if necessary the tuning range can be adjusted by altering it slightly.

Screen-grid voltage is fed to the loop end of the line, and the bypass capacitor arrangements should be similar to those for other tunable oscillators. A feedthrough insulator or small length of polystyrene rod drilled to pass the coupling lead from the anode of the valve to the mixer is set in the wall of the box close to the anode tag of the valveholder.

The power requirements of the circuit are very modest, a supply of 25–30 volts at less than 1 mA giving sufficient injection for most mixers, but if this proves insufficient the h.t. may be increased to about 200 volts without causing any serious deterioration in the frequency stability. Under these conditions, however, the anode resistor would of course need to have a higher power rating.

A well regulated h.t. supply is required for best results and a voltage-divider network connected across an 85 or 105 volt stabilized supply and passing 5–6 mA is quite satisfactory, the actual voltage on the oscillator being varied until the optimum injection to the mixer is obtained.

The frequency stability is such that a c.w. transmission may be held for considerable periods with only slight variations in note while no re-tuning is necessary during quite lengthy 'phone transmissions. A satisfactory degree of oscillator stability is achieved within a few minutes but it is advisable to arrange that the oscillator h.t. supply remains connected during periods of transmission.

CRYSTAL-CONTROLLED OSCILLATORS

In a crystal-controlled oscillator/multiplier chain for a converter it is usually desirable to multiply by not more than five or six times in any one stage. To decide what will be a suitable fundamental frequency a chart such as that shown in **Fig. 5.12** should be prepared.

As an example, it is found that if the local oscillator in the main receiver covers the range 26·5–28·5 Mc/s, no harmonics from it will fall into the 144 Mc/s band. If the main receiver has an i.f. of 500 kc/s and its oscillator operates above the signal frequency, a first i.f. of 26–28 Mc/s will be satisfactory. Since 26 Mc/s will be the tuning position for a signal on 144 Mc/s, the injection frequency for the converter will be 118 Mc/s (i.e. 144–26) and reference to the chart will show how this may be obtained.

It is normal practice to use an overtone oscillator in which the crystal vibrates mechanically at approximately three or five times its fundamental frequency depending upon the adjustment of the circuit. There is no output at the fundamental of the crystal and the harmonics present are related only to the overtone frequency. For example, a 5 Mc/s crystal operating on its third overtone would generate no energy below 15 Mc/s and behave like a crystal of that fundamental frequency.

Crystals specially manufactured for overtone operation are commercially available up to about 100 Mc/s and suitable

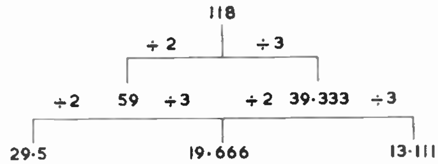


Fig. 5.12. Type of chart used for determining the fundamental crystal frequency required in an oscillator/multiplier chain.

circuits are supplied by the makers. The cost of such crystals may seem high when compared with surplus fundamental mode types but unless the frequency of operation is very high or the specified tolerance very close (which is unnecessary for most amateur purposes), careful consideration should be given to the purchase of an overtone crystal bearing in mind that such a crystal will give a greater certainty of trouble-free operation and, in the case of one of the high frequency crystals oscillating at 60 Mc/s or more, whether the greater cost is not offset by the saving in valves and components and the smaller space required for the converter.

In the example under consideration an ideal crystal frequency would be 59 Mc/s, necessitating only a doubler stage to arrive at the required injection frequency of 118 Mc/s.

For those who prefer to use the cheaper crystals other choices must be made with the aid of Fig. 5.12. It would be undesirable to start at 13·111 Mc/s since the second harmonic would fall at 26·222 Mc/s which is within the tuning range of the main receiver and would result in a spurious carrier usually referred to as a *birdie*. A frequency of 29·5 Mc/s appears to be a possibility and a 5·9 Mc/s crystal, provided it would operate on its fifth overtone, would satisfy the requirements. The oscillator unit could then comprise merely one double triode valve, the first section operating as a fifth

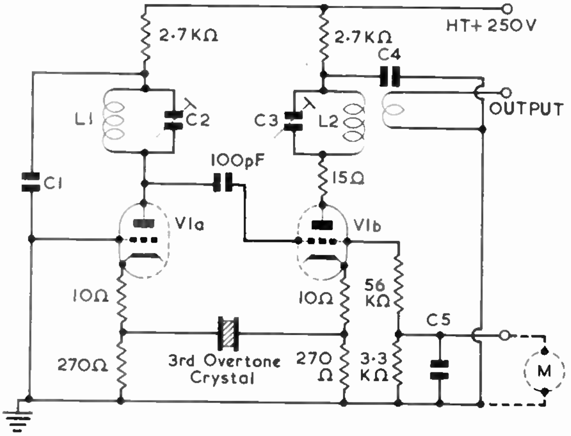


Fig. 5.13. Butler overtone crystal oscillator circuit. Output may be taken at two, three or four times the frequency of the overtone crystal. L1, C2 and L2, C3 tune to the overtone and desired output frequencies respectively. C1, C4 and C5 are bypass capacitors of normal values.

tance between the crystal electrodes and spurious oscillations may be generated unless the circuit is properly adjusted.

The Butler circuit (Fig. 5.13) is particularly useful in receivers and may be employed either as an overtone or as a fundamental oscillator with provision for taking an output at two, three or four times the frequency of the crystal, and, although two stages are required, the arrangement is very satisfactory when good stability is the main criterion. For use as a fundamental oscillator, L2 is omitted and the anode of V1b bypassed to earth by a suitable capacitor, the output then being taken from a link winding coupled to L1.

When setting up the oscillator a voltmeter may be connected as shown dotted in the diagram and the tuned circuits adjusted for maximum deflection. For highest stability the coils should both have a low Q value, and if necessary damping resistors may be connected across them for this purpose.

Typical circuits for other overtone oscillators are given in Chapter 7 (*V.H.F. Transmitters*).

overtone oscillator and the second section serving as a quadrupler.

Alternatively a start could be made at 19·65 Mc/s, obtained from a 6·55 Mc/s crystal vibrating on its third overtone, the output frequency being doubled or trebled in the second section of a double triode. The multiplier chain could then be completed in one section of a further double triode in which the remaining section served as a mixer stage.

Double triode valves are used extensively as oscillators in v.h.f. converters and among those suitable are the 6J6, 12AT7, 6BQ7A and ECC88. The triode section of the 6U8/ECF82 or the ECF804 is also suitable for use as an overtone oscillator. The 6J6 has a cathode common to both sections of the valve and this restricts its use in certain applications. The 6BQ7A and the ECC88 have an internal screen between the two anodes and are therefore particularly suitable for use when one section is to be employed as a mixer.

Several circuits have been described for overtone oscillators, some of which operate the crystal in the parallel (high impedance) mode, while in others the crystal is in the series resonant mode and therefore has a low impedance. The latter type of circuit is in more general use. A feature which is common to all the series-resonant circuits is that the crystal forms part of the feedback loop so that if it is replaced by a capacitor oscillations will take place at a frequency determined by the coil in the circuit together with its associated capacitances. This can, in fact, happen even with the crystal in circuit as a result of the capaci-

Frequency Multipliers

Unless the operating (overtone) frequency of a crystal is that required for injection into the mixer, some frequency multiplication will have to be arranged and circuits to do this will be found among the designs for converters described later in this chapter.

While basically similar to the frequency multiplying circuits in transmitters, the power output necessary for the satisfactory operation of a mixer is far smaller than would be required for transmitting applications and in consequence greater multiplication factors per stage can be realized.

It is always good practice to start with a reasonably high frequency crystal requiring fewer stages of multiplication

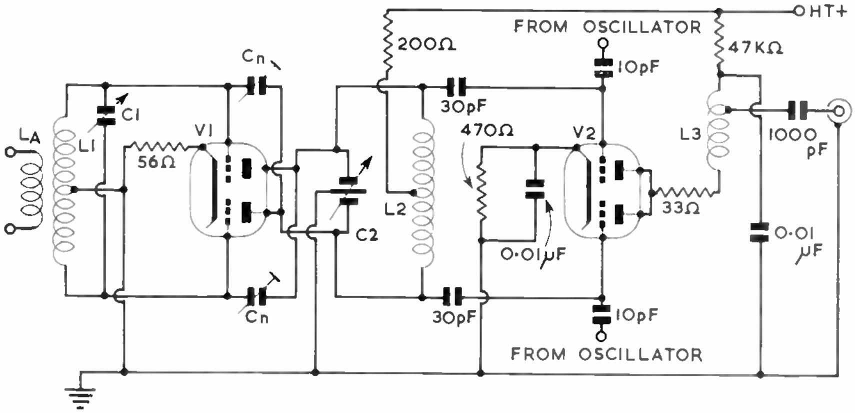


Fig. 5.14. Push-pull r.f. and mixer stages.

70 Mc/s

LA, 3 turns thin p.v.c. covered wire wound round L1.
L1, 9 turns 18 s.w.g. ½ in. diam., ½ in. long.
L2, 11 turns 18 s.w.g. ½ in. diam. self-supporting, 1¼ in. long.
L3, to suit i.f. in use.
C1, 4·7 pF.
C2, not used.
C_n, 0·5 – 4 pF.
V1, 2, 6J6.

144 Mc/s

LA, 3 turns thin p.v.c. wire wound round L1.
L1, 8 turns 20 s.w.g. ¼ in. outside diam. spaced one wire diam.
L2, 6 turns 20 s.w.g. ¼ in. diam. self-supporting approx. 1 in. long.
L3, to suit i.f. in use.
C1, not used.
C2, 15 + 15 pF split stator.
C_n, 0·5 – 4 pF.
V1, 2, 6J6.

rather than the reverse because there is a distinct possibility that one of the unwanted multiples of the crystal frequency might reach the mixer with sufficient power to convert—perhaps inefficiently—a strong local signal on a frequency right outside the band and cause its appearance within the range of the tunable i.f.

An example may make this point clear. An overtone crystal is used to produce a frequency of 28 Mc/s. This is multiplied by five in a subsequent stage to provide injection for the mixer at 140 Mc/s and a tunable i.f. of 4 to 6 Mc/s then covers the 144 to 146 Mc/s band. Although the output of the multiplier stage is tuned to five times its input frequency there may be sufficient output 84 Mc/s (multiplication of three), to beat with one of the BBC frequency modulated transmissions, say the Light Programme from Wrotham on 89·1 Mc/s, and so cause the f.m. signal to appear at 89·1 less 84·0 = 5·1 Mc/s, which could be the i.f. produced by an amateur station operating on 145·1 Mc/s. It will be appreciated that a similar result might obtain near another of the f.m. stations although the actual frequencies concerned would be different.

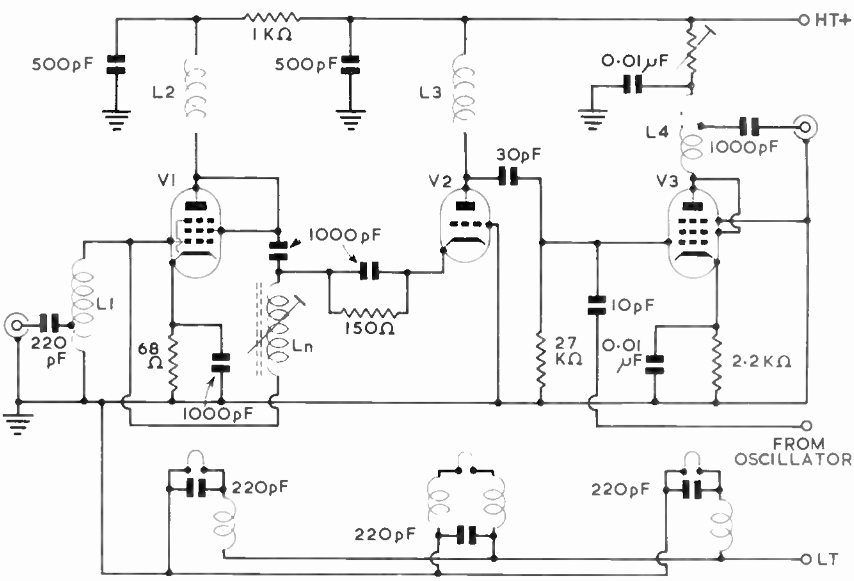


Fig. 5.15. The Wallman cascode circuit.

70 Mc/s

L1, 10½ turns 18 s.w.g. ½ in. inside diam. ⅞ in. long, tapped 4¼ turns from earthy end.
L2, 6 turns 20 s.w.g. on ⅜ in. diam. Aladdin former with dust iron core.
L3, 5½ turns 20 s.w.g. close wound on ⅜ in. diam. Aladdin former with dust iron core.
L_n, 16 turns 26 s.w.g. close wound on ⅜ in. diam. Aladdin former with dust iron core.
Heater chokes, 40 in. 28 s.w.g. enam. close wound on $\frac{3}{16}$ in. former.

144 Mc/s

L1, 5 turns 16 s.w.g. ⅜ in. inside diam. tapped 3 turns from earthy end.
L2, 4 turns 16 s.w.g. ⅜ in. inside diam.
L3, 3½ turns 16 s.w.g. ⅜ in. inside diam.
L_n, 11 turns 26 s.w.g. enam. close wound on ⅜ in. Aladdin former with dust iron core.
Heater chokes, 20 in. 28 s.w.g. enam. close wound on $\frac{3}{16}$ in. former.

V1, 3, 6AK5; V2, 6J4, EC91, 6L34 or half 6J6 (these types are not interchangeable). The pre-set variable resistor in the anode circuit of V3 should be adjusted for best signal-to-noise ratio and a fixed resistor of suitable value substituted. Initially, it may be 100 K ohms.

The procedure in such a case would be to make the output circuit of the frequency multiplying stage more selective, by improving its Q, so that the response at 84 Mc/s was reduced as much as possible, and then to couple to the mixer by means of a link winding rather than via a capacitor, thus attenuating the unwanted frequency still further.

Valves suitable for frequency multiplication are double-triodes, such as the 12AT7, small r.f. pentodes of the 6AK5 or EF91 (6AM6) class and triode-pentodes designed for v.h.f. frequency changer service such as the ECF80 (6BL8), ECF82 (6U8) and ECF804, the latter being particularly suitable due to the high mutual conductance of both the triode and pentode sections (11 and 7·2 mA/V respectively). The 6J6 (ECC91) is not so suitable for use in a crystal oscillator/frequency multiplier circuit because the two triode sections share a common cathode and cathode bias could not be employed. The use of a cathode bias resistor is not essential in a frequency multiplying stage, however, as the necessary working grid bias may be obtained by rectification of the drive voltage between grid and cathode, but with such an arrangement there would be no protective bias should the crystal fail to oscillate.

R.F. AMPLIFIERS

Although a large number of circuits for r.f. stages have been described from time to time, each with its advantages and disadvantages, many of them are only slight modifications of the better known arrangements. For simplicity only the three basic types are discussed here: they are all suitable for use on frequencies below about 150 Mc/s. Which ever is built, it is strongly recommended that a copper or brass chassis be used to ensure the best possible connection for earth returns and a consequent reduction in circulating currents in the chassis and therefore better stability and gain.

No over-riding qualities can be claimed for any one of these circuits: each has its attractive features, and indeed improved performance from one circuit is more often the result of a fortuitous combination of layout or choice of optimum component values than a definite technical superiority. So far as the newcomer to v.h.f work is concerned the simpler the circuit the greater the chance of success.

The choice of push-pull or single-ended input to the converter depends largely on whether a balanced or unbalanced aerial feeder is employed, and although a balanced-to-unbalanced transformer (balun) could be introduced to eliminate losses due to standing waves on the feeder, it may be preferable to avoid this complication by using an input circuit to suit the type of feeder in use.

Push-Pull 6J6 R.F. Amplifier

Fig. 5.14 shows the circuit of the front-end of a converter employing a push-pull circuit arrangement in both the r.f. and mixer stages, the object being to reduce the capacitance across the tuned circuits by effectively putting the input and output capacitances of the valves in series and to cancel out the effect of the cathode lead inductance. Some additional gain is achieved by the step-up in signal voltage between the aerial input and the grids of the r.f. stage.

With the addition of a push-pull oscillator such as the Kaliatron arrangement (Fig. 5.6) this circuit may be used as a complete converter; alternatively the r.f. stage alone may be built as a separate pre-amplifier for use with other converters. If this procedure is adopted an output link winding may conveniently be coupled to the centre of L2 and the inductance of this coil adjusted to maintain resonance.

The layout of components should match that of the circuit diagram closely and all leads should be of the shortest possible length. To obtain the full advantages from this circuit every effort should be made to achieve complete symmetry in both the electrical and the mechanical features. A copper screen should be fixed across the chassis in such a position that the grid and anode circuits of the r.f. stage are shielded from each other. This screen should fit across the valveholder and be soldered to the centre lug and to the earthed heater pin.

As the amplifier valve is a triode it is necessary to neutralize the feedback between anode and grid circuits, not only to maintain stability but to achieve the optimum signal-to-noise ratio. For the neutralizing capacitors miniature air spaced trimmers of the type made by Wingrove and Rogers and having a capacitance range of 0·5–4·5 pF are suitable.

Neutralizing is best accomplished by tuning to a strong signal and disconnecting the h.t. supply to the r.f. stage. The two neutralizing capacitors may then be adjusted in step, keeping the capacitances as nearly equal as possible, until the strength of the signal is reduced to a minimum, L1 and L2 being adjusted for maximum signal after each alteration of the neutralizing capacitors. The point of balance is quite sharp and care taken in this task will go far towards obtaining a good signal-to-noise ratio: for best results in this respect, however, a noise generator may be found more satisfactory. A detailed explanation of the use of noise generators is given in Chapter 19 (*Measurements*).

The 0·01 μF cathode bypass capacitor in the mixer cathode circuit serves as a short circuit for the i.f. currents and should be of the non-inductive type.

The Cascode R.F. Amplifier

In the circuit shown in **Fig. 5.15** the first stage, which is a triode-connected 6AK5, operates at approximately unity gain but enables an effective impedance match to be obtained between the signal input circuit and the cathode input circuit of the grounded grid triode stage V2. Neutralizing is necessary not so much to prevent instability in the first valve as to obtain the best noise factor, and is effected inductively by L_n. Owing to the heavy loading of L2 by the low input impedance of V2, the tuning of this circuit is very flat and its adjustment may therefore appear to be unimportant: however, much of the good performance of which the circuit is capable will be lost if care is not taken, preferably with the aid of a noise generator, to arrive at the optimum adjustment of L2 and L_n.

The coils L1 and L2 should be shielded from one another by a screen mounted across the first valveholder. The neutralizing inductance L_n should be mounted through a hole in this screen with a clearance of approximately $\frac{1}{8}$ in. all round and at right angles to L1 and L2.

With h.t. removed from V1 and a strong incoming signal, neutralization should be carried out by adjusting L_n for minimum response while keeping L1 and L2 tuned for maximum signal. Care should be taken to ensure that the anode circuit of V2 is tuned to the signal frequency and not to the image on the wrong side of the oscillator frequency. If it is tuned to the image frequency any adjustment of L3 will cause a decrease in signal strength whereas the background noise should rise to a distinct peak when L3 is tuned to the correct frequency. When the circuit is operating satisfac-

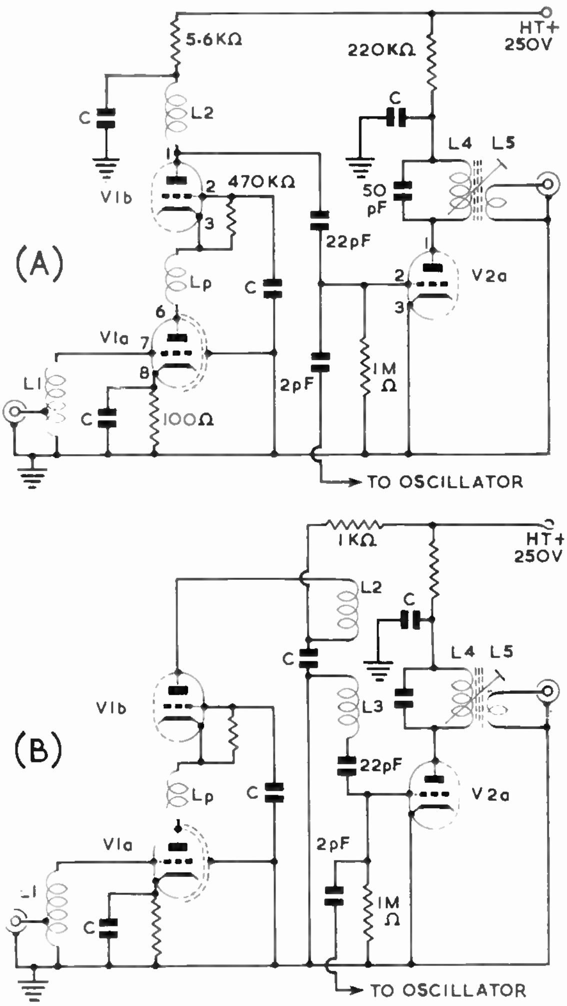


Fig. 5.16. Series cascode or driven grounded grid amplifiers with associated mixer stage. The arrangement in (B) gives appreciably more gain. All values in the two circuits are similar unless otherwise stated. All coils are $\frac{3}{8}$ in. diam. self-supporting and wound with 18 s.w.g. L1, 5½ turns tapped 2½ turns from earthy end; L2, 3½ turns (circuit (A)), 7½ turns (circuit (B)); L3 (circuit (B) only), 4½ turns: coupling approximately $\frac{1}{16}$ in. from L2, L4, L5, to suit i.f. in use; Lp, 5-7 turns 28 s.w.g. enam., $\frac{3}{16}$ in. diam. (exact size determined with the aid of noise generator); V1 6BZ7, 6BQ7A, ECC85 (with reduced h.t.); V2, 12AT7 (the other section may be used as the local oscillator). All capacitors marked C are 1000 pF hi-K feedthrough type.

torily the removal of V1 from its valveholder should result in a marked drop in the noise level.

The circuit gives slightly more gain than does the push-pull 6J6 arrangement and it is possible to achieve a noise factor of 4db at 145 Mc/s.

Series Cascode or Driven Grounded Grid Circuit

This is one of the simpler types of r.f. amplifier circuit but is nevertheless capable of excellent results. Two versions are shown in **Figs. 5.16 (a)** and **(b)**. The latter possesses an additional tuned circuit (L3) and gives appreciably higher gain but with some reduction in bandwidth although this is still adequate for complete coverage of the 2m band. No screening is necessary and there is no variable tuning, the inductances being adjusted by spreading or closing the turns. A grid dip oscillator is a useful aid although final adjustment should be made on incoming signals or with a noise generator.

As in any v.h.f. circuit, only top quality bypass capacitors are recommended and they should preferably be of the feed-through type soldered to the chassis. All r.f. leads should be as short as possible and may, with advantage, be thin strips of copper foil rather than wire in order to reduce the inductance.

The Grounded Grid Amplifier

With any triode r.f. amplifier some means must be adopted to prevent instability due to feedback between the input and output circuits. In the grounded grid configuration it is the grid, rather than the cathode, which is maintained at zero r.f. potential thus providing a screen between the cathode and anode which form the input and output electrodes respectively. Due to the low impedance of the cathode input circuit the grounded grid amplifier is by nature a wideband device and is thus well suited as a 144–146 Mc/s r.f. stage pretuned to the centre of the band.

The input impedance of a grounded grid triode is equal to *1/gm* where *gm* is the mutual conductance of the valve in *amps* per volt, so for *gm* in the range 6-15 mA/V this would imply input impedances between 167 and 67 ohms.

A complete 144 Mc/s converter employing two grounded grid triode r.f. amplifier stages is described later in this chapter.

Valves for V.H.F. Operation

Valves specially designed for grounded grid service include the 6J4, EC91 and 6L34 on the B7G base and later types capable of operating at frequencies of the order of 900 Mc/s—6AJ4, 6AM4, 6AN4, ECC88 and GEC A.2521 (6CR4), all on the B9A base. The GEC A.2599 (6CT4) and the RCA Nuvistor 6CW4 are excellent valves designed for the grounded cathode configuration.

TYPICAL TWO AND FOUR METRE DESIGNS

RSGB CONVERTER CIRCUIT

This converter, using a double triode valve as a combined grounded grid r.f. amplifier and mixer, was designed primarily as a simple circuit for the newcomer to v.h.f. Despite its simplicity it is, however, capable of satisfactory performance and is in use in many stations. The circuit is shown in **Fig. 5.17.**

One section of a 12AT7 double triode valve (V1) is connected as a grounded grid triode r.f. stage with the input

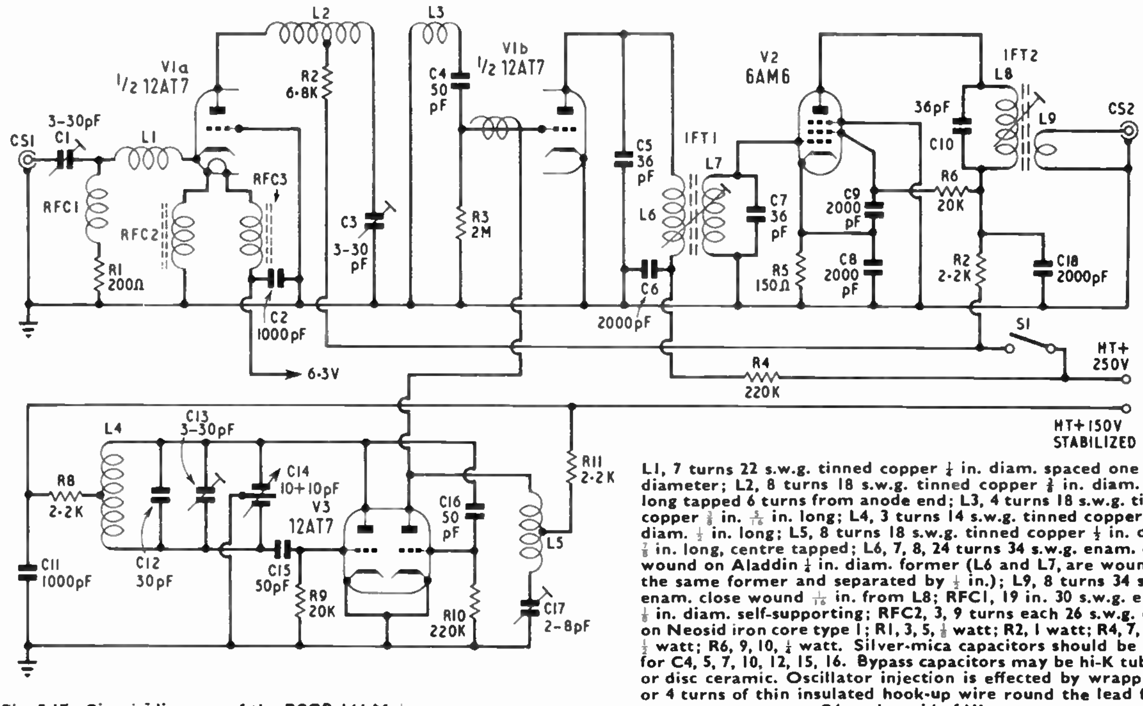
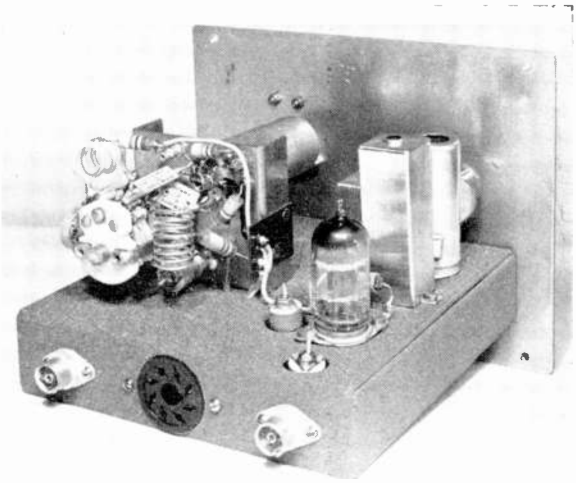


Fig. 5.17. Circuit diagram of the RSGB 144 Mc/s converter.

L1, 7 turns 22 s.w.g. tinned copper $\frac{1}{4}$ in. diam. spaced one wire diameter; L2, 8 turns 18 s.w.g. tinned copper $\frac{3}{8}$ in. diam. $\frac{3}{4}$ in. long tapped 6 turns from anode end; L3, 4 turns 18 s.w.g. tinned copper $\frac{3}{8}$ in. $\frac{5}{16}$ in. long; L4, 3 turns 14 s.w.g. tinned copper $\frac{3}{8}$ in. diam. $\frac{1}{2}$ in. long; L5, 8 turns 18 s.w.g. tinned copper $\frac{1}{2}$ in. diam. $\frac{7}{8}$ in. long, centre tapped; L6, 7, 8, 24 turns 34 s.w.g. enam. close wound on Aladdin $\frac{1}{4}$ in. diam. former (L6 and L7, are wound on the same former and separated by $\frac{1}{2}$ in.); L9, 8 turns 34 s.w.g. enam. close wound $\frac{1}{16}$ in. from L8; RFC1, 19 in. 30 s.w.g. enam. $\frac{1}{8}$ in. diam. self-supporting; RFC2, 3, 9 turns each 26 s.w.g. d.c.c. on Neosid iron core type 1; R1, 3, 5, $\frac{1}{8}$ watt; R2, 1 watt; R4, 7, 8, 11, $\frac{1}{2}$ watt; R6, 9, 10, $\frac{1}{4}$ watt. Silver-mica capacitors should be used for C4, 5, 7, 10, 12, 15, 16. Bypass capacitors may be hi-K tubular or disc ceramic. Oscillator injection is effected by wrapping 3 or 4 turns of thin insulated hook-up wire round the lead from C4 to the grid of V1b.



Rear view of the chassis showing the r.f./mixer valve in right foreground with trimmers C1 and C3. The oscillator/doubler assembly is to the left: C17 may be seen between the doubler coil L5 and the chassis.

circuit arranged to couple 80–100 ohm coaxial feeder to the cathode impedance. L1 serves to match this impedance to the cathode input circuit of V1a which is approximately 150 ohms. The prime function of the 3–30 pF trimmer capacitor C1 is to prevent short circuiting of the bias for V1a when the aerial feeder has d.c. continuity, but it also helps to tune out small amounts of reactance due to an imperfectly matched input system.

The coupled inductances L2 and L3 form an impedance matching network between the output of the g.g.t. stage and the triode mixer and it is here that sufficient gain is realized to provide adequate sensitivity and a good noise factor. L2 is resonated at the centre of the band by C3 and L3 by its inductance and the input capacity of the mixer. The degree of coupling between these coils determines the bandwidth.

To ensure the best signal-to-noise ratio it is necessary to run V1b at a low anode voltage, hence the high value of R4. By so doing, the gain of the stage is reduced to a certain extent, but as the noise falls more rapidly than the signal for reductions down to approximately 45 volts a net gain in the readability of weak signals is obtained.

C5 forms a bypass for signal and oscillator frequencies present at the mixer anode and as it is part of the tuning capacity for the primary of the i.f. transformer in conjunction with C6 it is possible to use a relatively large value without fear of bypassing the i.f. signal. To be fully effective in its dual role, C5 should be of good quality and connected by short leads between the anode tag of the valveholder and the common earthing point of the stage.

The i.f. amplifier (V2) is a 6AM6 but any r.f. pentode of similar characteristics could be employed in this position. The exact intermediate frequency is not important provided it is above 5 Mc/s; with the transformers described it is 10·2 Mc/s.

A second 12AT7 (V3) is used as a Hartley oscillator/doubler, the injection voltage for the mixer being taken from the anode of the doubler section. The injection frequency is on the *low* side of the signal. A split-stator tuning capacitor is used so that, as the rotor is left free, no r.f. current flows through the bearings, a major cause of noise and frequency jumping in v.h.f. oscillators. H.T. voltage to L4 is fed through

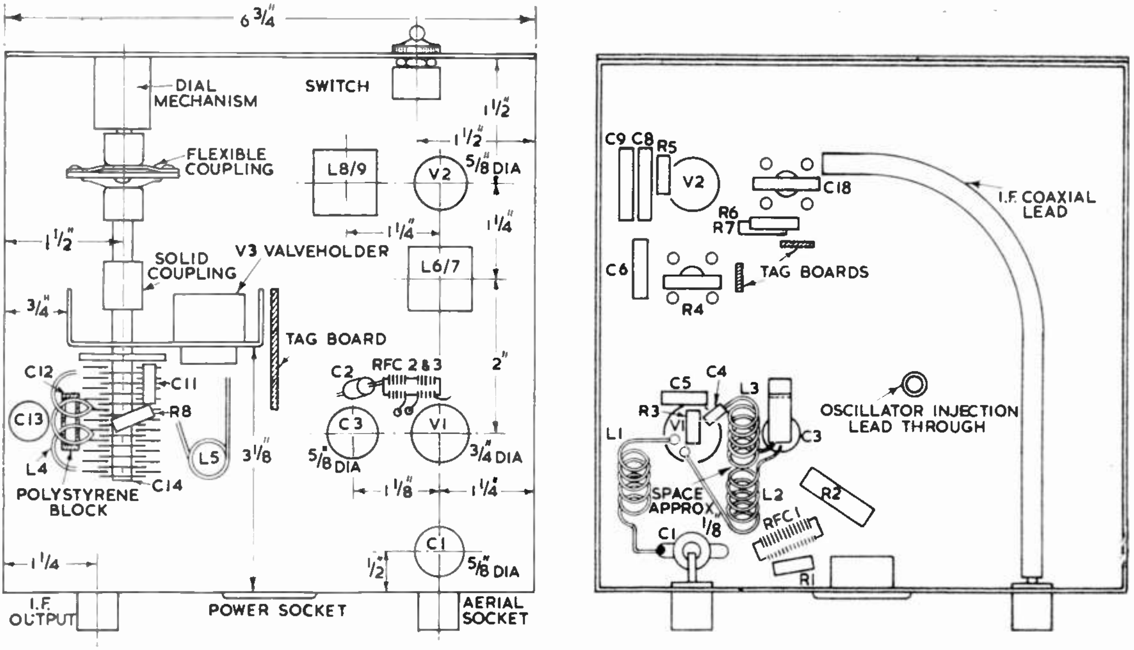


Fig. 5.18 (Left). Top view of the chassis showing the arrangement of the principal components. (Right). Layout of the components below the chassis.

R8 to the centre tap of the coil. It should be noted that this point is *not* bypassed to chassis, to do so would probably result in instability and a poor note.

The grid of the doubler section of V3 is driven via C16, bias being produced across the grid leak R10. The anode circuit comprises L5, series tuned by the 8 pF Philips trimmer C17.

Construction

The layout is shown in **Fig. 5.18** and is quite straightforward. The 30 pF concentric trimmer (C1) has its spigot bound to the centre conductor of the aerial socket with fine wire before soldering and is adjustable through the hole provided in the deck of the chassis. When making this joint care must be taken to avoid overheating, as both components are adversely affected by excessive heat. If the aerial plug is in position during the operation the pin of the socket will remain central, but the use of heat shunts in the form of crocodile clips is recommended.

C3 is mounted on a small brass bracket and protrudes through the top of the chassis. L2 is supported at one end by the anode tag of the valveholder (pin no. 1) and at the other is soldered to a lug on the body of C3. The earthy end of L3 is soldered to the bracket close to the point where the spigot of C3 passes through, the two coils being arranged so that their inner faces are about ⅛ in. apart. The other end of L3 is attached to the mixer grid capacitor C4 and no further support is necessary. The oscillator injection lead may be wrapped three times round the grid lead at this point.

The two r.f. chokes (RFC2 & 3) in the heater leads to V1 are mounted above the chassis and wound side by side on a Neosid iron core, the outer ends of the coils being soldered to the lead-out wires which then serve as supports: one end is connected to a tag placed under a bolt securing the p.t.f.e. valveholder for V1 and the other to the top of the bypass capacitor C2.

In the interests of freedom from pick-up of signals at the intermediate frequency, all leads to the i.f. amplifier should be short and the internal connections to the i.f. transformers should be so arranged that the external connections are as direct as possible. The shielding of the coaxial lead from L9 to the output socket should be connected to chassis at both ends though its length is unimportant.

All power supplies are taken through an octal plug and socket on the rear drop of the chassis. A switch is provided on the panel which, in its OFF position, removes h.t. from the r.f. and i.f. stages during transmission, leaving the mixer and oscillator operative so that it is possible to monitor a transmitter without overloading either the converter or the associated receiver.

Adjustment

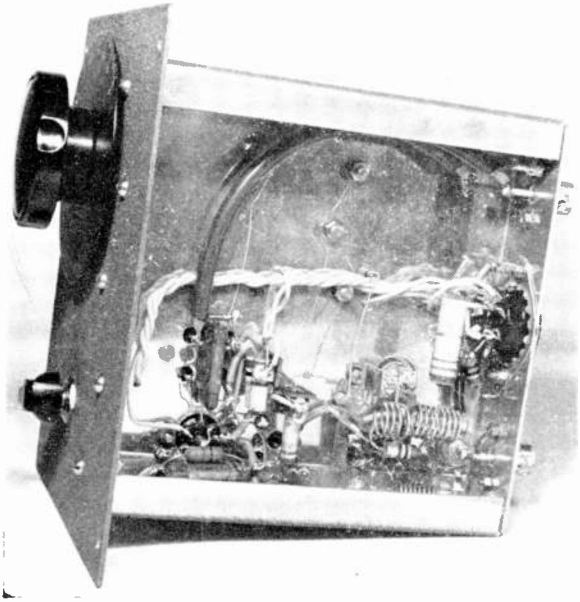
Set the main receiver to approximately 10·2 Mc/s and adjust the cores of L6, L7 and L8 for maximum noise. The exact frequency is not important and some latitude is possible to avoid unwanted signals at the intermediate frequency. The gain from V2 is such that the r.f./i.f. gain control setting in the main receiver will be quite low, resulting in no background noise being audible when the send/receive switch on the converter is at the SEND position.

The next step is to resonate L3. With no h.t. on the r.f. stage, the oscillator should be adjusted to give injection at 145 Mc/s. (This may require the temporary removal of C12). L5/C17 should tune to 145 Mc/s just short of minimum capacity. A meter with a f.s.d. of about 250 volts is then connected between the mixer anode and chassis and L3 adjusted by opening or closing the turns until the *highest* meter reading is obtained. C3 should be detuned to avoid inter-action between L2 and L3 and care taken that C17 maintains resonance in the output circuit of the doubler. C13 is then adjusted to set the oscillator to a frequency of (145 – i.f.)/2, or 67·4 Mc/s with the i.f. mentioned above. This may be done with the aid of a grid dip oscillator or a closely calibrated absorption wavemeter. In the latter case it will be necessary to monitor the anode current of the oscillator and watch for the "kick" when L4 passes through resonance.

The doubler anode circuit of V3 should be set to the required injection frequency, i.e. twice that of the oscillator. If the circuit values are as shown, the only resonance point within the range of C17 will be the correct one. This will be indicated by a voltmeter connected between mixer anode and chassis as already described. By adjusting the spacing of L4 and the capacity of C13 the 144–146 Mc/s band may be spread over the entire dial.

H.T. may next be applied to the r.f. and i.f. stages and C3 tuned to resonance, which will be indicated by a considerable rise in the noise level. It is essential that the aerial be connected to load V1a. It may be necessary to open or compress the turns of L2 so that resonance occurs at about mid-capacity of C3. Instability may be produced at low capacity settings of C3, but if this happens make sure by rotating C1 that this is not due to incorrect aerial loading.

All preliminary tuning adjustments should be made at mid-scale on C14 and if this capacitor is now tuned over its range the background noise should vary only slightly over the band. If there is a sharp falling off, tighter coupling



An underchassis view of the 144 Mc/s converter. L1 and L2 are in the right foreground. The co-axial lead feeds the i.f. signal from L9 to the output socket CS2 at the top right.

between L2 and L3 is required, followed by readjustment of C3 at 145 Mc/s.

C1 is not a tuning control and no resonance effects should be found when it is adjusted, but at low capacity settings the grounded grid stage will almost certainly go into oscillation. Slight adjustments to the inductance of L1 may be necessary and, if a noise generator is not available, these should be made for the best readability of a *weak* 'phone signal on approximately 145 Mc/s. With adjustments made as described there should be no difficulty in obtaining really satisfactory performance. The converter must not be connected to a balanced feeder without the addition of a balance-to-unbalance transformer or balun.

With this or any other converter the effects of varying the tuning or coupling adjustments as well as the h.t. voltage applied to the r.f. and mixer stages and the oscillator injection may best be examined with the aid of a noise generator, but although this is advisable to ensure optimum results it is possible to obtain an acceptable performance without this aid. It is most important to ensure that the heater voltage actually applied to the valve is in accordance with the maker's rating.

SIMPLE HIGH PERFORMANCE CONVERTERS FOR 70 AND 144 MC/S USING NUVISTOR R.F. STAGES

These converters, although reasonably simple to construct even by a beginner, give a performance very close to the best available but do not use any expensive valves.

As many operators use the RCA AR88D and similar communication receivers as a tunable i.f., relatively low frequencies are used; 2·1–2·7 Mc/s (for the 70 Mc/s converter) and 2–4 Mc/s (for the 144 Mc/s converter) where the AR88D has excellent bandspread and stability. A different intermediate frequency can be employed if desired and details are given later on this point.

Precautions have been taken to eliminate breakthrough, which can be troublesome at such low i.f.'s, and these (which include the use of a screen over the oscillator crystal) have proved effective.

Basic Circuitry

Both converters use an earthed cathode, capacity neutralized, 6CW4 Nuvistor r.f. amplifier inductively coupled into a triode mixer, in the anode circuit of which is connected a suitable i.f. tuning coil. To provide a low impedance for use with a coaxial cable, the output is fed to the main receiver through a triode cathode follower.

Whilst the converters are basically similar, the crystal oscillator/multiplier stages differ. In both instances the oscillator valve is an EF91 (6AM6) in a simple Colpitts circuit. This is to be preferred to the Squier oscillator which may prove difficult to use, particularly with surplus crystals.

In the case of the 70 Mc/s converter, an 8500 kc/s FT243 crystal is used, the oscillator anode circuit being tuned to the fourth harmonic of the crystal frequency. The output of this stage is capacity coupled to the grid of the EF91 doubler

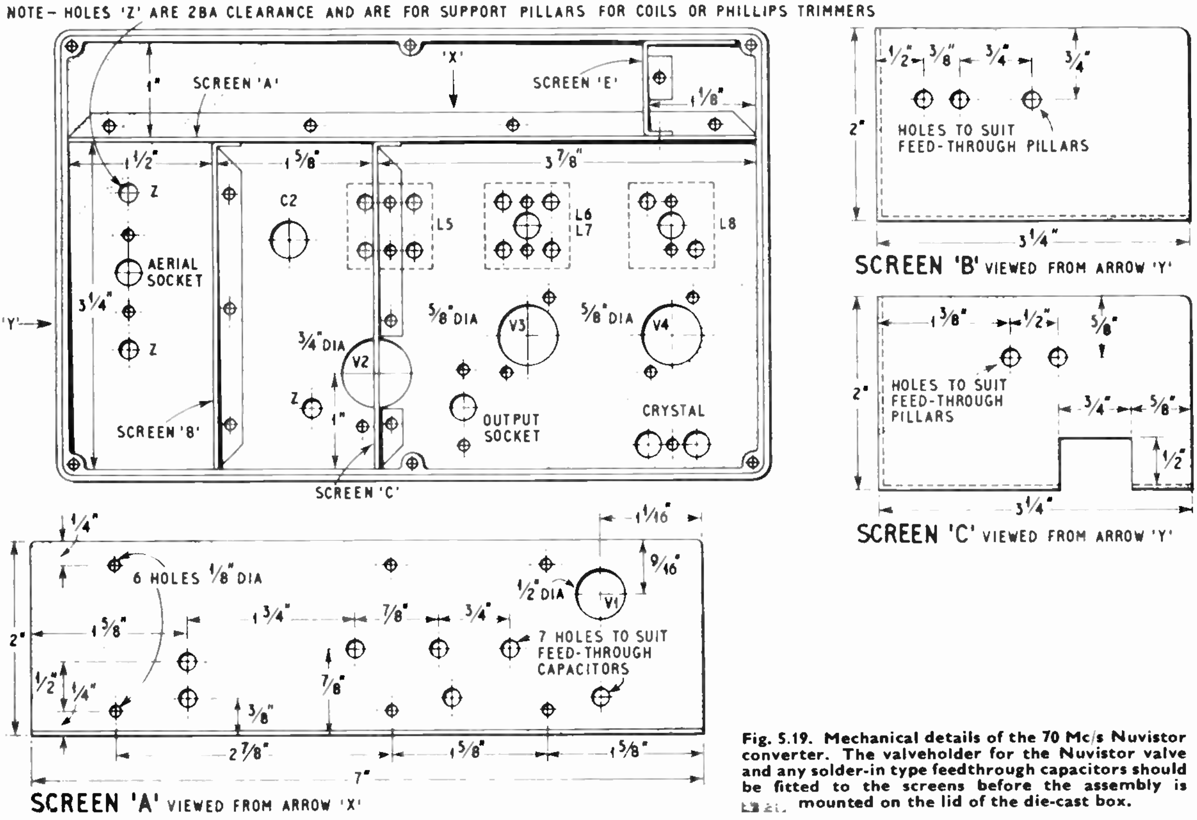


Fig. 5.19. Mechanical details of the 70 Mc/s Nuvistor converter. The valveholder for the Nuvistor valve and any solder-in type feedthrough capacitors should be fitted to the screens before the assembly is mounted on the lid of the die-cast box.

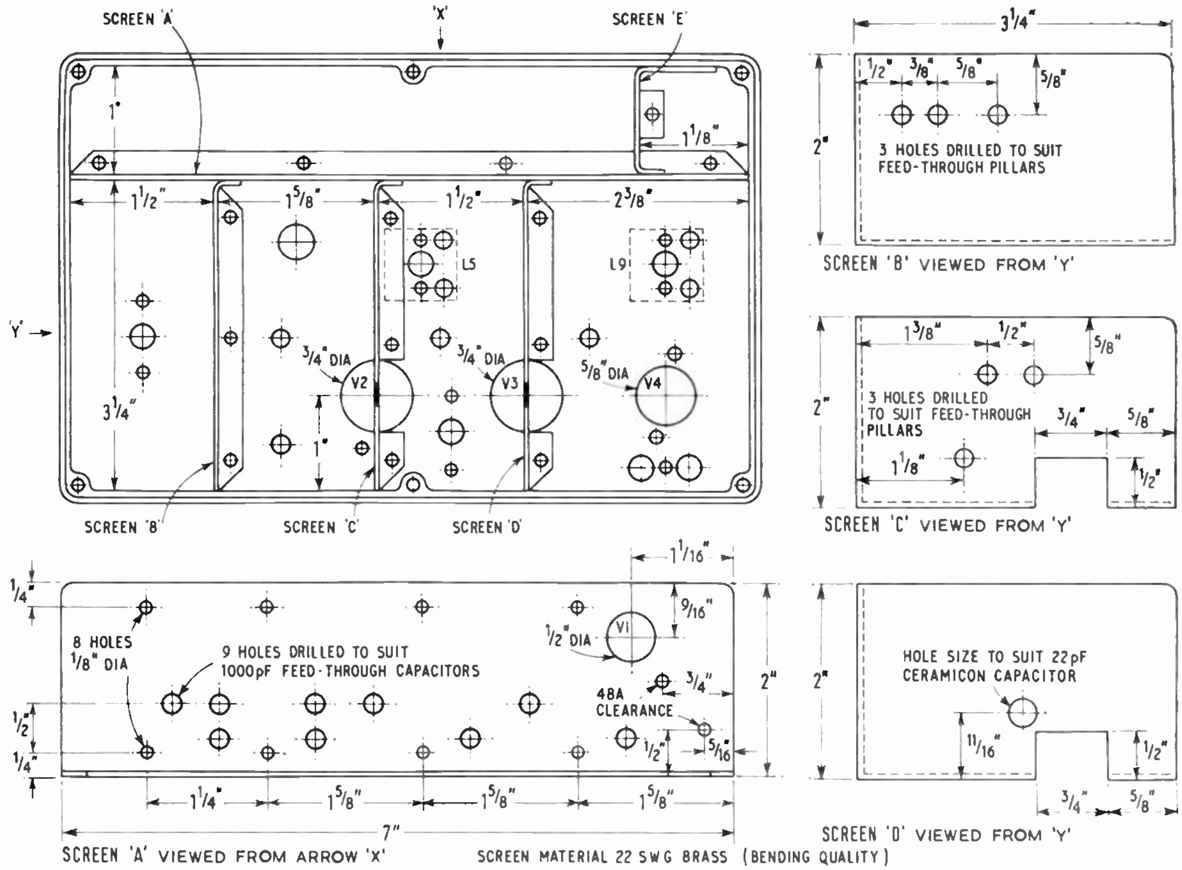


Fig. 5.20. Mechanical details of the 144 Mc/s converter.

stage, the anode of which is tuned to a frequency of 68·0 Mc/s, hence producing an i.f. of 2·1–2·7 Mc/s.

For the 144 Mc/s converter, a 7100 kc/s FT243 crystal is used but the oscillator anode circuit is tuned to the fifth harmonic of the crystal. If V4 fails to oscillate, reduce the value of the 2·2 K ohms cathode resistor; poor activity crystals may require a value as low as 100 ohms. The subsequent ECC81 (12AT7) double triode is connected in cascade and doubling in each half produces a final frequency of 142 Mc/s which results in an i.f. of 2–4 Mc/s.

In both converters the output from the local oscillator is link coupled to the mixer stage and more than sufficient output is available to ensure efficient mixing. No significant spurious responses from the crystal oscillator chain have been found within the tuning range of the converters.

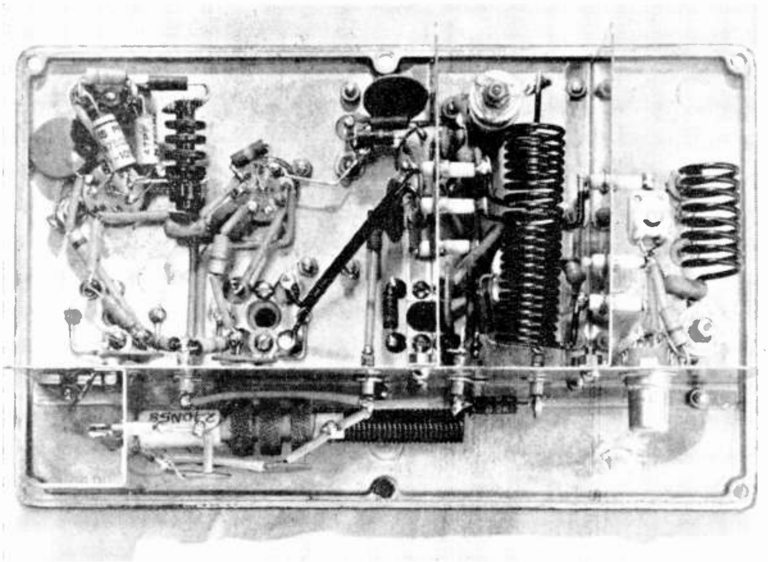
Mechanical Construction

Each converter is built on the lid of an Eddystone 7½ in. × 4½ in. No. 845 die-cast box. This ensures easy access for assembly and wiring. The construction of the interior screens, with dimensions, is shown in Figs. 5.19 and 5.20. If solder-in type feedthrough capacitors are used, these, together with the Nuvistor holder, should be soldered into position before attempting to bolt the screen sections together and to mount the assembly on the lid.

The only component mounted through the body of the box is the Bulgin P360 three-way miniature power input socket, a flying lead from which carries the power to the heater and h.t. feedthrough capacitors mounted on screen "*E*." This screen also serves to support RFC1 and its associated 0·01 μF ceramic capacitors. RFC2 is soldered directly to the heads of the appropriate feedthrough capacitors. Other aspects of construction may be seen from the photographs, but it should be noted that in the photograph of the top of the completed 144 Mc/s converter the aforementioned power socket is shown at the wrong end of the box.

Components and Wiring

Almost all the components used in both converters are standard types readily available. All valveholders should be of the low loss variety, preferably p.t.f.e. A special holder, Cinch Type 133-65-10-001, is required for the 6CW4 Nuvistor and is generally available from dealers stocking this type of valve. As the 6CW4 has a very low anode-grid capacity, it is essential that the neutralizing capacitor C_N should have a minimum capacity not greater than 1 pF. A Wingrove and Rogers or Plessey miniature 1-10 pF air-spaced trimmer is suitable. The Philips concentric type should not be used due to its relatively high minimum capacity.



An interior view of the 70 Mc/s converter. Note the use of ceramic feedthrough insulators for supporting the air-spaced coils.

All 0·01 μF capacitors are high K ceramic types. The 1000 pF feedthrough ceramic capacitors can either be the solder-in or nut secured types. All other fixed capacitors are ceramic, suitable types being marketed by Radiospares Ltd. All resistors are ¼ watt type except the 8·2 K ohms h.t. feed for V1, which as a ½ watt rating. A number of KLG feedthrough insulators are used both for support and feed-through purposes resulting in a neat and rigid assembly.

All wiring associated with the r.f. circuitry should be kept as short as possible. Decoupling and earth returns associated with each valve are made to solder tags which are held under the securing nuts for the valveholder concerned. In the case of the Nuvistor, tags 8 and 10 are wired to the frame lugs of the holder. The earth lead from the Bulgin power socket is connected to a solder tag under one of the securing bolts of screen "*E*."

The 70 Mc/s Nuvistor Converter

The circuit is shown in **Fig. 5.21** and mechanical details in Fig. 5.19. Coil details are given in **Table 5.1.**

To align the converter, disconnect the h.t. supply from V1, remove the cans from L7 and L8 and first adjust L8 for maximum indication on the r.f. checking meter—**Fig. 5.22.** Then adjust L7 in a similar manner. The tuning range of each coil is such that only the required harmonic should be selected but it is wise to check the actual frequencies with an absorption wavemeter or g.d.o. When this has been done the cans should be replaced on L7 and L8, and the two cores re-adjusted for maximum r.f. output from the coupling link L4. The output should be connected to the communications receiver, tuned to 2·4 Mc/s. L5 will be approximately

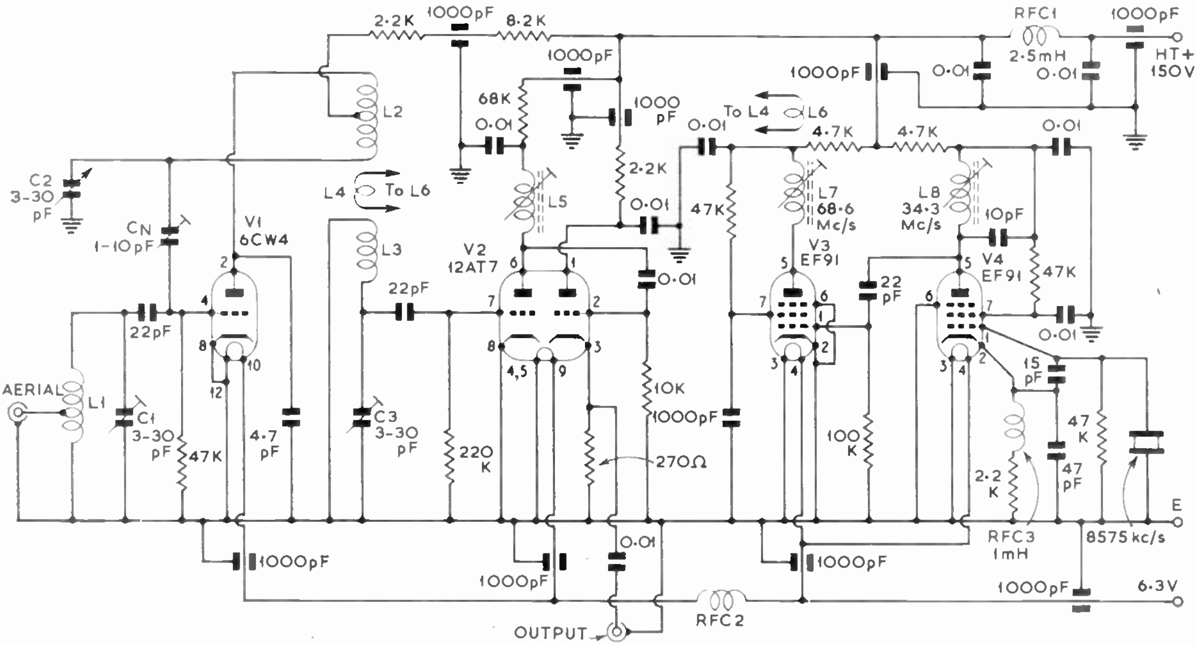


Fig. 5.21. Circuit diagram of the 70 Mc/s Nuvistor converter. A 22 pF ceramic capacitor should be connected from pin 6 (anode) of V2 to earth

TABLE 5.1

Inductor details for the 70 Mc/s Converter
Tunable i.f. 2·1 to 2·7 Mc/s. Crystal 8500 kc/s.

L1	8 turns 18 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length 1 in., tapped 3 turns from earth end, air spaced.
L2	13 turns 18 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length $1\frac{1}{4}$ in., tapped $5\frac{1}{2}$ turns from C_N air spaced.
L3	$8\frac{1}{2}$ turns 18 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length $\frac{7}{8}$ in., air spaced.
L4	2 turns 18 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel and placed between L2 and L3.
L5	Maxi-Q i.f. transformer type IFT 11/1·6, with silver mica capacitors in can removed and primary and secondary windings in series (top of lower layer winding to bottom of upper layer winding).
L6	2 turns 22 s.w.g. p.v.c. covered tinned copper wire wound at h.t. end of L7.
L7	10 turns 26 s.w.g. enam. wound on $\frac{1}{4}$ in. by $1\frac{3}{8}$ in. former, slug tuned (Aladdin type with can).
L8	10 turns 26 s.w.g. enam. wound on $\frac{1}{4}$ in. by $1\frac{3}{8}$ in. former, slug tuned (Aladdin type with can).
RFC1	2·5 mH r.f. choke.
RFC2	51 in. 18 s.w.g. enam. wound on $\frac{1}{4}$ in. mandrel, close spaced.
RFC3	1 mH r.f. choke.

correct. Adjust C3 for maximum hiss; two positions will be found, the one with the smaller capacity being the correct one.

A strong signal is then required (from a local transmitter or signal generator) and should be fed to the aerial socket. Adjust C1 and C2 followed by C3 and L5 for maximum output followed by adjustment of C_N, the neutralizing capacitor, with an insulated screwdriver, for *minimum* output. (In practice this is usually found to be near the minimum capacity of C_N.) This procedure should be repeated several times as there is some interaction between the adjustments.

The initial adjustments are now complete and h.t. may be reconnected to V1. If there is a tendency for oscillation, C_N is incorrectly adjusted; the minimum position referred to is very sharp and to a certain extent adjustment is interdependent with C2 and C1. No difficulty should be experienced, however, in obtaining the correct setting. For final adjustments a noise generator is desirable (see Chapter 19—*Measurements*) but if this is not available, C1, C2, C3 and L5 (also L7 and L8) should be adjusted for maximum output on a local signal. Finally the capacity of C1 should be increased slightly so that the circuit is detuned towards 70 Mc/s and the output just drops. This is near to the optimum position for the best signal-to-noise ratio. In practice a noise factor of better than 2·5db should be obtained.

The 144 Mc/s Nuvistor Converter

The circuit of the 144 Mc/s converter is shown in Fig. 5.23 and the mechanical arrangements in Fig. 5.19.

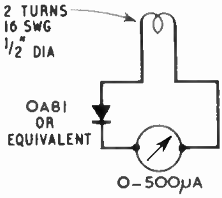


Fig. 5.22. R.F. checking meter

The adjustment of this converter is very similar to the 70 Mc/s unit except that the communications receiver is set to 3 Mc/s, corresponding to a signal frequency of 145 Mc/s. First power is applied with the h.t. disconnected from V1. The can is removed from L9 and the slug is adjusted for maximum r.f. indication followed by C5 (r.f. indication being

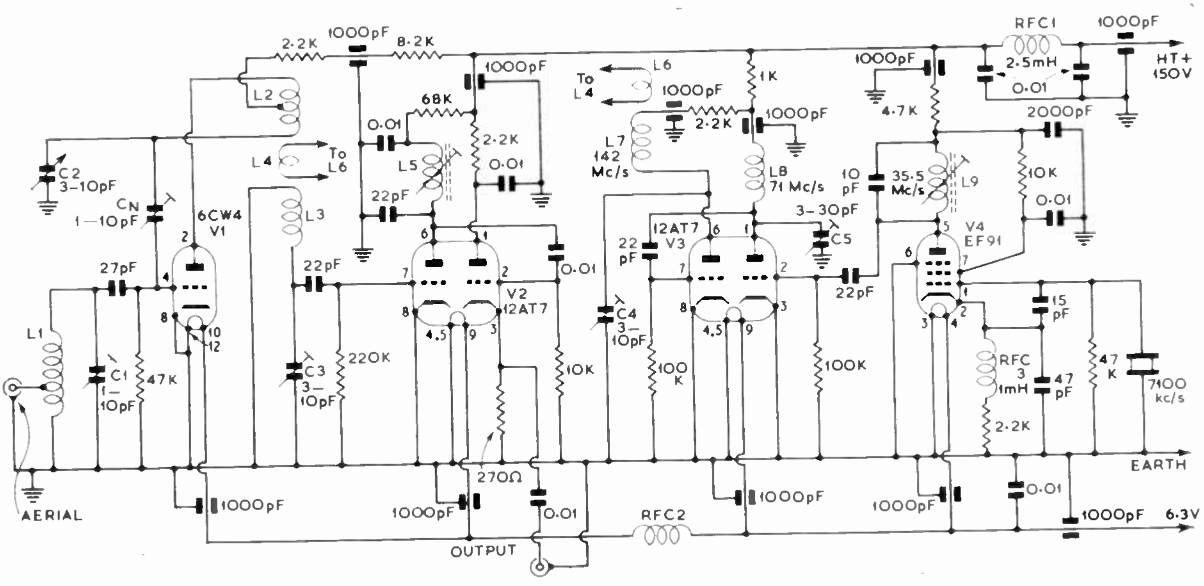
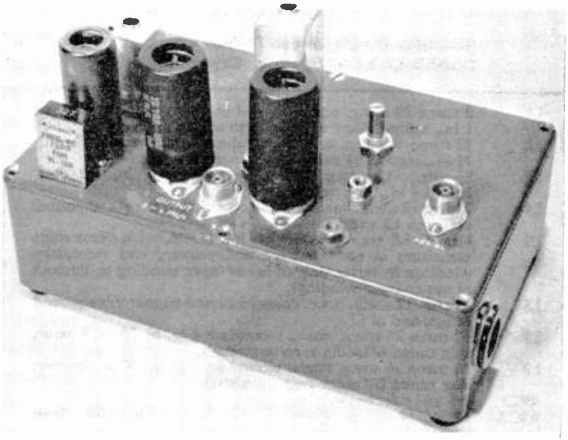


Fig. 5.23. Circuit diagram of the 144 Mc/s Nuvistor converter.

TABLE 5.2

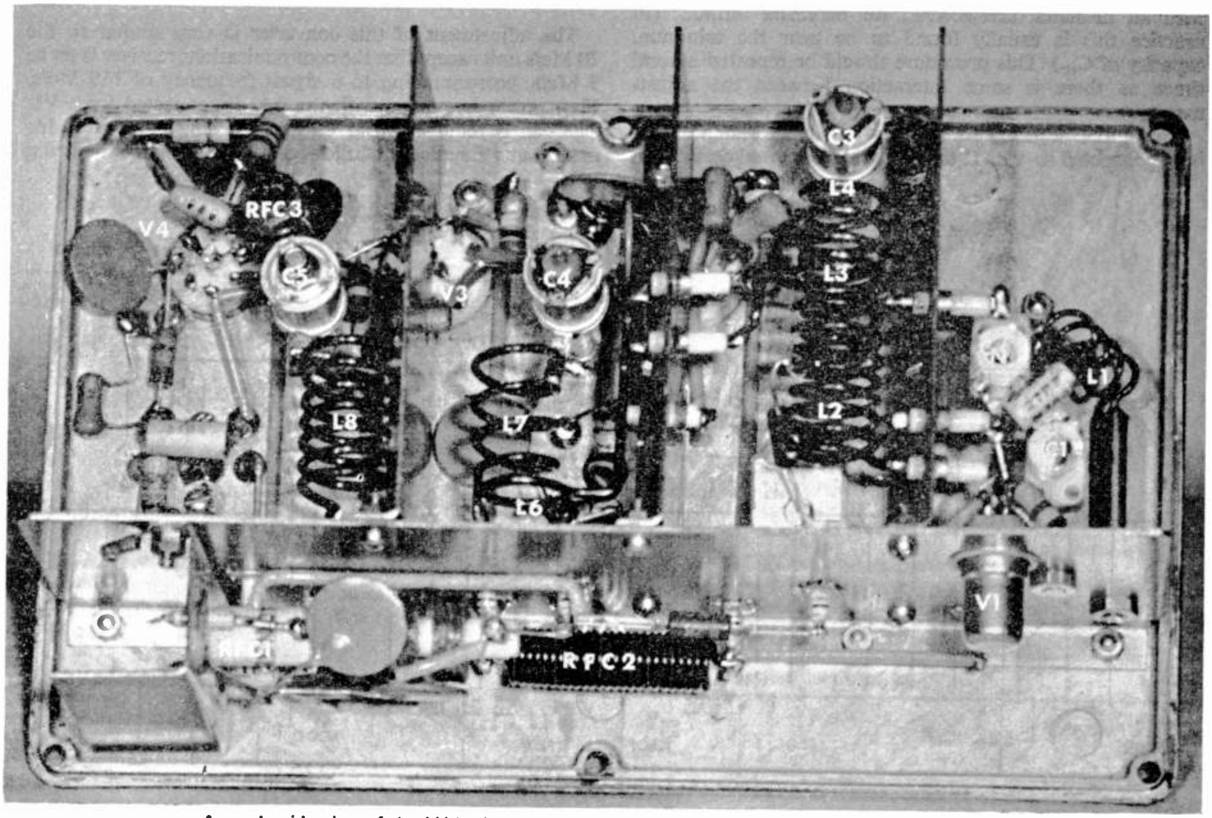
Inductor details for the 144 Mc/s Converter
Tunable i.f. 2 to 4 Mc/s. Crystal 7100 kc/s

L1	4 turns 18 s.w.g. enam. wound on $\frac{1}{4}$ in. mandrel, length $\frac{7}{8}$ in., tapped $1\frac{1}{2}$ turns from earthy end, air spaced.
L2	8 turns 16 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length 1 in., tapped $3\frac{1}{2}$ turns for C_N, air spaced.
L3	4 turns 16 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length $\frac{5}{8}$ in., air spaced.
L4	1 turn 16 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, interwound at earthy end of L3, air spaced.
L5	32 s.w.g. enam. wound on $\frac{1}{4}$ in. by $2\frac{1}{2}$ in. former (Aladdin type with can), 1 layer $1\frac{1}{2}$ in. long and 1 layer $\frac{3}{4}$ in. long. Tissue paper interleaving, secured with polystyrene cement. Tuned by two slugs. 3·3 K ohm resistor in parallel.
L6	as for L4 but mounted adjacent to the h.t. end of L7, air spaced.
L7	$3\frac{1}{2}$ turns 16 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length $\frac{3}{4}$ in., air spaced.
L8	7 turns 16 s.w.g. enam. wound on $\frac{7}{16}$ in. mandrel, length 1 in., air spaced.
L9	10 turns 26 s.w.g. enam. wound on $\frac{1}{4}$ in. by $1\frac{3}{8}$ in. former slug tuned (Aladdin type with can).
RFC1	2·5 mH r.f. choke.
RFC2	26 in. 18 s.w.g. enam. close wound on $\frac{1}{4}$ in. mandrel, self-supporting, air spaced.
RFC3	1 mH r.f. choke.



The 144 Mc's converter. The screening can for the crystal has been removed for this picture.

observed at L8 and L7 respectively), and then C4. Again, the tuning range of each tuned circuit is such that only the required harmonic should be selected but it is wise to check the actual frequencies with an absorption meter or g.d.o., if available. L5, which is very flat in its tuning, should be centred on 3 Mc/s. C1, C2 and C3 are then adjusted for maximum output on a strong local signal. C_N is adjusted for *minimum* output with an insulated screwdriver, the correct setting being very critical. When h.t. has been reconnected to V1 the converter is ready for use. Final adjustment



An underside view of the 144 Mc/s converter. Note that L3 should read L4 and vice versa.

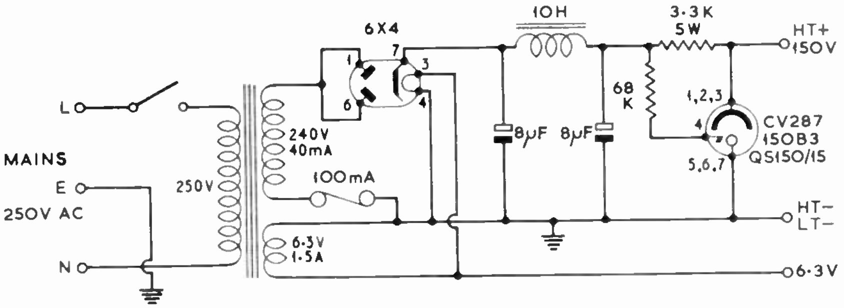


Fig. 5.24. A power supply suitable for use with either the 70 Mc/s or 144 Mc/s converter. A semi-conductor rectifier of suitable rating could be substituted for the 6X4 valve.

should again, if possible, be carried out with a noise generator.

Measured under laboratory conditions the noise factor of the original model was 2·9db, protection against i.f. breakthrough better than 80db, the image response better than 20db down at 144 Mc/s and the overall gain 23db.

Power Supply

A power supply which delivers 6·3 volts at 1·5 amps and 150 volts at 30 mA is adequate and a suitable circuit is shown in Fig. 5.24. The components are all of a modest size and the complete unit can be built into a 3½ in. × 4½ in. Eddystone No. 650 die-cast box. The smoothing capacitors and choke are mounted inside the box. A transmit/receive relay can be added if required to mute the converter whilst a transmitter is operating, by removing the h.t. supply.

Wiring Details

The orientation of the valveholders should be as follows: V2, V3, pins 1 and 9 are on the screen " A " side, V4, pins 1 and 7 are on the screen " A " side when viewed from the underside of the chassis. Earthing of the valveholder pins is important: all pins to be earthed are connected to the centre spigot of the valveholder. For V2 and V3, these are connected by as short a wire as possible, i.e., from the nearest earthed pin, to 6 B.A. solder tags under the valveholder retaining nuts nearer screen " A."

In the case of V4 the earth tag is mounted under the nut nearer the crystal holder. Screens " C " and " D " are also soldered to the centre spigots of their related valveholders.

144 MC/S BAND CONVERTER USING TRANSISTORS

With modern alloy diffusion and mesa epitaxial transistors, compact yet highly efficient converters for the amateur v.h.f. and u.h.f. bands may be constructed. Fig. 5.25 shows the circuit diagram of such a converter for the 144 to 146 Mc/s allocation.

A Siemens AF139 transistor (TR1), operates as a grounded base r.f. stage without input tuning, the feeder impedance forming a fair match with the emitter impedance. The collector circuit is resonated at the centre of the band by L1/C3, resistors R1, R2 and R3 setting the d.c. operating conditions so that approximately 2 mA is drawn from the 9 volt supply.

The degree of inductive coupling between L1 and L2 (tuned by C6) is a compromise between signal transfer, bandwidth and the rejection of second channel and intermediate frequencies. No neutralization or screening is necessary provided this coupling is not made too tight and chassis currents eliminated by the use of single point earthing of the various circuits. To this end, the base bypass, C2,

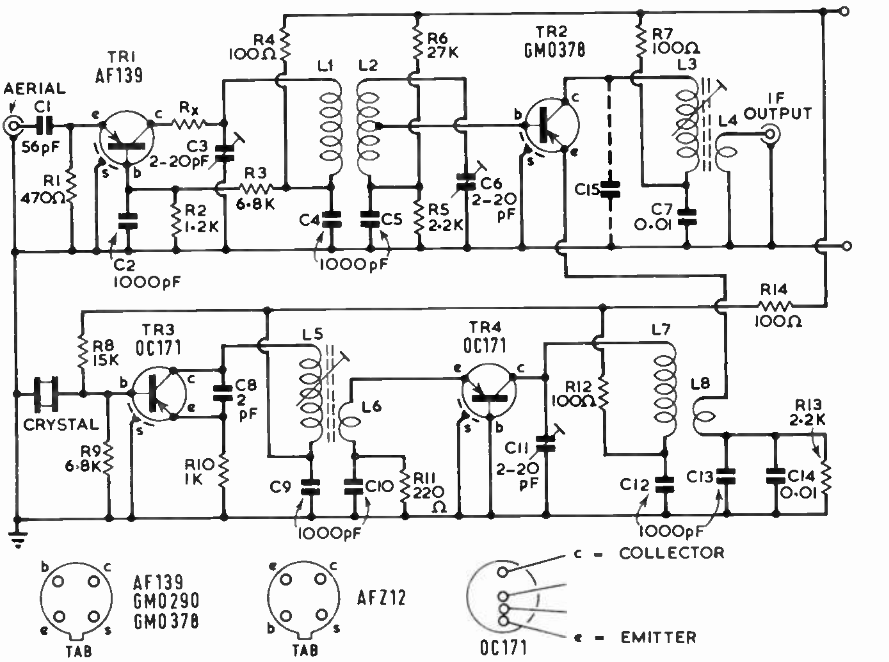


Fig. 5.25. Circuit diagram of the converter and transistor connections. For details of Rx see text. Crystal frequencies for other than the i.f. of 24 to 26 Mc/s described will be found in Table 5.3 together with values for C15 when this is required.

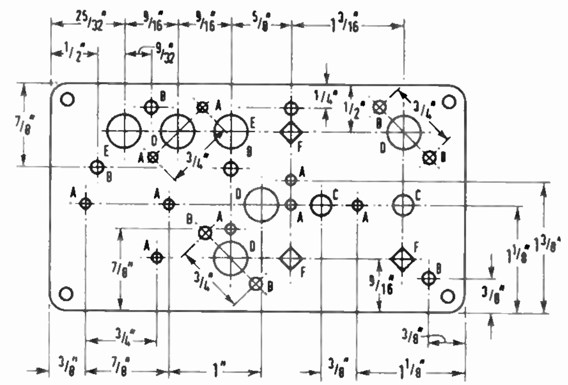


Fig. 5.26. Details for drilling the lid of the Eddystone die-cast box on which all components for the 144 Mc/s transistor converter are mounted.

is earthed to a tag under one of the bolts holding the coaxial aerial socket and the decoupling capacitor, C4, earthed as closely as possible to C3. Similar arrangements are made for C5 and C6. It will be noted that two bypass capacitors, C13 and C14, are connected in parallel across R13, the first having a low impedance to signal and oscillator frequencies and the second to the i.f. Both are earthed to the same tag as C5 and C6.

The mixer, TR2, is a type GMO378. The signal is fed to the base from a tapping on L2 and oscillator injection taken to the emitter via the coupling loop L8 from L7/C11 in the output of the crystal oscillator chain. R5, R6 and R13 control the d.c. operating conditions. Optimum performance requires a standing current of approximately 0·4 mA which rises to 0·6 mA when the injection is applied.

The inductance L3 is resonated by stray capacitance (approximately 5 pF) at the i.f. when this is in the range 24 to 26 Mc/s. Should it be desired to use an i.f. lower than 20 Mc/s, C15 may be required. Its connections are shown dotted on the circuit diagram. Suggested values are given in Table 5.3. The earthy ends of C7 and L4 (and C15 when used) are connected to a tag held by one of the bolts securing the i.f. output coaxial socket.

The oscillator chain is straightforward and with the component values given will work satisfactorily with overtone crystals in the range 30 to 47 Mc/s. An OC171 (TR3) serves as an overtone crystal oscillator in a circuit which has been found very reliable for crystals intended for this mode of operation on frequencies of this order. In this circuit configuration the transistor will oscillate only when a low impedance exists between base and earth. This condition is satisfied at the overtone frequency when the series impedance of the crystal falls to a low value. For an i.f. of 24 to 26 Mc/s the crystal overtone frequency is 40 Mc/s.

Another OC171 (TR4) acts as a tripler to 120 Mc/s. An alternative type of transistor would be the AFZ12. This stage is biased by emitter-base rectification, the emitter resistor R11 serving to limit the collector current to approximately 1 mA. The value of 220 ohms was arrived at as a compromise between efficiency and the danger of the transistor being damaged by excessive peak current.

TABLE 5.3

Circuit modifications for other intermediate frequencies and crystals

Tunable I.F.	Third Overtone Crystal *	Injection Frequency	Circuit Modifications
4·5–6·5 Mc/s	46·500 Mc/s	139·5 Mc/s	L7 wound ½ in. long with 38 s.w.g. enam. with 50 pF in parallel (C15).
9–11 Mc/s	45·000 Mc/s	135·0 Mc/s	C15 20 pF.
21–23 Mc/s	41·000 Mc/s	123·0 Mc/s	C15 5 pF.
27–29 Mc/s	39·000 Mc/s	117.0 Mc/s	None.
30–32 Mc/s	38·000 Mc/s	114·0 Mc/s	5 pF in parallel with L5.
33–35 Mc/s	37·000 Mc/s	111·0 Mc/s	5 pF in parallel with L5.

*Type HC6/U.

Construction

To prevent i.f. breakthrough the converter must be built in a metal case and an Eddystone die-cast box Type 845 measuring 4⅝ in. × 2⅜ in. × 1½ in. was used for the prototype. The components are all mounted on the lid of the box, drilling instructions being given in **Fig. 5.26**. The layout of the main components is shown in **Fig. 5.27**. Care must be taken to ensure that good contact exists between the earthy ends of the 2 to 20 pF trimming capacitors (C3, C6 and C11) and the case. In the prototype these were of the ceramic tube type made by Radiospares. An alternative would be the Erie type 535 which, however, requires slightly different sized fixing holes.

Adjustment

After checking the wiring a 9 volt battery may be connected, *taking care that the correct polarity is observed*. The total current should lie between 6 and 8 mA.

The oscillator should be adjusted first. As the core is inserted into L5/L6 the current increases to a maximum and then falls sharply: the core should be set just short of the maximum current position. It would, of course, be a great advantage when adjusting the crystal oscillator to be able to listen to the overtone frequency, but a receiver covering 40 Mc/s is something of a rarity. The oscillators in most receivers generate harmonics of sufficient amplitude however for a beat to be obtained on half the crystal frequency less half the i.f., i.e. if the required signal is on 40 Mc/s and the i.f. of the receiver is 465 kc/s, a beat should be obtained at approximately 20 Mc/s less 0·23 Mc/s or 19·77 Mc/s. It will probably be necessary in order to hear the beat to bring a lead from the aerial input of the receiver close to but *not* touching L5/L6.

The next step is to tune the circuits associated with TR4. As already mentioned the current passed by this stage should be approximately 1 mA and if necessary the value of R11 may be adjusted slightly to achieve this. C11 is tuned for a slight dip in the collector current towards the low capacitance end of its travel. After connecting the main receiver to the i.f. output socket the core of L3/L4 should be adjusted for maximum noise over the chosen i.f. range and a check made that the adjustment found for C11 also gives maximum noise. Removal of the crystal or coupling an absorption

TABLE 5.4

Inductor details for the 144 Mc/s Transistor Converter

L1	5 turns $\frac{5}{16}$ in. diam. $\frac{1}{2}$ in. long. 22 s.w.g. tinned copper wire.
L2	5 turns $\frac{5}{16}$ in. diam. $\frac{7}{16}$ in. long. 22 s.w.g. tinned copper tapped 1 turn from earthy end. L1/L2 spaced $\frac{7}{16}$ in. apart.
L3	50 turns on $\frac{3}{16}$ in. former $\frac{1}{2}$ in. long. 34 s.w.g. enam. Dust iron core.
L4	3 turns p.v.c. covered (7/40 s.w.g. "Radiospares") wound over L3.
L5	25 turns on $\frac{3}{16}$ in. former $\frac{1}{2}$ in. long. 26 s.w.g. enam. Dust iron core.
L6	2 turns p.v.c. covered (7/40 s.w.g. "Radiospares") wound over L5.
L7	5 turns $\frac{5}{16}$ in. diam. $\frac{1}{2}$ in. long. 22 s.w.g. tinned copper.
L8	1 turn p.v.c. covered (7/40 s.w.g. "Radiospares") interwound at earthy end of L7.

wavemeter close to L7 and tuned to its resonant frequency should both result in a reduction in noise output.

A strong local signal in the range 144 to 146 Mc/s is now required, failing which use may be made of a g.d.o., and C3 and C6 adjusted for maximum signal strength along with the core of L3/L4 and C11. Finally, using a weak signal, all the adjustments should be verified.

The use of a g.d.o., unless transistorized, is not recommended for setting coils in compact transistorized equipment. It is difficult to obtain a true reading of resonance in the small space available and all too easy to damage the transistors by the application of excessive r.f.

If a reliable noise source is available the coupling between L1 and L2 and the values of the biasing networks R1, R2, R3 and R5, R6, R13 may be adjusted for optimum noise figure, but the improvements obtainable are not likely to be great.

The performance of this converter will be found to be at least the equal of a good valve type using a 6CW4 r.f. stage.

Alternative Transistors

A GMO290 may be substituted for the AF139 in the r.f. stage with a marginal improvement in performance but it will be found necessary to include a resistor (R_X) of between 10 and 50 ohms between the collector and tuned circuit, as indicated on the circuit diagram, to overcome a slight tendency towards instability. This resistor was not found necessary when using the AF139.

Transistors which may also be employed in the mixer

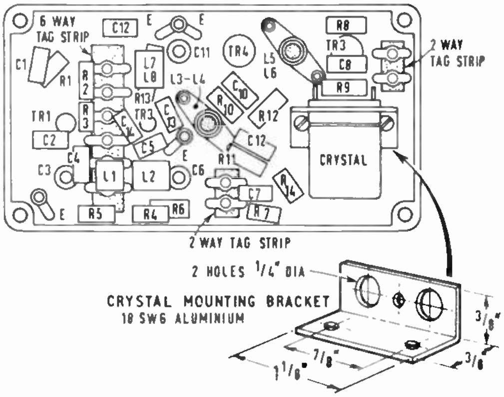
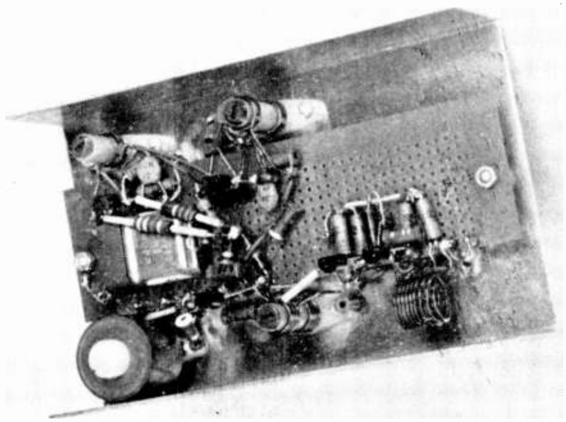


Fig. 5.27. Layout of the principal components on the underside of the lid of the box.



A 4m FET converter with a cascode amplifier.

stage include the AF139, AFZ12 and, with some degradation in performance, the AF114 or OC171. It should be noted that although the disposition of leads on the AFZ12 is similar to the GMO378, the base and emitter connections are interchanged.

Field Effect Transistor (FET) Converters for 70 and 144 Mc/s

The field effect transistor offers a performance rivalled only by the parametric amplifier. With the availability of such FETs as the Texas Instruments Ltd. 2N3819, it is practicable to design v.h.f. and u.h.f. receivers with extremely low noise levels and at the same time with an immunity to strong signal (cross-modulation) effects less than those shown by almost any valve converter, and far less than for r.f. and mixer stages using normal transistors.

Of the three FET converters described in this chapter, those for the 2 and 4m bands are almost identical and will, therefore, be treated together. The 70cm converter is somewhat different in design and is described later.

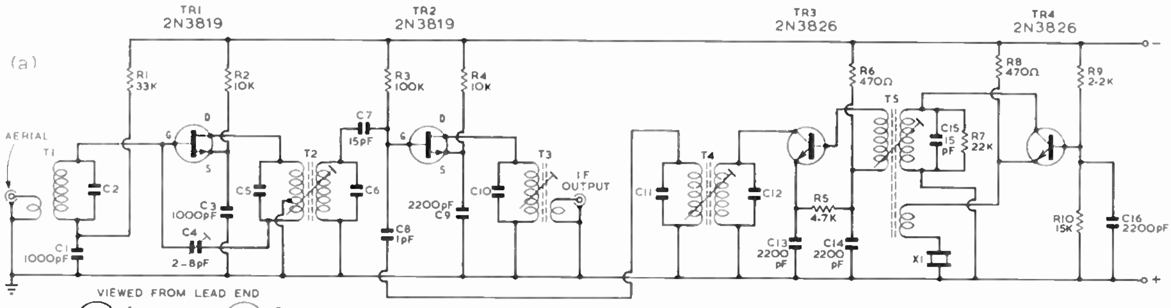
The elements in a FET have a different nomenclature from those in a normal transistor, and for ease of reference this is briefly mentioned.

FET	Normal Transistor	Valve (approx. only)
Source	Emitter	Cathode
Gate	Base	Grid
Drain	Collector	Anode

It should be noted that with the 2N3819, an *n*-channel device employed in these designs, the drain is connected to the positive side of the supply voltage while the gate is negatively biased and, as in a valve, draws negligible current.

The Two and Four Metre Converters

A common-source, neutralized r.f. stage feeds via double tuned circuits into a common source mixer (**Fig. 5.28(a)**). For those wishing to avoid even the simple procedure associated with neutralizing the r.f. stage, a cascode variant is shown **Fig. 5.28(b)** which does not appreciably alter the performance in any respect but does require another 2N3819. The local oscillator chain uses two cheap silicon *n-p-n* planar transistors, type 2N3826 (Texas Instruments Ltd.), one as an oscillator with an HC-6/U overtone crystal, operating in the 30 to



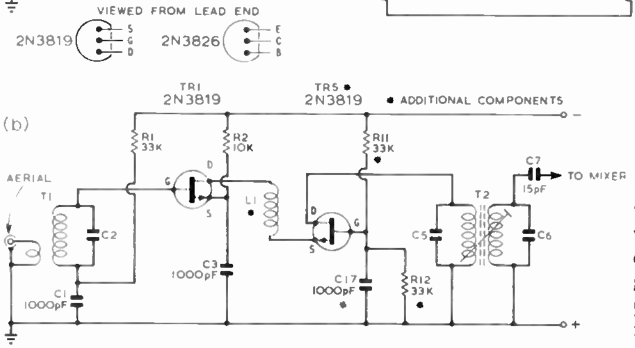


Fig. 5.28. (a) Circuit of the neutralised FET converter. (b) An alternative cascode input circuit. Operation is satisfactory for supply voltages between 10 and 14 volts.

35 Mc/s range, the other as a multiplier, doubling for 4m and quadrupling for 2m, with double tuned circuits feeding into the mixer.

The standard output i.f. for both converters is in the region of 3 Mc/s but design data is given for i.f.'s between 2 and 30 Mc/s for the 2m converter and between 2 and 10 Mc/s for the 4m. An Aladdin former, with enclosing Neosid pot core, is used to accommodate the output transformer for i.f.'s up to 16 Mc/s, but above this frequency, the Aladdin former only is required. To alter the i.f. it is only necessary to change the i.f. transformer coil and overtone crystal and perform a small amount of re-peaking.

Construction

The converters are constructed on a piece of Lektrokit pin board, using wiring pins pushed into the holes to support most components, including the crystal and the transistors (see **Fig. 5.29**). A standard, square piece of Lektrokit laminate board may be cut into three, each part of which may then be used to build a converter, although two fixing holes must be drilled appropriately in the centre of each piece. The board is mounted inside a simple, trough section, 20 s.w.g. aluminium chassis, using 6 B.A. screws and nuts, standing the board $\frac{5}{16}$ in. away from the chassis. The four coil formers, i.e. those associated with the oscillator and multiplier stages, together with the r.f.—mixer interstage coil and the i.f. transformer are all mounted on the chassis (see **Fig. 5.31**).

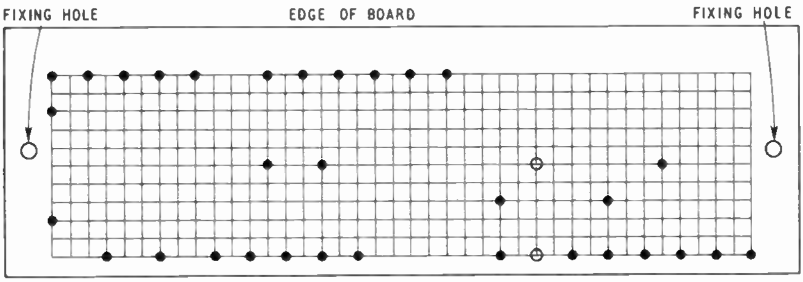
The aerial coil is air-spaced and is supported by wiring pins on the board. As the converters illustrated were intended for incorporation into a complete v.h.f. receiver, the aerial and i.f. terminations were, like the ground and input negative connections, made to wiring pins but, if required, two Belling-Lee co-axial sockets may be mounted on the chassis, after drilling the necessary holes. No extra shielding is required anywhere, the need for it being avoided by proper layout and coil positioning.

The logical steps for construction are:

(a) Cut and mark out the chassis.
(b) Bend up the chassis.
(c) Drill all holes.
(d) Fit the earthing tags and two Belling-Lee sockets, if required.
(e) Push the pins into the Lektrokit wiring board.
(f) Connect the leads on the underside of the board.
(g) Mount the Lektrokit board in the chassis.
(h) Mount the crystal and its holder on the board.
(i) Manufacture all the coils and fix the Aladdin formers to the chassis.
(j) Attach all the other components, input coil, etc. and wire up.

The metal work for the converters hardly needs any special comment as **Fig. 5.30** is self-explanatory.

The wiring pins should be pushed into the Lektrokit board from the underside so that about twice as much pro-



• WIRING PINS REQUIRED FOR BOTH 2 METRE AND 4 METRE CONVERTERS — 32 OFF
O EXTRA WIRING PINS FOR CASCODE MODIFICATION — 2 OFF

Fig. 5.29. Wiring pin arrangement.

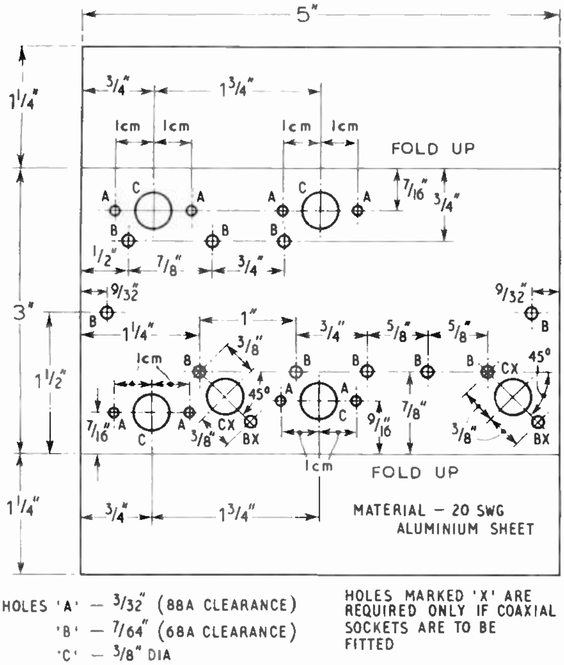


Fig. 5.30. Chassis drilling plan.

trudes above the board as is left below. The best type of pin to use is that with a shoulder and longitudinal splines (Vero Electronics Type 2143). Otherwise, Lektrokit LK3011 pins may be used. The finished converter looks neater if some wiring is done underneath the board, prior to mounting it in the chassis. In particular, the battery wiring and the earth wire from the crystal clip may be so inserted. If the cascode r.f. stage is to be used, the earth connection to the bottom of R2 may also be made under the board with advantage.

Mounting the Lektrokit board in the chassis requires the use of two 6 B.A. × ½ in. ch. hd. screws and nuts, spacing the board $\frac{5}{16}$ in. away from the chassis. The crystal holder is fixed to the board by soldering the connections to two wiring pins. The crystal is inserted and is then clamped in position by bending a 6 B.A. tag carefully over it, after the tag has been soldered to the earthed wiring pin provided.

The coils for T2 and T4 are produced by close-winding 22 s.w.g. enamelled wire on the shank of a ¼ in. drill (wind one more turn than required). If the wire is then pulled off carefully, it may be sprung, by rotation, on to the Aladdin former. Note that the earthed end of the secondary of T2 is nearer to the chassis. The primary is situated at the outer end of the former, with its " cold " end nearer to the secondary. The two coils are spaced by $\frac{3}{16}$ in. The earthed end of the primary of T4 is close to the chassis and the earthed end of the secondary (which is situated right at the other end of the former) is nearer to the primary. The cores for both T2 and T4 are prepared by cutting a standard core in half; they are locked in position, inside the former, by means of narrow strips of polythene sheet.

The main winding of the oscillator coil T5 is close-wound near the chassis end of the Aladdin former, its earthed end being close to the chassis. A layer of adhesive Melinex tape facilitates putting the other windings on top, the feedback coil at the cold end and the output winding near the centre of the main winding.

The i.f. transformer T3 should have each half of its tuned winding wound in two adjacent coil former sections, leaving one section available for the output-coupling secondary.

TABLE 5.5

Coil Details

Four Metres
L1, 30 turns, 36 s.w.g. enam., close wound on 100 K ohms ¼ watt resistor (required for cascode only).
T1, secondary: 8 turns, 22 s.w.g. enam., self-supporting (wound on ⅜ in. mandrel); Primary: 3 turns, 22 s.w.g. enam., pushed into " cold " end of secondary.
T2, Primary: 6 turns, tapped 2 turns from earthy end, 22 s.w.g. enam., close wound on 0·3 in. Aladdin former; Secondary: 6 turns, 22 s.w.g. enam., close wound nearer chassis, spaced $\frac{3}{16}$ in. from primary.
T3, see Table 5.6.
T4, Primary: 4 turns, 22 s.w.g. enam., close wound near chassis on 0·3 in. Aladdin former; Secondary: 4 turns, 22 s.w.g. enam., close wound, spaced ¼ in. from primary.
T5, Main: 11 turns, 28 s.w.g. enam., close wound on 0·3 in. Aladdin former; Emitter: 1 turn, 28 s.w.g. enam., overwound at " cold " end of main winding; Output: 2 turns, 28 s.w.g. enam., overwound at centre of main winding.
Two Metres
L1, 10 turns, 28 s.w.g. enam., close wound on 100 K ohms ¼ watt resistor (required for cascode only).
T1, Secondary: 4 turns, 22 s.w.g. enam., self-supporting (wound on ⅜ in. mandrel); Primary: 2 turns, 22 s.w.g. enam., pushed into " cold " end of secondary.
T2, Primary: 5 turns, tapped 2 turns from earthy end, 22 s.w.g. enam., close wound on 0·3 in. Aladdin former (ignore tap for cascode); Secondary: 5 turns, 22 s.w.g. enam., close wound nearer chassis, spaced $\frac{3}{16}$ in. from primary.
T3, See Table 5.6.
T4, 5, as for 4m.

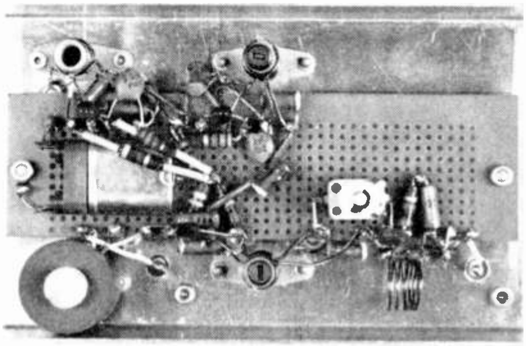
TABLE 5.6

Data for I.F. Transformer T3

Tunable I.F.	I.F. centre	Crystal freq.	T3 primary inductance	Turns	S.W.G.	Former	75 ohm coupling turns	600 ohm coupling turns
Four Metres								
1·8–2·4	2·1 Mc/s	34·15	205 μH	66	36	Pot-core	5	13
2·1–2·7	2·4	34·00	158	58	36	Pot-core	4	11
4·1–4·7	4·40	33·00	46·8	31	32	Pot-core	2	6
7·1–7·7	7·40	31·50	16·5	19	28	Pot-core	1	3
9·1–9·7	9·40	30·50	10·3	15	28	Pot-core	1	3
Two Metres								
1·8–3·8	2·8 Mc/s	35·55	369 μH	88	36	Pot-core	6	17
2·0–4·0	3·0	35·50	316	82	36	Pot-core	6	16
4·0–6·0	5·0	35·00	105	47	34	Pot-core	3	9
8·0–10·0	9·0	34·00	31·6	26	32	Pot-core	2	5
14·0–16·0	15·0	32·50	11·3	15	28	Pot-core	1	3
20·0–22·0	21·0	31·00	5·8	29 c/w	26	Aladdin former	2	5
24·0–26·0	25·0	30·00	4·1	27 c/w	26	Aladdin former	2	5
28·0–30·0	29·0	38·67	3·0	25 c/w	26	Aladdin former	2	5

Lining-up and Operation

The output i.f. tuned circuit will not normally require peaking for i.f.'s below 10 Mc/s. The crystal oscillator will only oscillate strongly on the correct overtone frequency and probably will not oscillate at all unlocked unless the crystal has an exceptionally high shunt capacitance. If a receiver in the 35 Mc/s region is not available, correct operation of the



A 2m FET converter using the neutralized input stage.

oscillator may be checked by the method described on page 520 in connection with the adjustment of the 2m transistorized converter. If the oscillator does not appear to work at all, the sense of the feedback winding should be reversed.

An r.f. detector loop and diode and a set of calibrated absorption wavemeters will be found useful when tuning up the multiplier coils. However, if these are not available, the 1 pF oscillator injection capacitor lead may be temporarily moved to the multiplier collector and the multiplier coil tuned to give a noise peak. Then the injection capacitor should be reconnected to the secondary winding of the multiplier coil which may itself be aligned, and the primary re-peaked.

It is next necessary to resonate the mixer gate circuit. In the case of i.f.'s below 6 Mc/s with the 2m converter and below 3 Mc/s with the 4m converter, it may be difficult to distinguish between the noise peaks due to operation above and below the frequency of the local oscillator. A very pronounced dip in noise between noise peaks probably indicates too much oscillator injection. This may be checked by measuring the mixer source current (connect a 0–10 voltmeter across R4). This current should not increase by even a few per cent when the local oscillator is switched on. If it does, reduce the coupling between the two parts of T4. Operation of the mixer at too high a level of oscillator injection will produce a very poor strong signal performance.

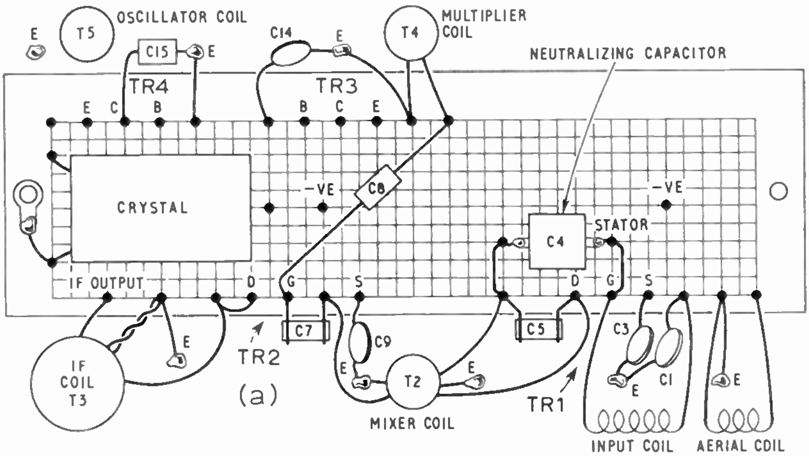
With a 75 ohm load resistor (or an aerial) plugged into the input socket, it should be found possible to tune the primary of T2 to resonance at signal frequency. At the same time (in the case of the single, earthed-source r.f. stage), the neutralizing capacitor may be adjusted for best stability. The input circuit may then be set up in the usual manner for valve r.f. amplifiers and the entire converter finally peaked using a signal generator or external signals.

Performance

One of the advantages of FET's over ordinary transistors, or even valves, is their ability to operate satisfactorily in the presence of strong signals, of the order of 100 mV or so, from other (unwanted) stations working in the same band. This is due to the curvature of the mutual characteristic of the FET being such that cross-modulation does not readily occur; at the same time the characteristics are almost ideal for the role of mixer. The absence of a heated, emissive cathode, apart from other considerations, is conducive to a low noise level. Both the converters are capable of providing a noise factor better than 3db provided that the noise factor of the main receiver (variable i.f. tuner) into which it is fed is not itself worse than about 10db. This latter proviso is a result of the converter gain having been held to a low value to minimize overall strong-signal effects. Should constructors wish to use tuners with rather poor noise performance the necessity of an extra amplifying stage may be avoided by simply reducing the values of the 10 K ohms r.f. stage and mixer source feed resistors, thus increasing the gain of the converter. The minimum advisable resistances are 500 ohms for the r.f. stage and 1 K ohm for the mixer.

To achieve gain equal to that of the original converters, but with the use of a 9 volt battery, the following component modifications should be made:

Change R2 to 6·8 K ohms
R4 to 6·8 K ohms
R10 to 10 K ohms.



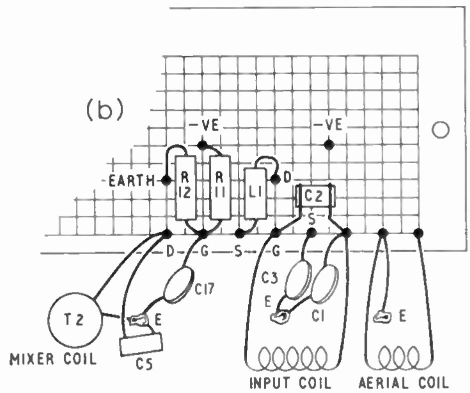


Fig. 5.31 (a) Wiring diagram for the two-FET converter. (b) Modifications for incorporating a cascode r.f. stage.

TABLE 5.7

Components List

C1, 3, 1000 pF disc ceramic.
C2, 4·7 pF ceramicon (4m), not required for 2m.
C4, 2–8 pF miniature trimmer, Plessey type 7365 (not required for cascode).
C5, 10 pF ceramicon (4m), 2·2 pF ceramicon (2m).
C6, 10 pF ceramicon (4m), not required for 2m.
C7, 15 pF ceramicon.
C8, 1 pF ceramicon.
C9, 13, 14, 16, 2200 pF disc ceramic.
C10, 22 pF ceramicon (4m), 6·8 pF ceramicon (2m).
C11, 12, 33 pF ceramicon (4m), 6·8 pF ceramicon (2m).
C15, 15 pF ceramicon.
R1, 33 K ohms.
R2, 4, 10 K ohms (10–14 volts), 6·8 K ohms (9 volts).
R3, 100 K ohms.
R5, 4·7 K ohms.
R6, 8, 470 ohms.
R7, 22 K ohms.
R9, 2·2 K ohms.
R10, 15 K ohms (10–14 volts), 10 K ohms (9 volts).
T2, 4, 5, 0·3 in. diam. nylon Aladdin former, with iron dust core.
T3, 0·3 in. diam. nylon Aladdin former, with iron dust core, for i.f.s above 16 Mc/s. Neosid pot core on Aladdin former for i.f.s below 16 Mc/s.
TR1, 2, 2N3819 FET (Texas Instruments Ltd.).
TR3, 4, 2N3826 (Texas Instruments Ltd.).
X1, HC-6/U crystal and holder, as required.
The following additional components are required for the cascode modifications:
C17, 1000 pF disc ceramic
R11, 12, 33 K ohms.
L1, 100 K ohms (see Table 5.5).
TR5, 2N3819 (Texas Instruments Ltd.).

A TRANSISTOR MAST-HEAD AMPLIFIER FOR 144 MC/S

Siting a v.h.f. aerial in the best position for transmission and reception often means the use of a long coaxial cable with consequent loss of the weaker incoming signals due to feeder attenuation. The use of a pre-amplifier mounted on the array itself goes a long way towards improving reception, and if a transistor pre-amplifier is used the difficulties of supplying high and low tension voltage to the unit are avoided.

The following design, giving about 8db of gain, was originated by L. W. Aurick, K3QAX/W2QEX and appeared in the Fall 1964 issue of *Ham Tips:* it is published here by kind permission of the copyright holder, the Radio Corporation of America.

By reference to **Fig. 5.32** it will be seen that, in the receive position, signals are applied through C1, the gamma-matching capacitor, via relay contacts *X* and *Y* to the base of a type 2N2708 transistor. L1, in parallel with C2, in the collector circuit is tuned to the centre of the band and the amplified signals pass, via C3 and relay contacts *A* and *B* to the inner conductor of a coaxial cable to the aerial change-over switch in the station and thence to the receiver. On transmission the relay is energized in the pre-amplifier, disconnecting the latter, and connecting the feeder directly to the aerial.

It will be seen that R1/R2 forms the base biasing potentiometer across the supply, R3 provides emitter bias and radio frequency chokes RFC1 and RFC2 are inserted to prevent a loss of signal through the biasing resistors. A 20 volt supply, with the positive side earthed as the 2N2708 in a *n-p-n* type transistor, is connected by one pair of leads in a three-core cable to terminals 1 and 2, the remaining lead being run between terminal 3 and one side of the pair of additional contacts on the aerial changeover switch previously mentioned.

In the original design the relay is a relatively inexpensive 12 volt open type, normally employed for aerial changeover service at the lower frequencies and suitable for transmitter inputs up to 100 watts. If the more expensive coaxial type were substituted power could be increased subject to the transistor not being overloaded by stray r.f.

The pre-amplifier is constructed on an aluminium plate 1½ in. square having a ½ in. lip for mounting on the wall of a metal box 4 in. × 2¼ in. × 2¼ in. All wiring should be as short as possible and the relay should be fixed to the opposite short wall of the box. The unit may be mounted on the aerial boom in any convenient manner, perhaps the easiest method being by means of a U-bolt fastened to the top of the box. A braided copper strap should be connected between a lug on the nearest wall of the box and a point on the boom directly in line with the gamma-match. As the earth strap could be a source of inductive reactance that must be cancelled by C1, it should be kept as short as possible. Capacitor C1 is connected to the gamma-match bar through a ceramic feed-through mounted on the box close to the relay. C1 must be isolated from the box by means of a small piece of polystyrene sheet and a hole provided so that adjustment may be effected by an insulated screwdriver. A cork or rubber bung should be inserted in the hole after adjustment to keep out damp. Resistor R4 should be of such a value that the relay operates satisfactorily from the 20 volt supply.

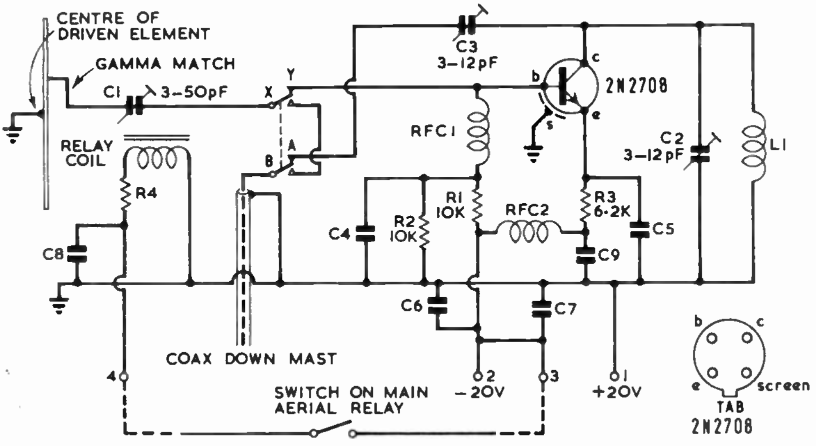


Fig. 5.32. Circuit diagram of a transistor mast head amplifier for the 2m band. C1 must be capable of carrying the transmitter aerial current. C4, 5, 9 470pF disc ceramic. C6, 7, 8 500pF feedthroughs. R4 must be of suitable rating and value to operate the relay from the 20 volt supply. Other resistors are of ¼ watt rating.

Setting-up—Transmission

With the relay in the TRANSMIT position, the length of the gamma matching bar and the value of C1 are adjusted with the aid of a standing wave ratio bridge, until the s.w.r. is as near 1 to 1 as possible.

Setting-up—Reception

Having ascertained, by means of a grid dip meter, that the circuit L1 C2 will tune to the centre of the band, an appropriate test signal is located on the receiver and C2 and C3 adjusted alternatively until the strongest result is obtained. This will be with C3 near maximum and C2 near minimum capacity.

U.H.F. RECEIVERS

The 70cm band involves an increase in frequency of three times that used for the 2m band and the 23cm band a further increase of three times. Since the losses in dielectrics, the r.f. skin effects in conductors and valve transit time, losses increase rapidly as the frequency is raised, particular care must be exercised in the choice of components, materials and valves for use on the u.h.f. bands if high performance is to be achieved.

Photograph of a B9A valve-holder with an earthing collar used to earth the grounded electrodes of a valve. Slots are provided adjacent to tags connected to the external circuit. Leadthrough bypass capacitors are mounted close to the socket.

Metal parts which form any part of, or are within the field of, a tuned circuit or carry r.f. currents, should be of copper or brass and not of steel or aluminium although these materials may be used if they are silver plated to an adequate thickness for the frequency in use. Where facilities are available, it is best to plate all such parts. They should, however, not be polished but left matt and protected from tarnish by lacquering. Insulating materials used in or near tuned circuits or in the manufacture of valveholders should be of p.t.f.e. (fluon), polystyrene or polythene. Such materials as Perspex, ceramic, mica or p.v.c. should be employed only with caution; Bakelite and nylon should not be used.

A number of valve types using glass bases require several of the pins to be effectively earthed to achieve stability; short lengths of wire to adjacent tags on the chassis are not sufficient for this purpose. One convenient method is to mount an earthing collar around the valveholder and to bolt it firmly or solder it to the chassis; the appropriate valveholder tags can then be soldered to the collar, as shown in the photograph. Where valve electrodes require decoupling to earth, feedthrough capacitors should be used. These should be mounted as closely as practicable to the valveholder tag. Those capacitors built into cavities and similar constructions should be made of p.t.f.e. or polythene film clamped between metal surfaces.

TUNED CIRCUITS

As the resonant frequency of a tuned circuit is raised the capacitance and inductance must be reduced until a point is reached at which the necessary inductance becomes so small that it is impracticable to use the ordinary type of coil construction and it assumes the form of a hairpin loop. **Fig. 5.33** illustrates how progressively high frequencies are attained. **Fig. 5.33 (b)** is frequently employed on 70cm while **Fig. 5.33 (c)** is similar to a butterfly circuit where the capacitor is of the split-stator type. **Fig. 5.33 (d)** represents a multiplication of this structure and in **Fig. 5.33 (e)** there is in effect an infinite number of loops in parallel, i.e. a cylinder closed at both ends with a central rod and known as a *coaxial cavity*. The capacitance element is provided by a threaded rod T, which passes through a threaded end-plate; a disc D1 is fixed to the lower end of the rod so that it is parallel to a similar disc D2 attached to a central support. The simple hairpin, shown at Fig. 5.33 (b), is a very convenient form of construction: it can be made of wide strip rather than wire and is especially suitable for push-pull circuits. It may be tuned by parallel capacitance at the open end P or by series capacitance inserted at the closed end S. In a modification of the hairpin loop, one side can consist of a straight wire or strip while the other side is a sheet of metal which can be part of the chassis, the strip being spaced from the metal; such an arrangement is known as a *strip line*. When very close to the metal sheet but insulated from it by a solid dielectric film, it is known as a *microstrip*.

All of these resonant elements can be either series or parallel tuned. Tuned circuits in the form of hairpin loops or strip lines can be used quite successfully up to about 700 Mc/s; above this frequency butterfly circuits or coaxial cavities must be employed. Because a cylindrical cavity usually requires the use of a lathe for its construction, a simpler type in the form of a trough which can be made with hand tools is frequently used: see **Fig. 5.34**. The open side may be closed with a well-fitting lid if required: the omission of the lid only reduces the Q slightly. One end of the trough is closed by a plate to which the inner line is secured and the

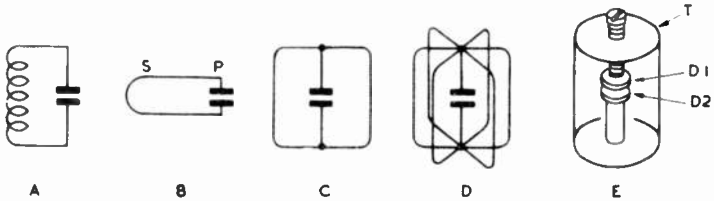


Fig. 5.33. Progressive development of tuned circuits from a coil to a cavity as the frequency is increased.

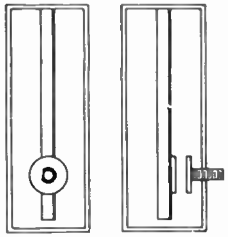


Fig. 5.34. A trough line tuned circuit.

three edges of the plate must make a sound low resistance connection with the trough, preferably by soldering. Tuning may be effected by a disc type capacitor or a small trimmer connected between it and the end of the inner line.

A cavity or trough line must have dimensions for the frequency used and the stray capacitance associated with the valve or crystal connected to it. The dimensions can be calculated from the basic equation:

$$l = \frac{\lambda}{2\pi} \tan^{-1} \frac{1}{2\pi f Z_0 C}$$

where l is the length in the same units as wavelength, f is the frequency in cycles per second, C is the capacitance (total) in Farads, Z_0 is the characteristic impedance in ohms.

The formula may be more conveniently rewritten as:

$$l = \frac{30{,}000}{2\pi f} \tan^{-1} \frac{10^6}{2\pi f Z_0 C}$$

where l is in centimetres, f is in Megacycles per second, C is in picofarads and Z_0 is in ohms.

The characteristic impedance Z_0 for a coaxial line is obtained from the formula $Z_0 = 138 \log_{10} \frac{b}{a}$ where b is the inside diameter of the outer tube and a is the outside diameter of the inner tube. For a square trough line the formula is:

$$Z_0 = 138 \log_{10} p + 6{\cdot}48 - 2{\cdot}34A - 0{\cdot}48B - 0{\cdot}12C$$

$$\text{where } p = \frac{\text{the inside dimension of the trough}}{\text{the outside diameter of the inner line}}$$

$$A = \frac{1 + 0{\cdot}405p^{-4}}{1 - 0{\cdot}405p^{-4}}$$

$$B = \frac{1 + 0{\cdot}163p^{-8}}{1 - 0{\cdot}163p^{-8}}$$

$$C = \frac{1 + 0{\cdot}067p^{-12}}{1 - 0{\cdot}067p^{-12}}$$

It can be seen from the formula for evaluating l that Z_0C occurs only in one portion of the equation. Quite clearly an infinite number of values of Z_0 and C will give the same multiple. Thus a trough line or cavity with a characteristic impedance Z_0 of 100 ohms and a capacitance of 8 pF will be the same length as one for which Z_0 is 200 ohms and C is 4 pF if they are both tuned to the same frequency. From this it follows that a graph can be plotted of frequency versus length of inner line for various values of Z_0C, as shown in Fig. 5.35. With its aid a trough line or cavity can be designed or the resonant frequency of an unknown cavity obtained.

The choice of the ratio of outer to inner dimensions is governed by electrical and mechanical considerations and the purpose. The value of the capacitance C is given by the valve capacitance, which may be appreciable, together with any tuning capacitance that may be present. It may prove to be so large that the length required for a high impedance cavity becomes so short as to be unmanageable: in extreme cases it might even be shorter than the valve which is to be mounted inside it.

Cavities are most readily constructed of lengths of standard sized brass or copper tubing and the choice of the ratio of outer to inner dimensions will of course be limited by the range of sizes available. For oscillator or amplifier anode tank circuits, cavities should in general have a low impedance (of the order of 30–40 ohms) whereas for input circuits the impedance may be as much as 100 ohms or higher. Optimum Q for an unloaded cavity occurs when the characteristic impedance Z_0 is 77 ohms. If it appears necessary either for achieving a proper impedance match or for reducing the capacitance loading, the connection from the valve or crystal can be tapped down to a " lower " point on the inner conductor (i.e. nearer the zero potential end). When this is done the value of C is calculated by dividing the valve or crystal capacitance in Farads by the total length of the line in inches, divided by the length from the short circuited end to the tapping point. The tapping point should not, however, be taken too far down the inner conductor otherwise the cavity will probably operate in some other mode than was intended and the resonant frequency will then be quite different from that expected.

Because of the variation in capacitance between various samples of the same valve or crystal (usually of the order of 30 per cent) provision should be made for varying the electrical length of a circuit if replacements are likely to be made. This can be achieved by sliding the line or cavity in and out, thus changing the effective length, but since a very high mechanical precision is essential and the sliding surfaces must be plated with non-tarnishable material such as rhodium, it is not recommended for amateur purposes.

A more convenient method is to use a small trimmer capacitance formed of two discs, one mounted on the inner conductor and the other carried on a screwed stem mounted in a threaded bush fitted in the outer conductor. Some means of locking or gripping the thread of the adjustable element is desirable to ensure not only a sound contact but also to prevent any wobbling of the disc as it is adjusted.

A graph relating the diameters and capacitance ranges of disc type capacitors is given in Fig. 7.7 in Chapter 7 (*V.H.F./U.H.F. Transmitters*).

Input and Output Coupling

Three methods may be used to couple power into or out of a resonant cavity or line: (a) a tapping may be made on to the inner line near the closed end (**Fig. 5.36(a)**); (b) a coupling loop may be inserted near the inner line towards the closed end (**Fig. 5.36(b)**); (c) a capacity probe may be inserted near the open end of the inner line (**Fig. 5.36(c)**).

The first method allows no easy means of adjustment and for low impedances the tapping point is often inconveniently near the closed end. The second is more versatile since the

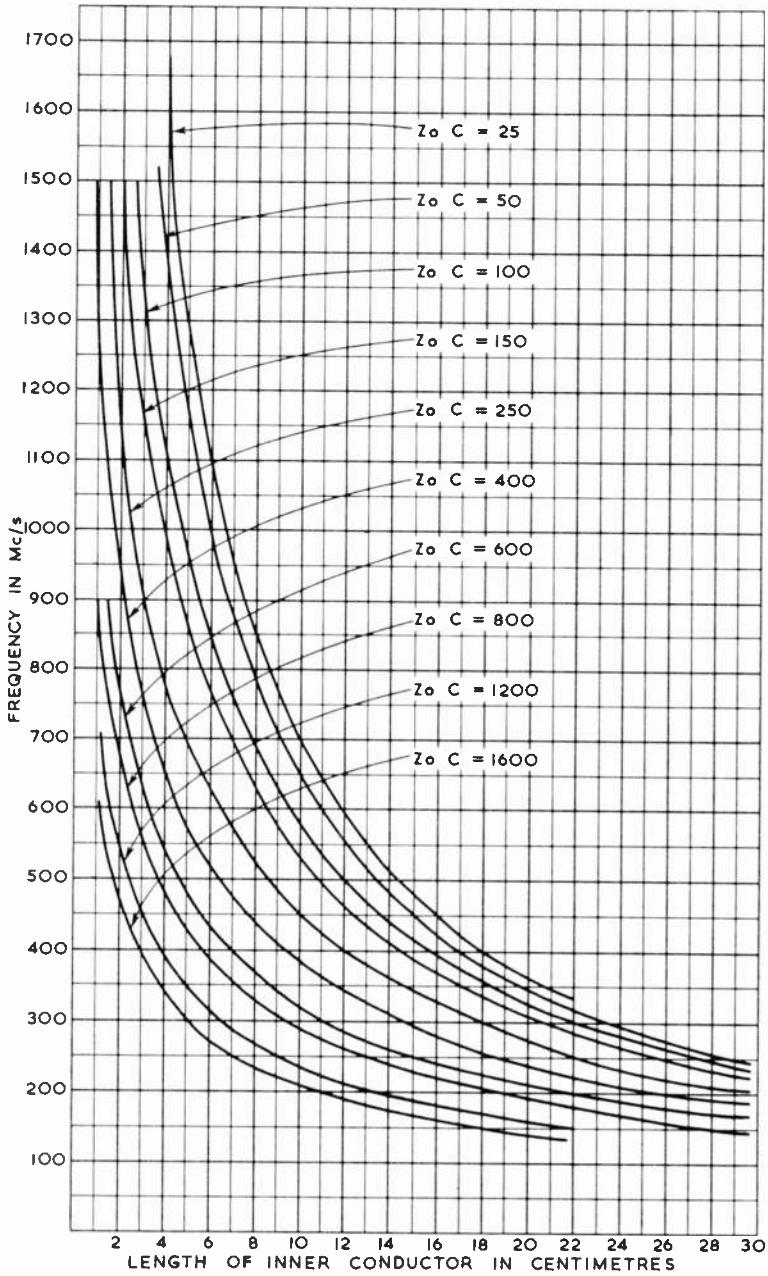


Fig. 5.35. Chart plotting frequency against length of inner line for various values of the characteristic impedance multiplied by the total capacitance.

degree of coupling can be readily adjusted either mechanically by rotating the loop or varying its size, or electrically by tuning the loop with a series or parallel tuning capacitor: when very tight coupling is required the loop should be boot shaped as shown in Fig. 5.36 (b). The third method is generally suitable only for high impedance coupling, the position of the probe varying the coupling, but if required it can be matched to a low impedance circuit, for instance, by inserting a quarter-wave line: details of this technique will be found in Chapter 14 (*V.H.F. Aerials*).

THE MIXER

Two basic types of mixer are used in u.h.f. receivers—triodes (generally connected in a grounded grid circuit) or crystal diodes. In some cases a thermionic diode may be used instead of the crystal type but the performance may be inferior because its capacitance is higher. For the 70cm band either type may be used, but triode valves are more often used than crystals because they are more readily available and have a conversion gain, whereas crystals and thermionic diodes have a conversion loss. For the 23cm band and below (i.e. frequencies above 1250 Mc/s) the mixer is almost invariably a crystal.

Triode Mixers

A triode valve can be used as a mixer in two essentially different ways. The non-linear characteristic necessary for mixing results from biasing the valve to operate either on the anode bend or on the grid current bend (leaky grid). In the anode bend method the application of the local oscillator voltage will drive the anode current upwards; in the leaky grid method it will drive the anode current downwards. Provided that the recommended oscillator injection voltage applied is correct the resultant anode current will be the same by either method and the conversion gain and the noise factor will be the same. For anode

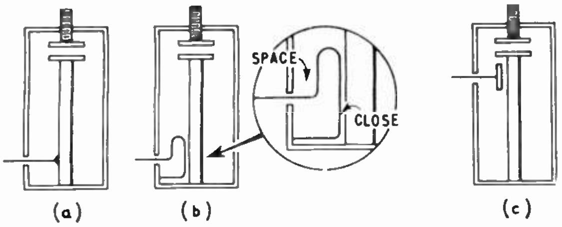


Fig. 5.36. Three methods of coupling into a line or cavity. (a) Tapping the inner conductor. (b) Boot-shaped coupling loop at low impedance end of inner conductor. (c) Probe at high impedance end of inner conductor.

bend operation, valves having indirectly heated cathodes can be provided with automatic bias. Those u.h.f. valves with a common heater and cathode connection, however, will need a separate heater winding. When disc-seal valves mounted in cavities are used the grid disc or ring will have to be insulated from the cavity if they are to be used as leaky grid mixers.

The choice of method will also depend on whether the valve is connected in a grounded grid or grounded cathode circuit. If trouble is experienced with i.f. breakthrough the mixer input electrode (the grid or the cathode as the case may be) should be bypassed to earth with a reactance that is low for the i.f. employed. A typical 70cm mixer is shown as part of Fig. 5.41 later in this chapter.

The advantages of a triode mixer are the useful power gain that it provides and the reduction in the number of i.f. stages resulting from the fact that it acts as an i.f. amplifier itself. The exact value of oscillator injection is not very critical and a valve is not so readily damaged by accidental pick up of r.f. power from the transmitter as a crystal, but it is of course more expensive and the alternative of including an additional i.f. stage may be considered better value. In a well-designed arrangement the noise factor is of the same order.

Crystal Mixers

As a crystal mixer acts as a diode and operates non-linearly around the bottom bend in the characteristic, the oscillator injection voltage applied must be sufficient to swing the operating point well round the bend. If the voltage is too low the conversion efficiency will be low because the curvature of the working portion of the characteristic will be relatively slight. On the other hand, if the voltage is too high the conversion efficiency is no greater but more noise is produced because the noise output is proportional to the oscillator voltage input. It is clear therefore that the injection voltage employed is fairly critical if optimum noise factor is to be achieved and some means of adjusting its value should be provided in the receiver. Preferably, a meter should be installed permanently in circuit so that the oscillator voltage can be set for the best performance. The meter should be a microammeter connected in series with the crystal, the applied voltage being varied to produce a predetermined current which will vary according to the type of crystal: even different specimens of the same type will produce different current levels. The current is likely to be in the range 0·1–1 mA.

Some crystals require a small negative bias for optimum performance and a suitable bias voltage can be provided by including a resistor of 500–1000 ohms in the circuit, but it should be borne in mind that some microammeters have a resistance of this order and some may even have too high a resistance. The resistor and meter must be adequately by-passed for both r.f. and i.f. currents. A number of crystals suitable for 70cm are available, examples being the GEX66 (CV2290) or USA CK710, and for 23cm the type CS2A (similar to the CV103) or 1N21 or SIM2.

Fig. 5.37 shows a typical mixer cavity for 23cm operation.

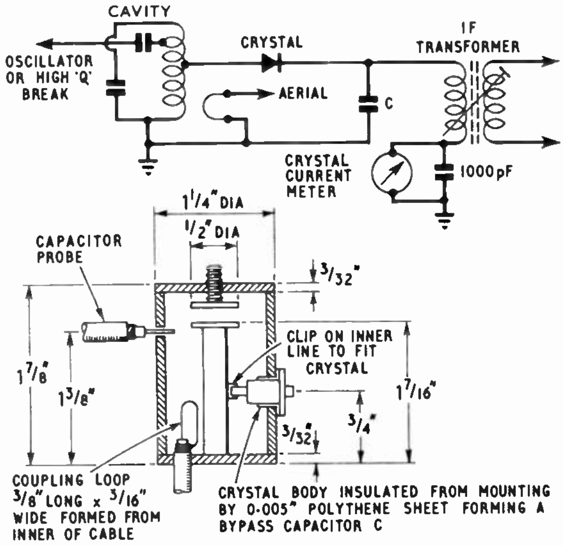
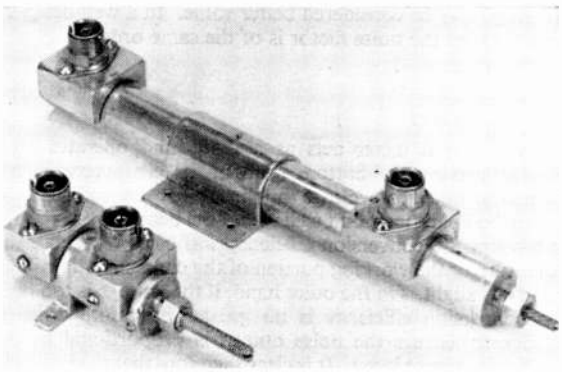


Fig. 5.37. Typical mixer cavity and circuit for 1296 Mc/s.

High Q Break

The local oscillator voltage in a receiver may be generated by a single valve or, as mentioned in the v.h.f. receivers section, at a lower frequency to obtain better stability, and the frequency multiplied by one or more doublers or treblers. The output of an oscillator contains frequencies other than the required one (i.e., harmonics in addition to the fundamental) together with some noise; the output of a chain of multiplier stages includes lower frequencies in addition to the required frequency. Ideally the local oscillator voltage injected into the mixer should be a pure sine wave of the required frequency; other frequencies merely generate more noise in the mixer without contributing at all to the signal. Since a low noise level is an essential requirement in u.h.f. receivers, the voltage injected into the mixer should be at the fundamental only and all other frequencies should be removed by a filter. Such a filter is known as a *high Q break* and usually consists of a coaxial cavity with a coupling loop at the low potential (" cold ") end and a probe coupling at the high potential (" hot ") end, both being fairly loosely coupled so that a high Q value is achieved. Examples of 70cm and 23cm high Q breaks are shown overleaf. Typical performance of a 70cm filter would be an attenuation of −15db at a frequency 3·5 Mc/s away from the operating



Two high Q Breaks, the longer for 430 Mc/s, the shorter for 1296 Mc/s.

frequency. This would ensure that, provided that the original source is higher than 3·5 Mc/s, no other frequency of sufficient magnitude to matter would be fed into the mixer. Where the last valve in a frequency multiplied oscillator chain is of the disc-seal or coaxial type with a tuned cavity anode circuit, this cavity can readily be made sufficiently selective to operate as a high Q break.

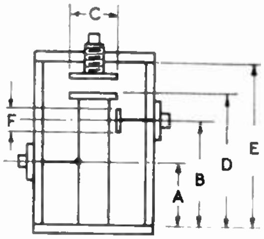


Fig. 5.38. Essential dimensions for high Q breaks.

	23 cm	30 cm	70 cm
Inside diameter of outer conductor	43/64 in.	43/64 in.	43/64 in. *
Outside diameter of inner conductor	1/4 in.	1/4 in.	1/4 in.
Dimension A (end to tap on inner)	5/16 in.	5/16 in.	3/8 in.
Dimension B (end to coupling probe)	1 1/4 in.	1 3/4 in.	5 3/4 in.
Dimension C (diam. of trimmer)	3/8 in.	3/8 in.	3/8 in.
Dimension D (length of inner conductor)	1 1/8 in.	2 in.	6 in.
Dimension E (length of outer conductor	1 3/4 in.	2 3/8 in.	6 1/2 in.
Dimension F (diam. of probe)	1/4 in.	1/4 in.	1/4 in.

*3/4 in. o.d. tube with 20 s.w.g. wall.

THE INTERMEDIATE FREQUENCY

The same basic considerations apply to the choice of the first intermediate frequency to be used for the u.h.f. bands as for 70 and 144 Mc/s. It should be borne in mind, however, that the frequency ranges to be covered are somewhat wider and therefore it is desirable to adopt a higher first i.f. for a u.h.f. converter. This also helps to improve second channel rejection.

Although interference from second channel signals is almost unknown in the u.h.f. ranges poor second channel ratio will cause a deterioration in the noise performance of the receiver; for example, a 1 : 1 ratio would increase the noise factor by 3db. Trouble may be experienced, however, in the mixer if the i.f. is too high, because the signal circuit is then so far removed in frequency from the oscillator injection frequency that it becomes difficult to maintain an adequate mixer heterodyne voltage. A suitable first i.f. for a converter for 70cm will be found between 20–30 Mc/s and for 23cm between 30–40 Mc/s; a 70cm receiver which incorporates a good r.f. stage may use the standard i.f. of 10·7 Mc/s or a lower frequency if preferred.

First I.F. Amplifier

The early stages of the i.f. amplifier following a crystal or thermionic diode mixer must be of the low noise type because this type of mixer has a gain less than unity. Low noise is most readily achieved with a double triode valve such as the E88CC, ECC89, 6BQ7A, or ECC84 (6CW7) used as a cascode; see Fig. 5.16. After the early stages, an ordinary pentode may be used but one having a low equivalent noise resistance is preferable.

When the mixer is a triode, connected either in a grounded cathode or grounded grid circuit, a pentode type of first i.f. stage is satisfactory because in effect the first i.f. stage is already a triode.

The mixer output impedance must be suitably matched to the input impedance of the first i.f. amplifier valve. This is particularly important if a crystal or thermionic diode is used because in both of these types the impedance is relatively low (about 500 ohms as a mixer). The circuit design and the precautions necessary to avoid i.f. breakthrough are similar to those relating to v.h.f. receivers.

An i.f. amplifier following a crystal mixer should have a noise factor of 2–4db and following a triode mixer, 6–8db.

LOCAL OSCILLATORS

The local oscillator voltage for a u.h.f. converter may be generated either at a frequency near to the signal frequency (i.e., plus or minus the first i.f.) or at a lower frequency and multiplied. These alternatives have been discussed earlier in this chapter in connection with 2m and 4m converters. The same considerations apply to u.h.f. receivers but because of the higher frequencies involved frequency stability is even more important. Receivers for telephony reception on 70cm can be built using a fundamental frequency tunable oscillator but for c.w. the oscillator must be on a lower frequency (for example, 200 Mc/s). All practical receivers for 23cm employ a crystal oscillator and frequency multiplier chain, as do many for 70cm.

Self-excited Oscillators

In addition to the various stable oscillators already described for use in 2m and 4m equipment, a type designed by RCA in Switzerland is useful over the frequency range between 420 and 1300 Mc/s. It can be a basis for experimental work as a receiver oscillator and is also invaluable as a signal source and as a form of grid dip oscillator. It is relatively easy to construct and employs a readily available receiving type valve, the 6AF4A.

The basic circuit is shown in **Fig. 5.39** while, a picture of such an oscillator is shown, with the end cover removed. The tuned circuit is of the butterfly type, but the inductive element is formed by the metal box which contains the oscillator and is tuned by a split-stator capacitor con-

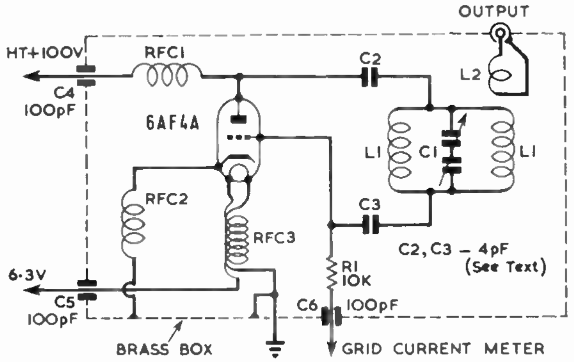


Fig. 5.39. Circuit of u.h.f. oscillator. C1, see Table 5.7; C2, 3, approximately 4 pF (see text, value between 2 and 6 pF); C4, 5, 6, 100 pF leadthrough capacitors; L1, sides of box (see Table 5.7); L2, coupling loop; RFC1, 15 turns 30 s.w.g. enam. $\frac{1}{8}$ in. i.d. $\frac{1}{2}$ in. long; RFC2, 8 turns 28 s.w.g. enam. $\frac{1}{8}$ in. i.d. $\frac{3}{16}$ in. long; RFC3, 15 turns 28 s.w.g. enam. $\frac{1}{8}$ in. i.d. $\frac{1}{2}$ in. long (two wires wound as a twisted pair).

TABLE 5.8

Dimensions of boxes for use with four types of capacitor in the oscillator circuit of Fig. 5.39

Capacitor Eddystone Type No.	Capacitance	Inside dimensions of box			Approx. frequency range
		Length	Width	Depth	
551	24 + 24 pF	$1\frac{3}{8}$ in.	$\frac{3}{4}$ in.	$1\frac{3}{8}$ in.	800-1150 Mc/s
587	10 + 10 pF	$2\frac{3}{4}$ in.	$1\frac{3}{8}$ in.	$2\frac{5}{8}$ in.	610-925 Mc/s
584	34 + 34 pF	$2\frac{3}{4}$ in.	$1\frac{3}{8}$ in.	$2\frac{1}{4}$ in.	420-850 Mc/s
1068	40 + 40 pF	5 in.	$2\frac{3}{4}$ in.	$2\frac{1}{2}$ in.	300-550 Mc/s

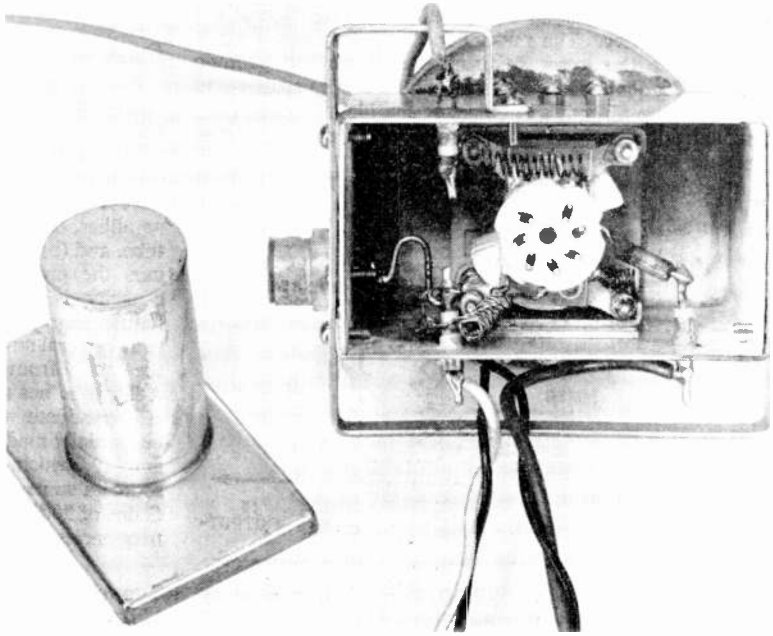
nected across the opposite sides of the box. This results in a rigid construction having the wide tuning range associated with butterfly circuits. Drift due to variations in valve capacitance and loading are reduced by effectively tapping the grid and anode of the valve down the tuned circuit by capacitive potentiometers formed by built-in capacitors and the valve electrode capacitance. The effect of the valve on the circuit can be reduced still further by making these capacitors smaller but the output voltage will fall, the valve ceasing to oscillate if the values are too small; conversely if the values are increased, the output will increase but the drift will also be increased. Typical values are between 2–6 pF.

The box is made of about 18 s.w.g. brass or copper sheet and the seams soldered: its height should be just sufficient to accommodate the split-stator capacitor with the valve holder mounted on the end, the valve itself being shielded by a metal chimney or a well fitting cover. If the box is too high the end space will act as a cavity and produce spurious modes of oscillation. The split-stator capacitor should have brass plates soldered across the sides of the fixed vanes and the plates clamped to the sides of the box by a double row of screws. Unless the sides of the box are in intimate contact with the capacitor throughout its length the Q of the circuit will be reduced. The width of the capacitor with the plates in position decides the internal width of the box: the remaining sides of the box should have a dimension of twice the width.

The mid-frequency and tuning range depend on the size of the box and the capacitance swing of the split-stator capacitor. As relatively few suitable capacitors are available the choice is somewhat limited but Table 5.8 gives four types of Eddystone capacitor, the frequency range for each and other relevant dimensions. The built-in capacitors are formed of two brass plates fixed by small screws (10 B.A.) to the fixed vanes of the tuning capacitor but insulated therefrom by several sheets of thin polythene or p.t.f.e. film (about 0·01 in.). The drilling and tapping must be done with great care.

The two anode tags of the valveholder, which must be made of p.t.f.e., are soldered directly to one plate and the two grid tags to the other. The heater, cathode and anode supplies are filtered with chokes and by-passed to the box with leadthrough capacitors which provide the external connections. A coupling loop is mounted in the box at the end remote from the valve and adjusted so that connection of an external 70 ohm load does not cause the oscillation to cease.

The earthy end of the grid leak is brought out through a leadthrough capacitor and is available to check the grid current. With no anode voltage applied the coupling loop may be connected by a short length of cable to another loop coupled to another oscillator or transmitter, and the tuned box used as an absorption wavemeter, the grid current providing indication. Alternatively, with anode voltage



The u.h.f. oscillator of Fig. 5.39 with the valve removed showing the construction.

applied, the grid current will kick downwards when an external circuit is tuned through the oscillator frequency, the box acting as a grid dip oscillator. At u.h.f., however, it is not recommended that the oscillator be tuned across the band to find the resonant frequency of an unknown circuit because there are various points at which dips occur, due to minor irregularities in the box and coupling loops which make identification difficult.

When the oscillator is used as a source of signal it may be anode modulated with tone by a small a.f. oscillator, or frequency modulated by 50 c/s if the bottom end of the grid leak is connected to the live 6·3 volt heater supply.

Crystal Oscillator Multiplier Chains

The provision of the final oscillator frequency by multiplication from a lower frequency crystal has been referred to already in connection with 2 and 4m receivers. The information generally applies to 70 and 23cm equipment apart from the fact that more multiplying stages are required. Because the injection frequency required is much higher it is however desirable to use a higher frequency crystal in order to keep the number of stages down to a reasonable figure. Whilst lower frequency crystals are cheaper the increased cost of the extra stages must be considered. An additional advantage of the higher frequency crystals is that less difficulty is likely to arise due to self-generated whistles in the receiver and there is less likelihood of interference to other users by spurious radiation. For example, a 70 Mc/s overtone crystal is quite satisfactory for both the 70 and 23cm bands and the frequency is clear of Bands I, II and III television and sound broadcasting. Wherever possible, the crystals chosen should have a low temperature drift. Alternatively, they can be mounted in a simple oven to give a reasonable control of temperature.

Overtone crystal oscillators tend to give rather lower output than those designed for use with fundamental crystals and this may involve an additional amplifying stage at the crystal frequency. The multiplier stages may use a small pentode for the lower frequencies such as the pentode section of a 6U8 or a 6AK5 and double triodes such as the 6J6 used in push-pull as treblers at the higher frequencies. Small double tetrodes such as the QQV02/6 are very useful as frequency treblers up to 600 Mc/s.

In a 23cm multiplier chain the last doubler from 600–1200 Mc/s presents a problem, and one solution is to use a triode of the planar or disc-seal type in a suitable cavity. Valves that may be employed are the types DET29, 2C40, 446A, 6BY4 or 7077. It is possible, however, to employ a semiconductor diode as a frequency multiplier, the circuit in Fig. 5.40 using a 1N64 crystal multiplying from 200–600 Mc/s. This arrangement will furnish a crystal current of 400 μA in a normal crystal mixer. Similar methods can be employed at lower frequencies or at higher frequencies for doubling to 1200 Mc/s although for the latter the crystal would need to be mounted in a cavity similar to that for a crystal mixer. A suitable crystal for multiplier service has a low forward resistance and is capable of passing a high current of the order of 50 mA. The high current is necessary because the harmonic component is only a small percentage of the total current; a low current would result in a very low output. Crystal multipliers must be followed by selective tuned circuits or a high Q break because the output of lower frequencies is considerable and will deteriorate the mixer noise factor.

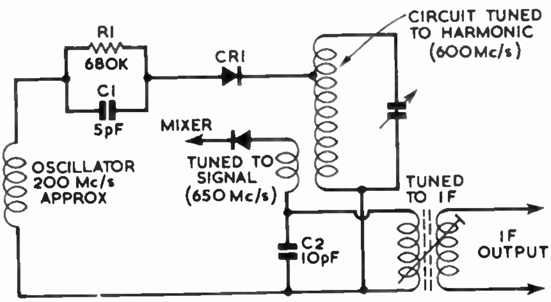


Fig. 5.40. Circuit of a crystal frequency multiplier from 200 to 600 Mc/s. The resistor-capacitor combination in series with the crystal is to provide automatic bias to the correct operating point for multiplication.

R.F. AMPLIFIERS

Because mixers of any type are inherently more noisy than amplifiers it is desirable that the mixer in a low noise u.h.f. receiver be preceded by at least one r.f. amplifier stage. Such a stage, because it usually involves additional tuned circuits, improves the second channel ratio and further reduces the noise. For this reason alone it is often worthwhile.

R.F. amplification on the u.h.f. bands can be achieved by using (a) grounded grid triodes, (b) travelling wave tubes, (c) parametric amplifiers, (d) tunnel diodes, and (e) masers. Travelling wave tubes are very expensive and do not effect much improvement on triode valves at frequencies below 1300 Mc/s.

Parametric Amplifiers

Parametric amplifiers are a comparatively recent development in the field of low noise amplification. As the name suggests, they operate by virtue of a varying circuit parameter, in this case, reactance, which is made to store and release signal energy without adding noise. Two basic types have been developed to date: (a) the beam parametric amplifier, or Adler tube, which resembles a travelling wave tube, and (b) the semi-conductor parametric amplifier which uses the property of a semi-conductor that its junction capacity is proportional to the applied voltage. Both require an external low power " pump " oscillator operating at an integral multiple of the signal frequency, which provides the energy from which the amplified signal is derived. The Adler tube has superior properties and performance to the semi-conductor type, but has the disadvantage that it must be specially made for the frequency band over which it is to amplify, which makes the tube and its associated pump source a somewhat expensive proposition for the amateur. Gains of 20db with noise factors of 1–3db are obtainable at frequencies in the 200–800 Mc/s region.

Tunnel Diodes

The high frequency performance of the tunnel diode has been steadily improved due to better manufacturing tech-

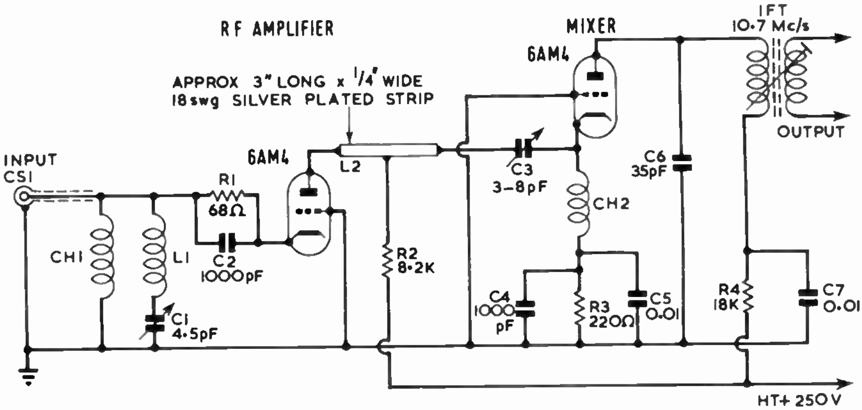


Fig. 5.41. R.F. amplifier and mixer for 430 Mc/s. C1, 4·5 pF trimmer (Wingrove and Rogers type C.3201); C2, 4, 1000 pF (T.C.C. type CTH310); C3, 3-8 pF Philips trimmer; C5, 7, 0·01 μF (T.C.C. type 543); C6, 35 pF (T.C.C. type SNB101); CH1, 2, 24 s.w.g. enam close wound 1/8 in. i.d. 1 in. long; L1, hairpin loop 1 in. long 3/16 in. wide 18 s.w.g.; L2, silver plated brass strip 3 in. long 1/4 in. wide spaced 1/8 in. from chassis; R3 220 ohms.

nique and in suitable circuitry is capable of low noise amplification up to several hundred megacycles. The tunnel diode is a negative resistance device analogous to a Q multiplier and a typical method of use is to connect the diode across a part of the tuned input circuit of a receiver, so raising the effective Q of the input circuit and increasing the gain from the aerial input to the input of the following r.f. stage. Since the tunnel diode operates at a current of the order of 1 mA, and its major noise contribution is due to shot noise, low noise amplification results. In practice it is desirable to incorporate some form of d.c. stabilization of the operating point, as negative resistance devices are prone to instability unless this precaution is taken. A germanium tunnel diode used in this way is capable of gains of the order of 25db with 3db noise factor at 100 Mc/s.

Masers

The maser (microwave amplification by stimulated emission of radiation) is capable of very low noise amplification in the microwave region and noise factors of 0·1db are theoretically possible when the crystal and associated circuits are cooled to the temperature of liquid helium. The mode of operation is briefly as follows. Certain crystalline materials such as ruby or rutile when doped with traces of chromium exhibit the property of amplifying microwave energy within certain well defined frequency bands because of resonances occurring within the structure of the molecules of the material. The maser may take two forms: (a) the crystal is mounted in a cavity and operates as a relatively narrow band two terminal device, or (b) the crystal material may be distributed along a slow wave structure similar to that used in a travelling wave tube when a unilateral four terminal device results with a much greater bandwidth. Current development is directed at improving the travelling wave maser.

Valve U.H.F. Amplifiers

As r.f. amplifiers for u.h.f tend to be more experimental than the remainder of a receiver it is convenient to construct these stages as separate pre-amplifiers connected between the aerial feeder and the mixer stage. In this way, various types can be tried and modifications easily made without involving a complete reconstruction of the receiver or converter.

A 70cm r.f. amplifier and triode mixer using 6AM4 valves is shown in **Fig. 5.41**. The amplifier and mixer are operated in the grounded grid configuration with a strip line as the coupling element between the valves. This line is series tuned at the mixer end with a form of pi network which matches the impedance of the line down to the cathode impedance of the mixer. The aerial feeder is connected directly to the r.f. stage cathode via an isolating capacitor but a tuned input circuit may be included if trouble is experienced from interference from local stations or i.f. breakthrough.

The local oscillator injection may be coupled to the strip line by a loop near the line or coupled directly to the line by a small capacitor (about 1 or 2 pF) connected near either end. The tuning capacitor of the first i.f. transformer must be wired from the mixer anode pin directly to chassis in order to bypass the signal and oscillator frequencies, and should not be across the coil in the i.f. transformer can. The cathode bypass capacitance comprises two capacitors, one for signal frequencies and one for i.f. The h.t. is connected to the line via a decoupling resistor at a voltage anti-node about halfway along the line, somewhat nearer to the r.f. valve than the centre. This point is found by temporarily connecting the h.t. to the centre and by touching the line with a pencil and retuning if necessary until a point is reached where the pencil has no effect. The h.t. should then be permanently connected to this point. This arrangement has a bandwidth at the 3db points of about 6 Mc/s and used with a normal 10·7 Mc/s i.f. has a noise factor of 7–8db.

Fig. 5.42 is the circuit of a two stage pre-amplifier for 70cm using two CV354 (DET23 or TD03–5) disc seal valves. It uses a trough line tuned in the same way as the 6AM4 r.f. amplifier. It requires a separate heater transformer with a winding for each valve as the types suggested have a common heater and cathode connection. The bandwidth is about 6 Mc/s, the gain 20db and the noise factor, used in front of a receiver of noise factor 8db, is between 3 and 4db.

The type 12AT7 and 6BQ7A valves may be used in r.f. stages for 70cm and can be conveniently arranged as pushpull grounded grid amplifiers using two parallel lines series tuned at the end remote from the valve, with h.t. fed on to the lines at anti-nodal points. They require neutralizing, how-

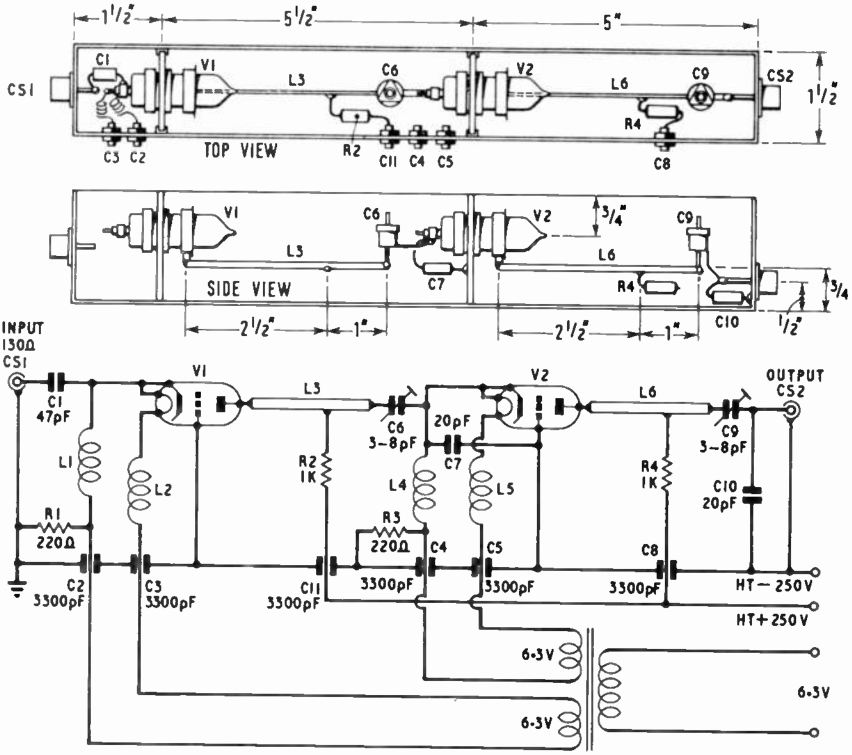


Fig. 5.42. Two stage 430 Mc/s pre-amplifier using CV354 valves coupled with strip lines in a trough. C1, 47 pF (T.C.C. type SCT1); C2, 3, 4, 5, 8, 11, 3300 pF (T.C.C. type CTH315/LT); C6, 9, 3-8 pF Philips trimmer; C7, 10, 20 pF silvered mica (T.C.C. type SMB101 or SCT11); L1, 2, 4, 5, r.f. chokes ⅛ in. diam. 1 in. long wound with 24 s.w.g. enam.; L3, 6, 3½ in. by ¼ in. silver plated 18 s.w.g. brass strip; V1, 2, DET23 (CV354). CS1 and CS2 are Belling & Lee co-ax sockets type L34S.

ever, which can be readily done with two short lengths of wire (about 1 in.) cross-connected from the cathodes and lying near to the opposite anode lines. The 6BQ7A is the slightly better valve. Noise factors of about 10db are obtainable from such a stage.

TABLE 5.9

R.F. AMPLIFIER VALVES

Type No.	Construction	Mounting	70cm Gain db	70cm N.F. db	23cm Gain db	23cm N.F. db	Slope mA/V
12AT7	Normal miniature	Glass B9A base	10	10	—	—	5.5
6BQ7A	Normal miniature	Glass B9A base	12	9	—	—	6.4
6AM4	Frame grid	Glass B9A base	14	8	—	—	9.8
PC86	Frame grid	Glass B9A base	15	7	—	—	14.0
A.2521/A.2599	Planar	Glass B9A base	18	7	—	—	15.0
DET29	Disc seal	Coaxial	—	—	13*	11*	14.0
EC56	Disc seal	Coaxial	—	—	—	—	16.0
6299	Disc seal	Coaxial	17.5	4.5	16	8.5	12.0
6280/416A	Disc seal	Coaxial	20	4.0	—	—	50.0
6BY4	Stacked Ceramic	Coaxial	14	6.0	14	8.5+	6.0
7077	Stacked Ceramic	Coaxial	15	4.5	15	8.0	9.0
6CW4	Nuvistor	Twelvar 5 p.	20	6	—	—	12.5

*at 2300 Mc/s   + at 900 Mc/s

Note: gain and noise figures are approximate depending on circuit and bandwidth.

There are at present few valves suitable for r.f. stages at 23cm and all are of the coaxial disc seal type. Examples are the DET29, EC56, 6299, 6280, 6BY4 and 7077. Abridged data, unfortunately incomplete on these valves and others for 70cm and 23cm, is shown in **Table 5.9.**

Input Matching

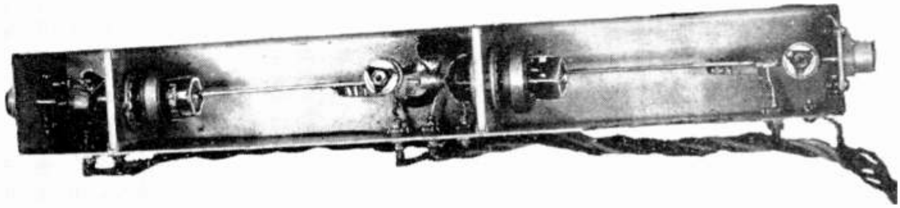
In order to achieve maximum efficiency, it is necessary that the aerial feeders be matched to the receiver input. This is even more important on 23cm because the mismatch is likely to be more severe. An improvement of the order of 3db or more can be obtained by using a matching device. One such device is a two stub tuner, which consists of two adjustable short-circuited stubs of maximum length ⅜ wavelength spaced ⅛ wavelength apart, which can be connected by plugs and sockets between the feeder and receiver input. The device should be adjusted on a signal for the best signal-to-noise performance.

TYPICAL U.H.F. CONVERTERS

SIMPLE 420 MC/S CONVERTER

The circuit of a simple but effective 70cm converter designed by G2DD is shown in **Fig. 5.43** (a). It can be built on a standard Eddystone die-cast box type 650 measuring 4¼ in. × 3¾ in. × 2 in.

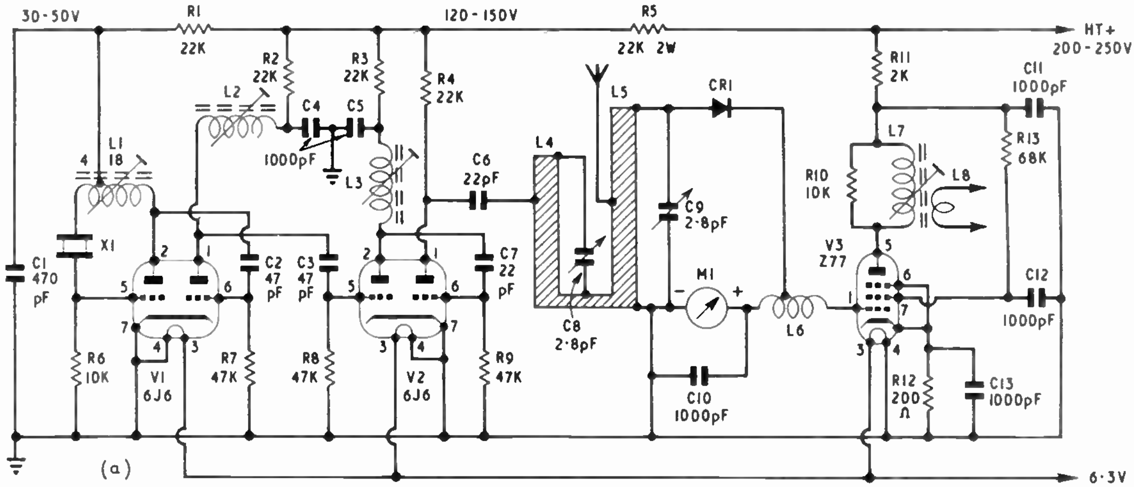
A crystal frequency of about 7740·6 kc/s will produce an i.f. in the 17 Mc/s range but if a higher i.f. is desired a crystal of 7555·5 kc/s would produce an i.f. of 27 Mc/s. The first



The two stage 430 Mc/s pre-amplifier of Fig. 5.42. The chassis should be made from 18 s.w.g. copper or brass silver plated.

triode of V1 together with L1 is the oscillator circuit and operates on the third overtone of the crystal driving the second triode of V1 which operates as a tripler giving an output of 69 Mc/s. The second double triode, V2, doubles to 139 Mc/s in its first half, and triples to 418 Mc/s in the second. The 418 Mc/s output is developed across the strip line circuit, L4, C8. Another strip line, L5, tuned by C9, comprises the tuned circuit for the mixer crystal, the feeder being tapped on to the strip, the d.c. crystal current being monitored by the meter.

The strip lines L4 and L5, are formed from one piece of brass as shown in Fig. 5.43 (b). The output from the crystal oscillator chain is injected into the mixer crystal circuit by mutual coupling between L4 and L5. With no feeder connected, a crystal current of about 300 μA can be obtained but this value will fall about 25 per cent when the feeder is connected. The optimum value for the mixer crystal specified is between 60 and 120 μA with L5, C9 resonated at the signal frequency. The coils L1, L2, L3, L6 and L7 are wound on Aladdin moulded formers and tuned by dust cores, which should be of the high frequency type (coded yellow). Holes for access for adjustment are drilled in the outside of the box.



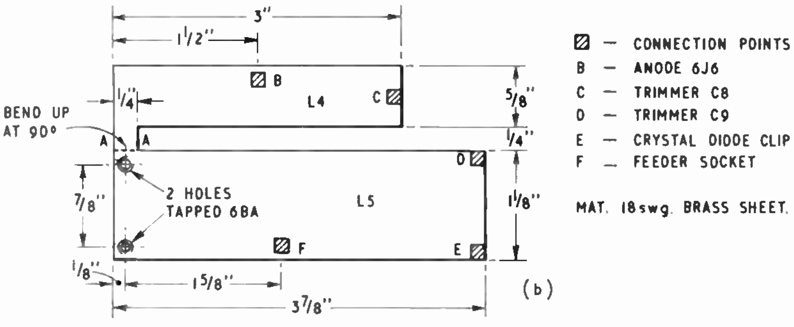
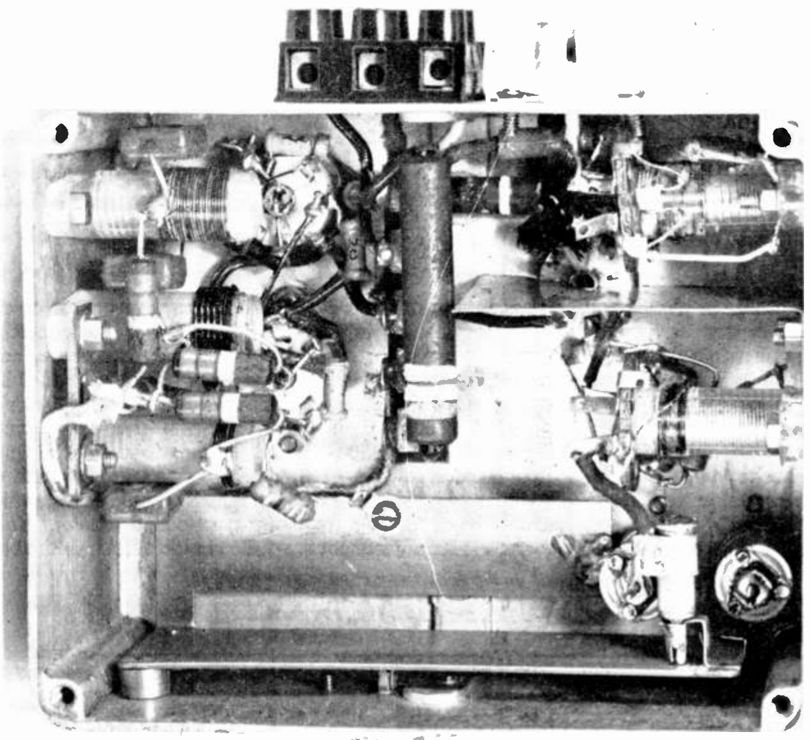


Fig. 5.43(b). Dimensions of the strip line L4 and L5 which is made from a single piece of 16 s.w.g. brass and spaced $\frac{1}{16}$ in. from the chassis, fixed at right angles to each other in a corner of the chassis.

Fig. 5.43(a). C1, 4, 5, 6, 10, 11, 12, 13, midget mica; C2, 3, 7, ceramicons; C8, 9, 3-8 pF Philips trimmers; CR1, silicon diode type CV102 or CV103; L1, 22 turns, 22 s.w.g. enam., close wound, tapped at four turns; L2, six turns, 20 s.w.g. enam., close wound; L3, two turns, 20 s.w.g. enam., close wound; L4, 5 see Fig. 5.43(b); L6, 22 turns, 32 s.w.g. d.s.c., close wound, centre tapped; L7, 37 turns 32 s.w.g. d.s.c. close wound; L8, four turns, 32 s.w.g. d.s.c., wound at earthy end of L7 (all coils except L4, 5, are close wound on ⅜ in. diam. Aladdin formers, and should have high frequency cores); M1, 0-500 μA; X1, 7740·6 kc/s or 23·2 Mc/s overtone crystal.



A view of the underside of the simple 430 Mc/s crystal controlled converter showing the layout and the strip line.

The i.f. output from the mixer crystal is stepped up in L6 and amplified by the pentode V3 and low impedance output to the main receiver obtained from L8.

The underside of the converter from which the layout can be seen is shown in the photograph.

If a tunable converter is preferred so that a fixed i.f. can be used or an i.f. selected which is free from breakthrough, the converter can be modified so that the first section of V2 is used as an oscillator on 104 Mc/s or on 139 Mc/s, the second section acting as a frequency quadrupler or trebler respectively. The oscillator can be bandspread to cover the portion of the 70cm band required. A suitable oscillator coil would comprise 3 turns ⅜ in. i.d. centre-tapped and tuned by a bandset capacitor of 30 pF (Philips trimmer) and bandspread with a 5 pF variable. The second section should be loosely coupled to the oscillator by a wire close to the anode lead.

HIGH GAIN 430 MC/S CONVERTER

This converter employs two A.2521 (6CR4) grounded-grid amplifiers in cascade, an A.2521 quintupler in the frequency multiplier chain and conventional coils in all the tuned circuits. The circuit diagram is shown in Fig. 5.44.

In order to obtain the best noise factor an input tuned circuit is provided in the cathode of V1 and the feeder connected to a tap *via* C32. Aerial matching and tuning are quite critical, the correct tuning position for optimum noise factor being slightly below the signal frequency. Provided that the aerial is properly matched to its feeder and a u.h.f. socket (in preference to a television type) is used on the converter for the aerial connection, the optimum position for the tap will not be far from the centre point.

The coupling between the first and second r.f. stages consists of a tuned transformer comprising two identical coils of 1¾ turns of 18 s.w.g. enamelled copper wire with a ⅜ in. internal diameter, the turns and the coils being spaced by one wire diameter. The coupling between V2 and the crystal mixer (CR1) is identical except that the secondary is untuned.

All the r.f. chokes in the heater and cathode leads are air-cored, consisting of 23cm of 24 s.w.g. wire, formed by close winding on a ⅛ in. diameter rod. Those in the anode circuits, however, are slightly damped by fully overwinding a ¼ watt 100 K ohm resistor with 24 s.w.g. wire.

Local Oscillator Chain

The local oscillator chain uses a double triode valve B.719/ECC85 or 6AQ8 (V5), the first triode operating as a Squier oscillator and the second as a quadrupler. An A.2521/6CR4 (V6) acts as quintupler multiplier to the final frequency. High orders of multiplication have been used to reduce the number of crystal harmonics which can produce spurious responses. The anode circuit of V6 uses coils wound to the same specification as the signal frequency

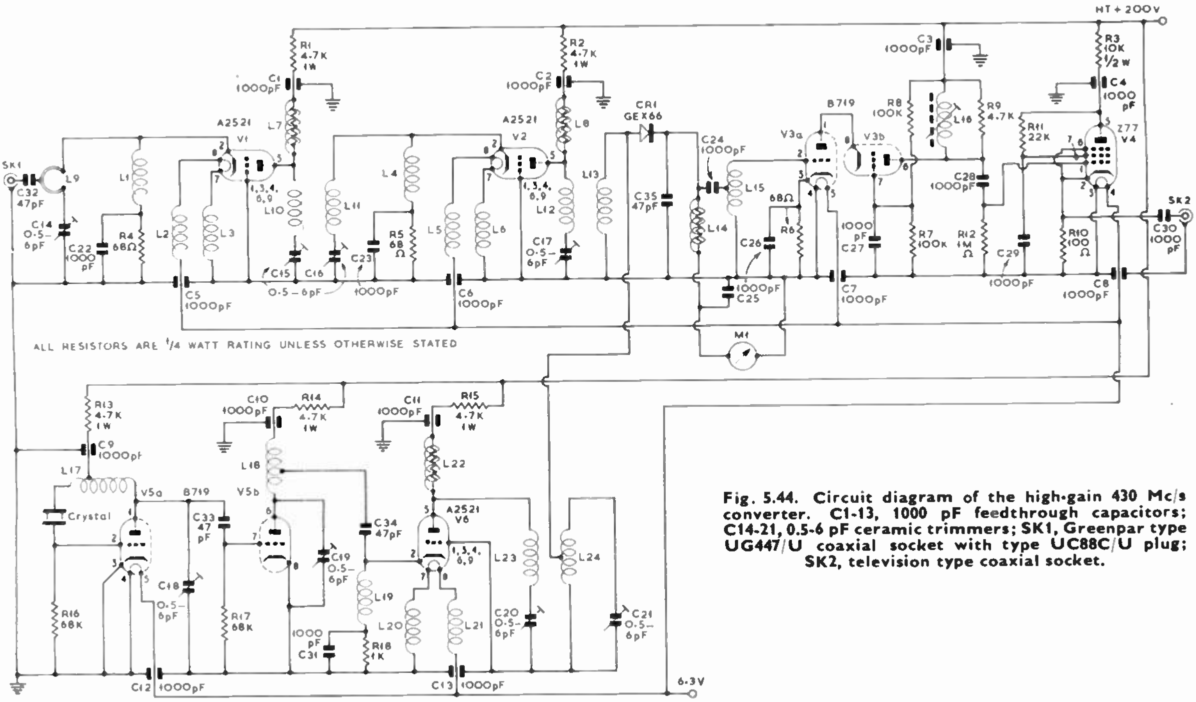


Fig. 5.44. Circuit diagram of the high-gain 430 Mc/s converter. C1-13, 1000 pF feedthrough capacitors; C14-21, 0.5-6 pF ceramic trimmers; SK1, Greenpar type UG447/U coaxial socket with type UC88C/U plug; SK2, television type coaxial socket.

transformers, but the two coils are spaced by about 0·6 in. to assist in the rejection of unwanted harmonics.

A suitable crystal would be one with a frequency of 20·375 Mc/s, giving an i.f. varying between 24 and 30 Mc/s for a signal frequency range of 431·5 to 437·5 Mc/s.

Mixer and I.F. Amplifier

The outputs of the r.f. amplifier and the local oscillator are fed to the GEX66 mixer (CR1) and the i.f. output is taken to a tap on the input coil of the cascode amplifier (V3). The position of this tap has a bearing on the noise performance of the cascode stage, but in view of the generous amount of r.f. gain the noise contributed by V3 should be negligible. A cathode follower is used to convert the output to low impedance before transfer to the tunable main receiver. With some receivers it may be desirable to reduce the gain of the cascode stage to avoid cross-modulation.

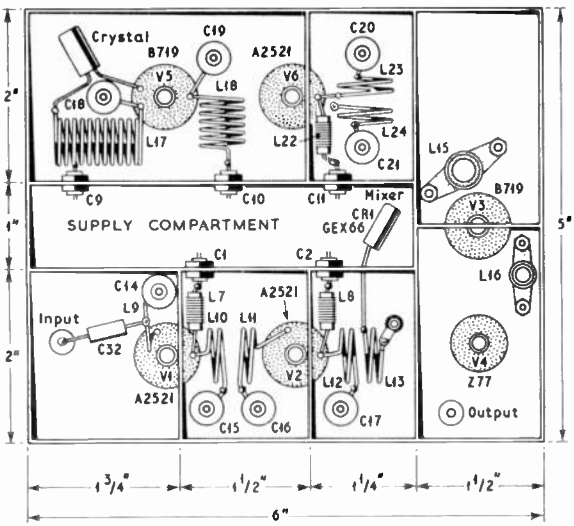


Fig. 5.45. Layout of the principal components in the 430 Mc/s converter.

Construction

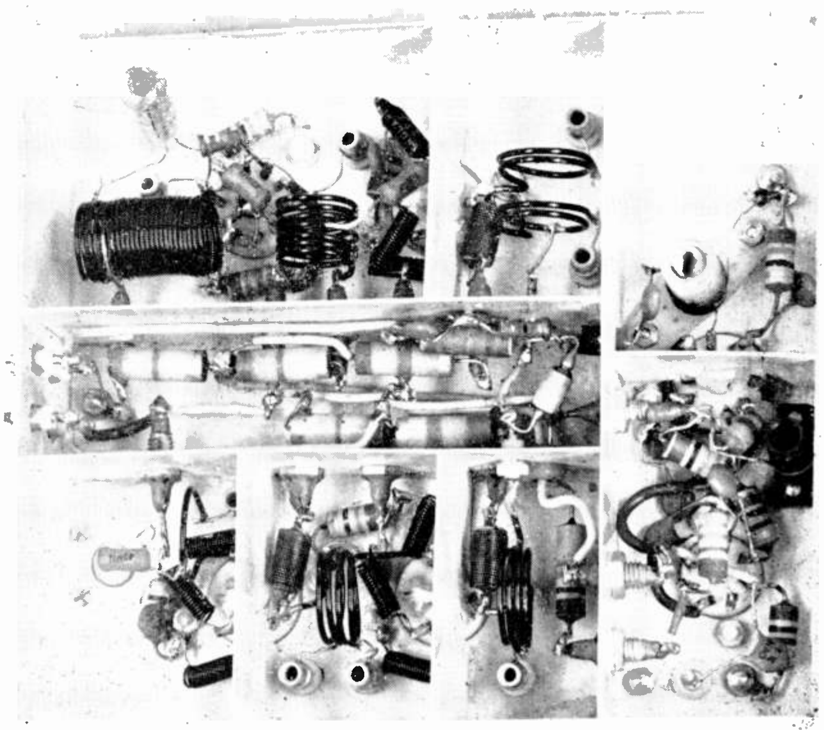
The converter is constructed on a chassis of 20 s.w.g. silver-plated brass measuring 6 in. × 5 in. × 1·5 in. deep. It is divided into compartments by soldering cross screens to the main chassis. See Fig. 5.45. The screens across the A.2521 valveholders need to be particularly close fitting and

TABLE 5.10

Inductor details for the high gain 430 Mc/s converter

L1–6	23cm 24 s.w.g., ⅛ in. diam., air cored.
L7, 8	24 s.w.g., close wound on 100 K ohm ¼ watt resistor.
L9	¾ turn 18 s.w.g., ⅜ in. diam., air cored.
L10–13	1¾ turns 18 s.w.g., ⅜ in. diam., air cored.
L14	44 s.w.g. d.s.c., close wound on 100 K ohm ¼ watt resistor.
L15	25 turns 26 s.w.g., ⅜ in. diam., tapped 7 turns from earthy end.
L16	23 turns 24 s.w.g., ¼ in. diam., ferrite slug tuned.
L17	25 turns 20 s.w.g., ½ in. diam., tapped 5½ turns from crystal end.
L18	6½ turns 18 s.w.g., ½ in. diam., tapped at 3½ turns.
L19	3 μH choke (Painton type 200151).
L20, 21	24cm 24 s.w.g., ⅛ in. diam., air cored.
L22	26 s.w.g., close wound on 100 K ohm ¼ watt resistor.
L23	1¾ turns 18 s.w.g., ⅜ in. diam., air cored.
L24	1¾ turns 18 s.w.g., ⅜ in. diam., tapped at ¾ turn, air cored.

Enamelled copper wire throughout, except where otherwise stated.



An under-chassis view of the converter. The components may be identified by reference to the layout diagram Fig. 5.45.

all the grid tags of the valveholder should be soldered to the screen and not to the main chassis. It is prudent to put a scrap B9A valve into the holder while this is being done, so that the contacts are not pulled out of alignment. P.T.F.E. valveholders were used throughout in the prototype.

A power supply compartment runs down the centre of the chassis and contains the h.t. filter resistors and feed-through capacitors which transfer the supplies to the various compartments. The mixer crystal lies across the end of the power supply compartment.

The converter requires an h.t. supply of 200 volts at 60 mA and a heater supply of 6·3 volts at 1·8 amp. Individual valve currents are:

Stage		*Anode Current*
Oscillator	V5	9·5 mA (total)
Quintupler	V6	8 mA
R.F. amplifiers	V1, V2	15 mA each
I.F. amplifier	V3	5 mA
	V4	5·5 mA
Mixer	CR1	0·5 mA

Performance

Measured under laboratory conditions, the following performance figures were obtained:

Centre frequency	433 Mc/s
R.F. gain before mixer	25db
Noise factor	7db
R.F. bandwidth	4 Mc/s
Image rejection	50db
Intermediate frequency	26·25 Mc/s

A TRANSISTORIZED CONVERTER FOR 432 MC/S

This converter, the circuit of which is given in Fig. 5.46, employs u.h.f. transistors whose prices compare favourably with valves of similar performance. No apology is made for the use of "lumped" tuned circuits which are generally considered to be poor practice at frequencies in excess of 400 Mc/s. They were found to be far easier to tame and to construct than the accepted tuned lines and no difficulty was experienced in achieving the theoretical gain and noise figure for the transistors concerned.

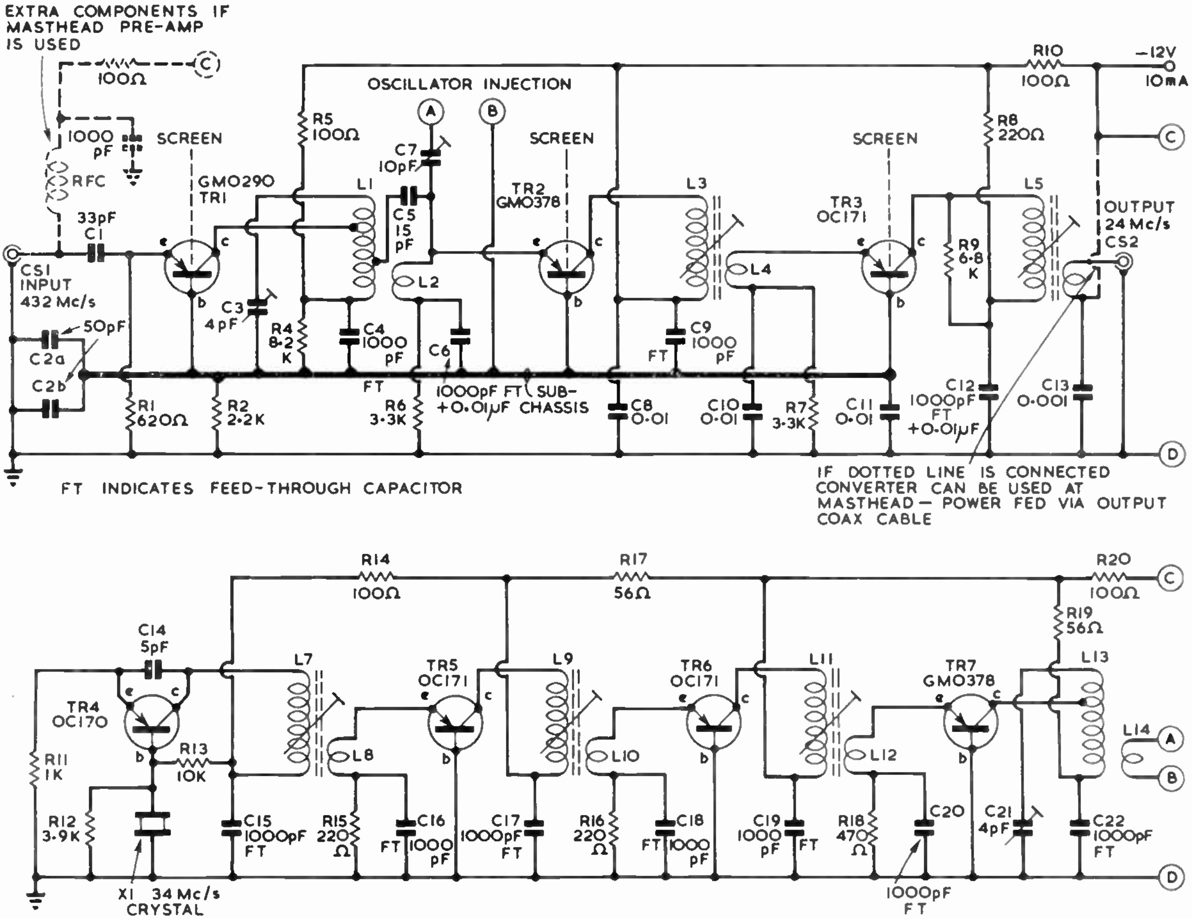
With the GM0290 r.f. stage the noise factor is of the order of 5db and a gain of some 14db is obtainable with a bandwidth of at least 2 Mc/s. If slightly inferior performance can be accepted the somewhat cheaper GM0378 may be substituted for the GM0290 r.f. amplifier.

U.H.F. transistors are particularly sensitive to stray r.f. and every care must be taken to guard against power from the transmitter leaking across the relay employed for change-over switching between receive and transmit. The r.f. stage transistor can easily be ruined if such precautions are not taken.

Circuit

In Fig. 5.46, the upper section of the diagram comprises the r.f., mixer and i.f. amplifier stages and the lower section the crystal controlled oscillator-multiplier chain, the output of which is fed to the mixer, TR2.

TR1 is operated as a grounded base amplifier without input tuning. The output is tuned to 432 Mc/s by L1/C3, the collector being tapped down L1 in the interests of stability. Capacity coupling is employed between TR1 and



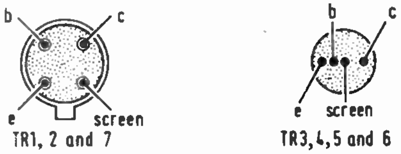


Fig. 5.46. The r.f., mixer and i.f. stages of the transistorised 432 Mc/s converter (upper diagram) and overtone oscillator and multiplier chain (lower diagram).

TR2 as link coupling with the necessarily diminutive coils proved impracticable. L2 may appear to contradict this statement but its actual function is that of an r.f. choke placed in the field of L1 to increase the coupling co-efficient without introducing a high impedance at the intermediate frequencies between base and emitter of the mixer: the inclusion of such an impedance is a common cause of low mixer sensitivity.

Oscillator injection at 408 Mc/s is applied to the emitter of TR2. The capacitor C7 together with the inductance of the coupling loop L14 and the 1 in. connecting lead form a series resonant circuit at 408 Mc/s making for efficient transfer of energy between oscillator chain and mixer.

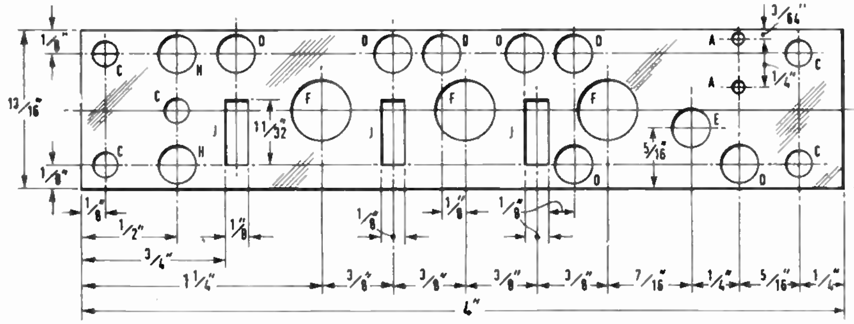
L3 and the output capacitance of TR2 form a resonant circuit at the intermediate frequency of 24 Mc/s. Precise adjustment of the frequency of this circuit is effected by the core of L3. In order to damp L3 in the interests of bandwidth and stability, the coupling coil L4 is made larger than usual, a method found to be more effective than loading L3 with a resistor.

The inclusion of an i.f. amplifier on the converter chassis has certain advantages. It allows L3 and L5 to be tuned for a more level response over the bandwidth of prime interest, and by increasing the level of the i.f. output helps to reduce breakthrough on poorly screened receivers used as the tunable i.f.

TABLE 5.11

COMPONENT DETAILS

C3, C21, 4 pF Wingrove and Rogers airspaced trimmers.
C7, 10 pF Wingrove and Rogers airspaced trimmer.
C6, C8, C10, C11, C12, Mullard miniature paper capacitors, 30V working.
CS1, CS2, Belling Lee L734/S flush mounting co-axial sockets.
Other capacitors are miniature ceramic, except for C13 which is a Radiospares 0·001 μF Stand-off type, and C9 1000 pF Radiospares feed-through capacitors.
Eddystone 650 die-cast box.
All resistors are Radiospares miniature ½ watt.
GM0290 (1 off) and GM0378 (2 off) transistors are manufactured by: Texas Instruments Ltd., Manton Lane, Bedford.
Xtal, Miniature 3rd overtone type HC-6/U, ½ in. spacing, with holder. (Henrys Radio Ltd., Edgware Road, London W.2.)



HOLES 'A' – 10 BA CLEARANCE 1/16" (·062") DIA
'C' – 6 BA CLEARANCE 7/64" (·109") DIA
'D' – 3/16" (·187") DIA OPENED AS NECESSARY FOR FEED-THROUGH CAPACITORS
'E' – 3/16" (·187") DIA OPENED AS NECESSARY FOR FITTING TR7
'F' – DRILL 1/4" DIA FILE TO SUIT 5/16" DIA POLYSTYRENE COIL FORMER FIXED WITH POLYSTYRENE CEMENT
'H' – 3/16" DIA APPROX OPENED FOR CRYSTAL HOLDER
'J' – 1/8"x 11/32" CUT-OUTS TO ACCOMODATE LEADS OF TR4, 5 AND 6

Fig. 5.47. The drilling details for unit 2, viewed from the copper side.

The i.f. amplifier, TR3, requires no particular description. Recourse is made to resistor damping across L5 to assist in achieving the required bandwidth.

In the oscillator-multiplier chain, TR4 operates in an overtone circuit using a miniature HC-6/U crystal at 34 Mc/s. This circuit is perfectly satisfactory for overtone crystals of this type but difficulty may be experienced if FT243 crystals are substituted.

L7 and the output capacitance of TR4 tune to 34 Mc/s. This circuit is link coupled to TR5 by L8, the output of TR5 being tuned to 68 Mc/s, bias being obtained by base/emitter rectification of the drive from L8. Supplementary bias is derived from R15 which has been chosen for maximum output from TR5. To some extent the value of R15 is a compromise between efficiency, output, and the danger of thermal runaway. Increasing its value increases the efficiency of the stage but at the expense of peak output: reducing it below the value given, while enhancing the output, holds the danger of thermal runaway under conditions of high ambient temperature.

TABLE 5.12
COIL DETAILS

- L1 2 turns, ¼ in. diam. ½ in. long, 18 s.w.g. tinned copper wire, with leads ⅜ in. long. The collector of TR1 is connected to main body of coil at C3 end. C5 is connected to main body of coil at C4 end. Self supporting.
- L2 1 turn, ¼ in. diam., 22 s.w.g. tinned copper wire, spaced ¼ in. from and connected " in phase " with L1. Self supporting.
- L3 5/16 in. diam., ⅜ in. long, 36 s.w.g. enamel close wound. Dust iron slug tuned.
- L4 4 turns, 7/42 p.v.c. insulated (Radiospares) wound over L3.
- L5 5/16 in. diam., ⅜ in. long, 36 s.w.g. enamel, close wound. Dust iron-slug tuned.
- L6 2 turns 7/42 p.v.c. insulated (Radiospares) wound over L5.
- L7 5/16 in. diam., 9/16 in. long, 26 s.w.g. enamel, close wound. Dust iron slug tuned.
- L8 3 turns 7/42 p.v.c. insulated (Radiospares) wound over L7.
- L9 5/16 in. diam., ¼ in. long, 26 s.w.g. enamel, close wound. Tuned with Salford 200 Mc/s dust core, coded blue.
- L10 1½ turns, 7/42 p.v.c. insulated, wound over L9.
- L11 8 turns, 5/16 in. diam., 22 s.w.g. tinned copper, spaced by wire diameter. Dust iron slug tuned, as L9.
- L12 1 turn, 7/42 p.v.c. insulated wire, wound over L11.
- L13 2 turns, ¼ in. diam., ¼ in. long, 22 s.w.g. tinned copper, with ½ in. leads at C21 end (TR7 collector is tapped into the main body of coil), and ¾ in. long at C22 end. Self supporting.
- L14 1 turn, ¼ in. diam., 22 s.w.g. tinned copper, with ⅜ in. and ¼ in. leads interwound with L13 (⅛ in. spacing from coil L13).

The output of TR5 is coupled to TR6 by L10, TR6 acting as a doubler from 68 to 136 Mc/s. This circuit is identical with that of the previous stage, the output coil, L11, being resonated by the output capacitance of the transistor. Finally, TR7 operates as a frequency tripler from 136 Mc/s to 408 Mc/s. L13 and C21 are tuned to the latter frequency, the collector being tapped down L13 in order to achieve the optimum Q. The output link, L14, which has been mentioned previously when dealing with the mixer, is positioned in the centre of L13.

Construction

The converter is in two units, the r.f. section and the oscillator-multiplier circuits, and both are housed in an Eddystone die-cast box type 650 which provides complete screening.

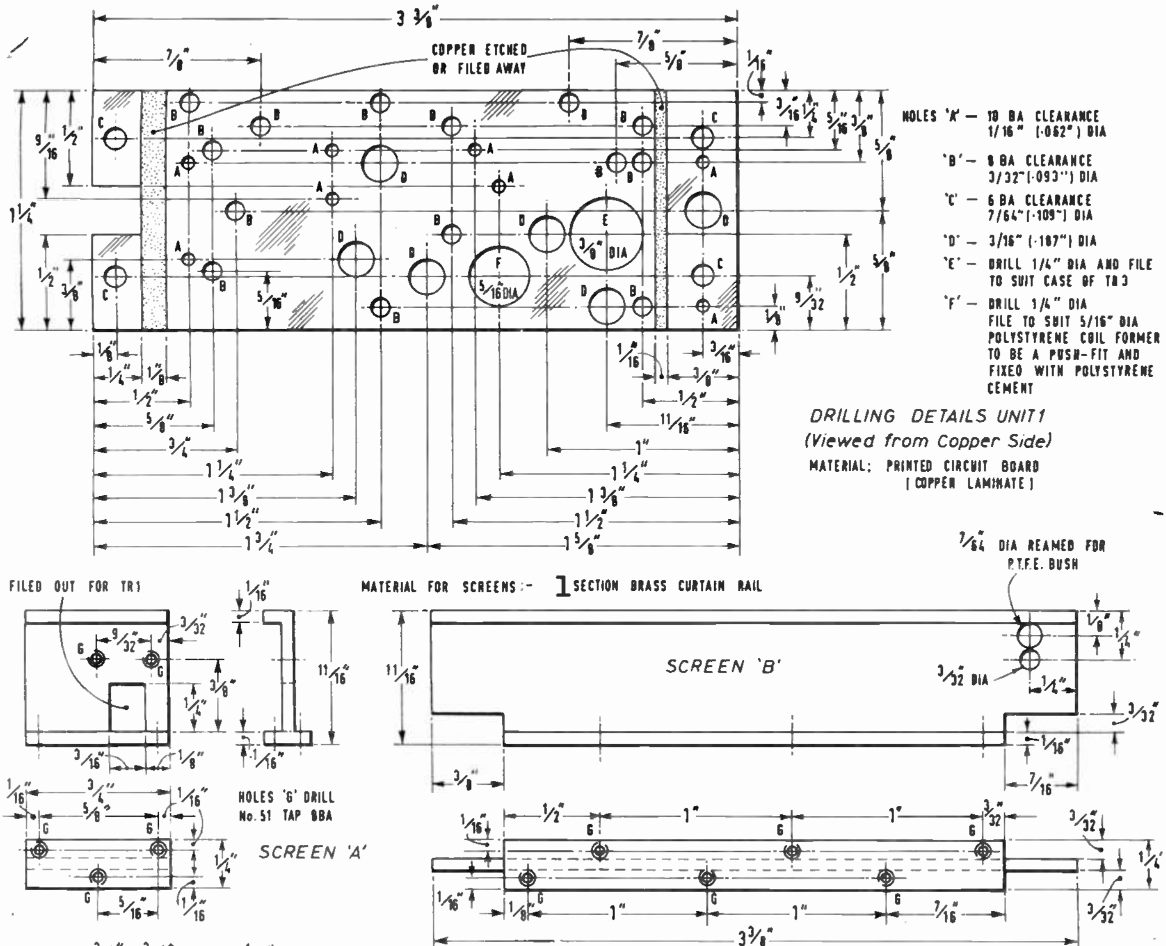
The working diagrams, **Figs. 5.47 to 5.49** show in detail the component parts, construction and assembly of the converter. Fig. 5.47 shows the drilling of the copper laminate board used for the oscillator-multiplier chain, Fig. 5.48 gives not only the drilling details of the copper laminate board for the r.f. section but also details of the screens. Fig. 5.49 shows how the assemblies are fitted into the die-cast box.

One detail which is not illustrated but does require mention is the mounting of the sub-assemblies on the lid of the box. Each sub-assembly is mounted on four brass pillars so that it stands away from the lid. The pillars supporting the r.f. board are ⅜ in. long and those for the oscillator-multiplier board are 1⅛ in. long. Both sets of pillars are drilled and tapped 6 B.A. at each end.

Adjustment

Transistors are easily destroyed by the application of incorrect voltages, so before applying power to the converter a most careful visual check should be made to ensure that all connections are correct. Above all, make sure that the battery polarity is correct.

The oscillator-multiplier chain should be adjusted first. While monitoring 34 Mc/s, either on a receiver or by means of a sensitive wavemeter, adjust the core of L7 to the point where there is no appreciable change in frequency even though the core is adjusted slightly above and below this



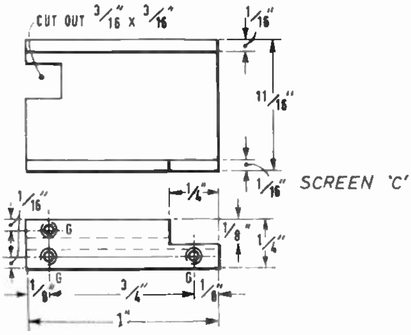


Fig. 5.48. The dimensions for unit 1, with the associated screens.

setting. By switching the power on and off, check that the oscillator remains on frequency. If it does not, then the oscillation is not controlled by the crystal, and further adjustment to the core of L7 is needed.

Should the crystal fail to oscillate at all, either try another OC170, adjust the bias to this stage, or ease off the coupling winding L8 until the circuit is operating in a satisfactory manner.

Once the crystal oscillator is functioning, connect a voltmeter across R15 and adjust the coupling of L8 to L7 for maximum drive to TR5 consistent with reliable operation of the crystal oscillator. Not less than 0·2 volts should be obtained.

As a final check on the crystal oscillator, and with the meter across R15, remove the crystal. No voltage should now be present across R15. If by chance there is, then the crystal oscillator stage is self-oscillating.

Connect the voltmeter across R16 and adjust the slug in L9 for maximum reading. Check that L9 is tuned to 68 Mc/s by means of a wavemeter.

Connect the voltmeter across R18, adjust the slug of L11 for maximum indication on the meter and check that the frequency at L11 is 136 Mc/s.

As the tuning range of the combination C21/L13 in the collector circuit of TR7 is limited, a wavemeter is not absolutely necessary to check the frequency of this circuit. To adjust L13, temporarily disconnect C7, and from point A on the output link of L14 connect a crystal diode and sensitive microammeter in series to earth, bypassing the

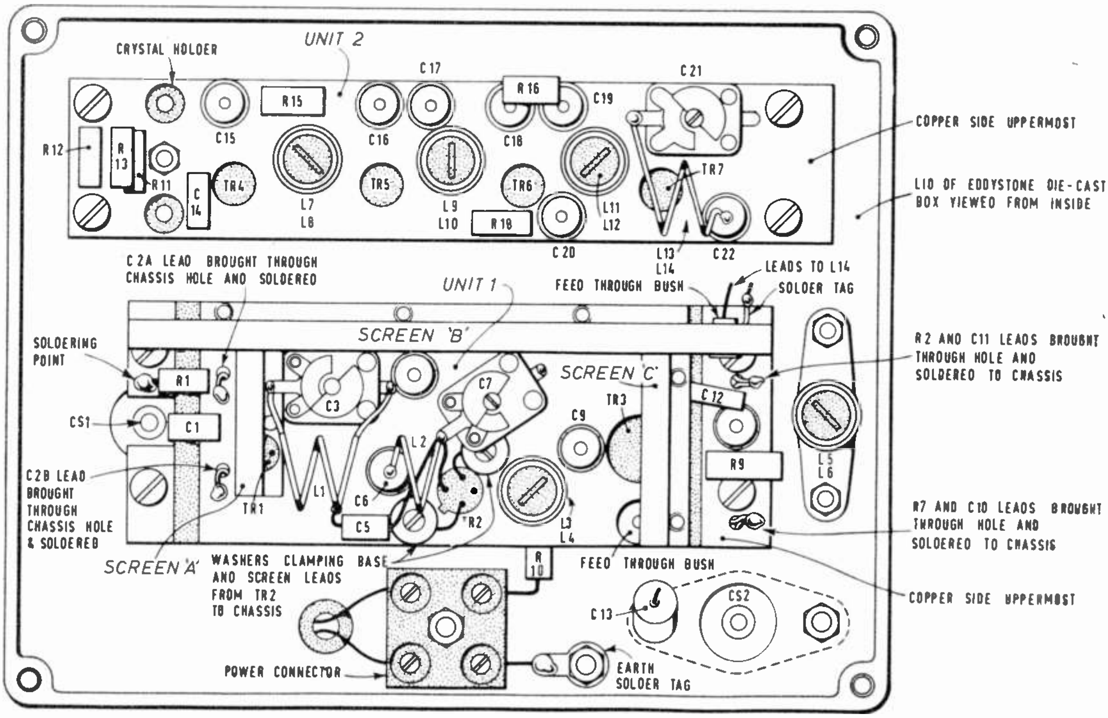


Fig. 5.49. View showing position of units when fitted into main housing. Location of major components of units also shown.

meter with a small capacitor. Adjust C21 for maximum indication on the microammeter.

Alternatively, a voltmeter may be connected across R6, and with C7 set to half capacity, C21 should be adjusted for a rise in the standing reading. This rise will be quite small, but just perceptible on the 10 volt range of an Avo Model 7 (1,000 ohms per volt). Finally, C7 is adjusted to the position which causes a further rise in the standing voltage.

As a final check that all is well with the oscillator-multiplier chain, remove the crystal. All drive to the mixer TR2 should cease.

The converter may now be connected to the main receiver via a length of coaxial cable. Tune the receiver between 24 Mc/s and 26 Mc/s and adjust the cores of L3 and L5 to give an even noise response over this frequency range. Attention to the values of R7 and R9 and/or the direction of the windings of L5 and L6 should overcome any regeneration causing peaks in the noise response. When making these adjustments it may be necessary to remove the crystal and also R1. Such steps are however only needed if it is suspected that part of the noise being heard comes from the front-end of the converter, and not from the i.f. amplifier alone. After restoring any connections removed, adjust C3 for maximum noise.

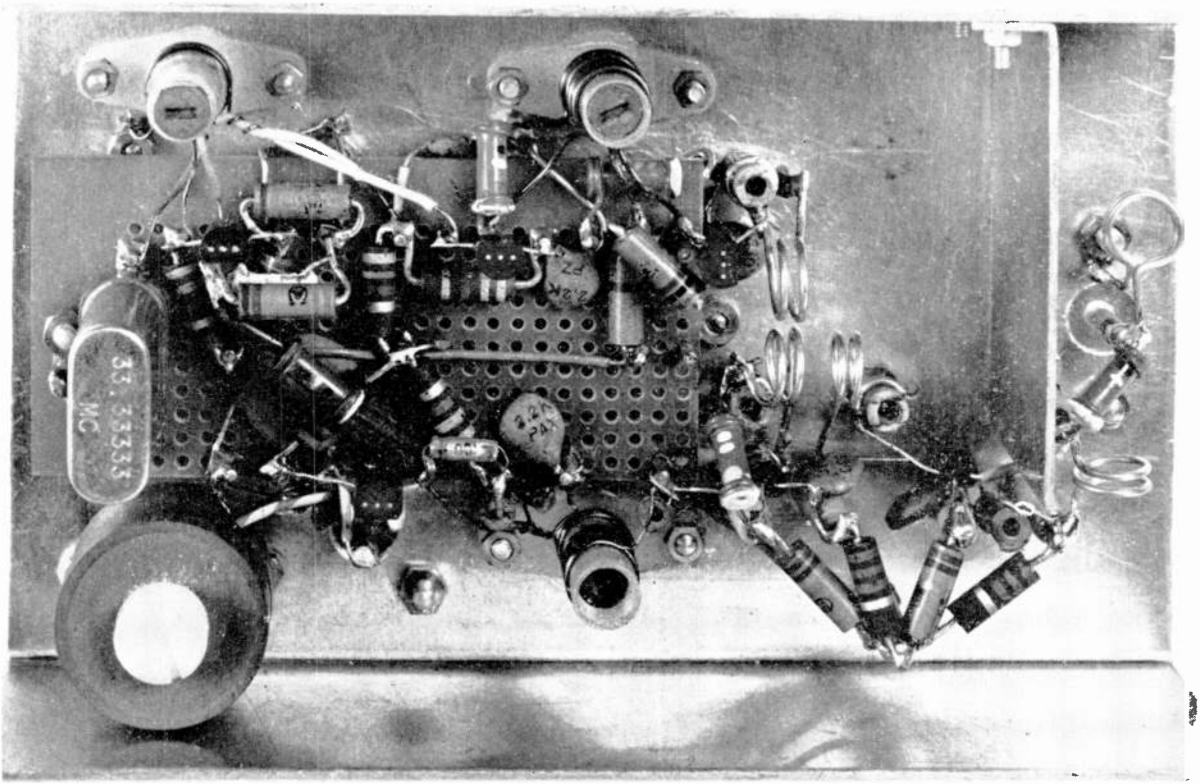
For the next adjustments either a local signal on the 430 Mc/s band is required, or a signal from a g.d.o. or a signal generator. If either of the last two sources is being utilized and it is necessary to employ a harmonic, be quite sure that the correct harmonic has been identified and that in addition reception is not occurring on the image frequency which is 384 Mc/s (408 Mc/s less than the i.f.).

Adjust C3, C7 and C21, in that order, for maximum output. Final adjustments are made on weak signals in the usual manner.

This converter has been found to have a weak signal performance rather better than a valve type using an A.2521 valve. There are, however, certain shortcomings which are small in the light of the overall performance. These are a tendency to cross-modulation from very powerful nearby stations, and also, under conditions of very high local field strength, stations in the 90 Mc/s f.m. band can beat with the 68 Mc/s multiplier stage to produce the i.f. of 24 Mc/s. Both of these faults could be cleared, the first by arranging a gain control on the r.f. amplifier, and the second by the use of a suitable trap in the input circuit. However, since the conditions which cause these effects are rarities, their inclusion was not felt to be warranted.

GM0290 and GM0378 transistors may be mounted in TO-15-2 encapsulations, designated GM0290A and GM0378A respectively. The connections are in line and with the pip on the case downwards, the left-hand lead is the base, followed by the collector, emitter and case in sequence.

The dotted connections and components in Fig. 5.46 associated with the input and output of the converter are, in the case of the input, for feeding power up the coaxial feeder to a masthead pre-amplifier, should this be required. Similarly the addition of the dotted connection to the output link would enable the whole converter (suitably weatherproofed) to be fitted at the top of the mast and supplied with power from the ground via the coaxial cable joining converter and main receiver.



The completed 430 Mc/s converter using 2N3819 field effect transistor.

FET Converter for 430 Mc/s

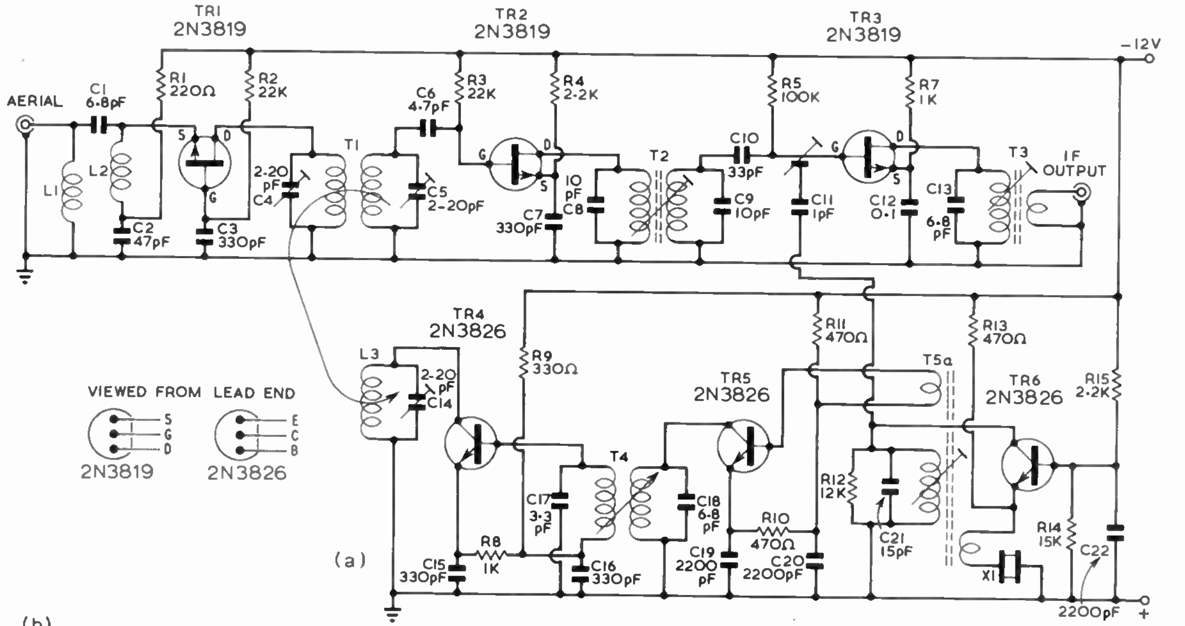
The converter shown in Fig. 5.50 follows the general form of the 70 and 144 Mc/s units described earlier and like them is inexpensive and easily put together. It nevertheless gives a very high performance and measurements show the overall noise factor to be in the region of 5db, the spread being from about 4 to 5·5db. Double frequency changing is employed within the converter to reduce second-channel effects to an absolute minimum.

The main reasons for using FET's, rather than ordinary bipolar transistors, in the r.f. and mixer stages of this converter are rather different from those which suggested their use in the 70 and 144 Mc/s converters. The 2N3819, as a common gate r.f. amplifier at 430 Mc/s, provides a satisfactory, stable, insertion gain of from 12 to 14db, together with an excellent noise performance. In the case of the mixer, the most useful feature of the FET is the high input impedance, in the region of 1 K ohm, attainable in common source at u.h.f., together with the good noise performance. The former characteristic eases the design of the local oscillator chain which, in this case, requires only three stages in all, using cheap silicon planar devices to provide adequate local oscillator injection. The excellent strong signal properties of the FET assist in preventing overload from nearby transmitters although this ability is not such a necessity on 70cm due to the lower density of activity as compared with the v.h.f. bands. A modification to the standard design is therefore included in Fig. 5.50(b) in which a bipolar transistor may be substituted for the FET in the second mixer. In this way, the total cost of the converter is reduced. However, the strong-signal performance is degraded to the extent that the front-end will only accept signals up to a few millivolts without exhibiting serious non-linearity whereas, using the FET, several 100 mV signals are acceptable. There is no other significant change in performance produced by this modification, apart from a small increase in the insertion gain. The second mixer stage is not protected by any selectivity to interfering in-band signals and yet has 20db or so of gain ahead of it. The modified design will handle most signals from distances of more than a mile or so, even with mutual beam alignment. However, if the full strong-signal performance is required, the all-FET design is recommended.

Circuit

An un-neutralized common-gate r.f. stage, using a 2N3819 (Texas Instruments) is inductively coupled into a common-source first mixer, also a 2N3819. It is followed by a further mixer employing a 2N3819, also in common-source or, in the alternative design, a common-emitter 2N3826 silicon planar, with local-oscillator injection at the crystal frequency itself.

Double tuned circuits, with mutual inductance coupling, are employed between the r.f. stage drain and first mixer gate. The aerial input is fed into the source of the common-gate r.f. stage via a circuit which is a double pi-coupler, with two



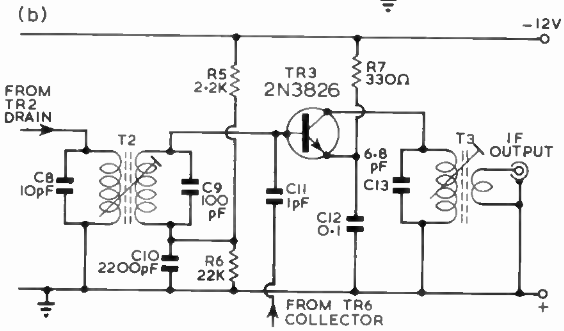


Fig. 5.50. (a) Converter for 70cm using three FETs; (b) an alternative second mixer employing a cheaper bipolar transistor.

shunt inductors and a series capacitor. This has the advantage that it tends to simulate a high-pass rather than a low-pass filter, at frequencies far outside the matching band, and this helps to reduce possible i.f. interference. The only slight disadvantage relative to the more normal pi-coupler is that mutual coupling between the two coils must be kept fairly low.

Double tuned circuits, at the first i.f. in the 36 Mc/s region, are used to feed the second mixer stage. The second i.f. extends from 2·0 to 4·0 Mc/s, for 432·0 to 434·0 Mc/s coverage, in the present design, using a crystal on or near to 33·077 Mc/s. The output i.f. may, of course, be put anywhere in the range up to about 15 Mc/s by appropriate choice of the crystal and by making modifications to the first (T2) and second (T3) i.f. transformers. The changes to be made to T2 are relatively minor and are easily estimated. Sample designs for T3, suitable for several different i.f.'s, may be found in the i.f. transformer table (Table 5.6). The formula for calculating the crystal frequency is:

$$\textit{Crystal frequency} = \frac{432{\cdot}0 - (\textit{lowest frequency of tunable i.f.})}{13}$$

For example, another suitable i.f. would be 3·0 to 5·0 Mc/s, requiring a 33·0 Mc/s crystal. The design for T2 would be essentially unaltered and that for T3 would be as follows:

Primary: 58 turns of 34 s.w.g. enamelled.
Secondary: 5 turns of 28 s.w.g. enamelled (for 75 ohms output).
13 turns of 28 s.w.g. enamelled (for 600 ohms output) in the pot-core, as before.

The local oscillator chain consists of three stages,

TABLE 5.13

Coil Details for the 430 Mc/s FET Converter

L1, 2 turns, 20 s.w.g. tinned copper, wound on ¼ in. mandrel, with $\frac{3}{16}$ in. long leads.
L2, As L1, but with ⅛ in. leads.
L3, As L1, but with one $\frac{3}{16}$ in. lead and one ⅛ in. lead.
T1, *Primary*: as L1, but with one ⅛ in. lead and one 1 in. lead; *Secondary*: as L1, but with one $\frac{3}{16}$ in. lead and one 2 in. lead. Primary and secondary co-axial, earthy ends near together with approx. ⅛ in. spacing.
(*Note: The underside photograph shows clearly the shapes and relative positions of the coils L1, L2, L3 and T1.*)
T2, *Primary*: 13 turns, 28 s.w.g. enamelled c/w on 0·3 in. Aladdin former; *Secondary*: 13 turns, 28 s.w.g. enamelled c/w (2N3819 second mixer); 4 turns 28 s.w.g. enam. c/w (2N3826 second mixer).
Both coils wound initially upon ¼ in. drill and allowed to spring off. Earthy ends near together on former, separated by ⅛ in. Secondary near to chassis end of former.
T3, *Primary*: 82 turns 36 s.w.g. enam., 41 turns wound in two adjacent sections of pot-core former (2·0 to 4·0 Mc/s i.f.); *Secondary*: 6 turns 28 s.w.g. (for 75 ohm i.f. output impedance) or 16 turns 28 s.w.g. (for 600 ohm i.f. output impedance), in remaining section.
T4, *Primary*: 5 turns 22 s.w.g. enam. c/w on 0·3 in. Aladdin former; *Secondary*: 5 turns 22 s.w.g. c/w on 0·3 in. Aladdin former. Coils mounted with earthy ends near together, space $\frac{3}{16}$ in. Primary near to chassis end of former. Both coils wound initially on ¼ in. drill and allowed to spring off.
T5, *Main*: 11 turns 28 s.w.g. enam., close wound on 0·3 in. Aladdin former; *Emitter*: 1 turn 28 s.w.g. enam., overwound at "cold" end of main winding; *Output*: 2 turns 28 s.w.g. enam., overwound at centre of main winding.

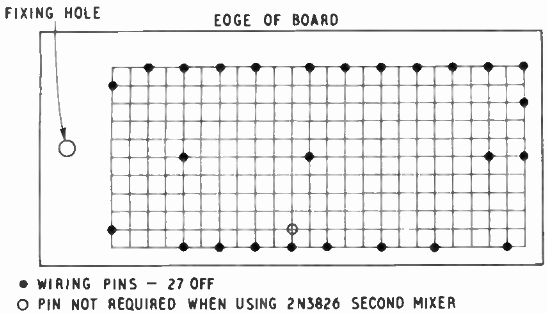


Fig. 5.51. Wiring pin arrangement on the Lektrokit circuit board.

employing 2N3826s, the first as an overtone crystal oscillator in the 33 Mc/s region. The second triples to 99 Mc/s, with double tuned interstage coupling circuits; the final stage quadrulples to about 396 Mc/s and is coupled into the first mixer by the mutual inductance between L3 and the secondary of T1.

Construction

Like the two v.h.f. converters, the 430 Mc/s unit uses a piece of Lektrokit board mounted inside a small, 20 s.w.g. aluminium U-channel chassis. The chassis for all three converters are identical but, for the 430 Mc/s unit, a shorter piece of Lektrokit board is used than for the other two. It is spaced about $\frac{5}{16}$ in. away from the chassis, with one end fixed as in the v.h.f. designs the other being secured by soldering a wiring pin to an earthed tag bolted to the chassis. Wiring pins in the board serve to support most components but those operating at signal or first oscillator frequency are mounted either upon ceramic pillars or on small tubular trimmers. A small screen between the input and output circuits of the r.f. stage ensures good stability. The single fixing hole in the screen tab should be marked and drilled only after the two holes have been drilled in the main flange, so allowing the screen to be fitted.

Apart from these details, the construction of the 70cm unit is similar to that for the 2m and 4m converters. The aerial input and i.f. output terminations may be either internal or external, using Belling-Lee coaxial sockets.

The coils associated with L3 and T1 are tuned with tubular trimmers. The type specified was chosen for its mechanical suitability but it has an unmodified capacitance range of from 2 to 20 pF. This minimum capacity is too high and should be reduced by removing about $\frac{3}{16}$ in. of silvering completely from the normally clamped end of the trimmer using a file and a piece of medium grade emery cloth.

The coils for L3 and T1 are mounted directly upon their respective trimmers but their earthy ends are returned to the stage common-earthing point (**Fig. 5.53**). The gate lead of TR1 is shown, for clarity, connected directly to the appropriate stand-off. A shorter gate-earth lead results however if C3 is grounded by a very short wire and the gate of TR1 is connected as close as possible to C3 itself. The mixer, TR2, source lead should also be kept short. The bypassing of TR4 is improved if C15 and C16 are taken to

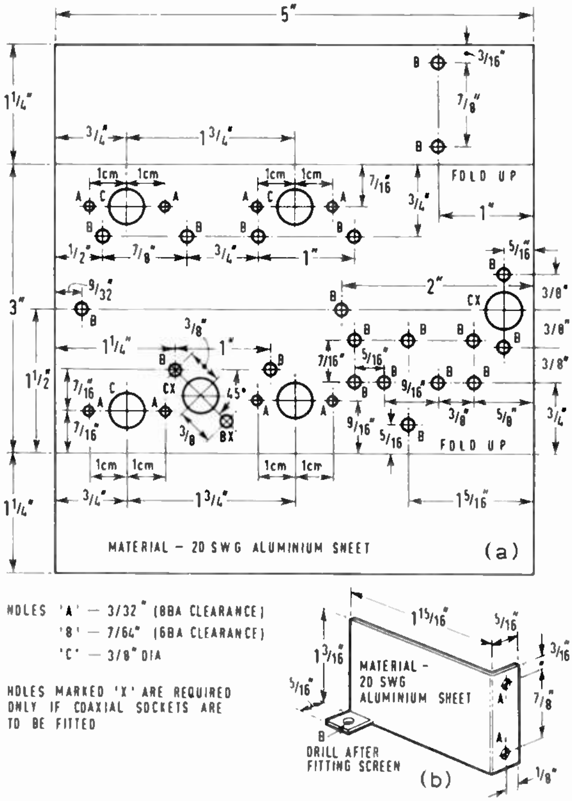


Fig. 5.52. Dimensions of the chassis and screen for the input stage.

TABLE 5.14

Components List for the 430 Mc/s FET Converter

C1, 13, 18, 6·8 pF ceramicon.
C2, 47 pF ceramicon.
C3, 7, 15, 16, 330 pF disc ceramic.
C4, 5, 14, 2 to 20 pF tubular ceramic trimmers.
C6, 4·7 pF ceramicon.
C8, 10 pF ceramicon.
C9, 10 pF ceramicon (FET mixer), 100 pF ceramicon (bipolar mixer).
C10, 33 pF ceramicon (FET mixer), 2200 pF disc ceramic (bipolar mixer).
C11, 1·0 pF ceramicon.
C12, 0·1 μF 30V disc ceramic.
C17, 3·3 pF ceramicon.
C19, 20, 22, 2200 pF disc ceramic.
C21, 15 pF ceramicon.
R1, 220 ohms.
R2, 3, 22 K ohms.
R4, 15, 2·2 K ohms.
R5, 100 K ohms (FET mixer), 2·2 K ohms (bipolar mixer).
R6, not required (FET mixer), 22 K ohms (bipolar mixer).
R7, 1 K ohm (FET mixer), 330 ohms (bipolar mixer).
R8, 1 K ohm
R9, 330 ohms.
R10, 11, 13, 470 ohms.
R12, 12 K ohms.
R14, 15 K ohms.
T2, 4, 5, 0·3 in. diam. nylon Aladdin former, with iron dust core.
T3, Denco Neosid pot-core.
TR1, 2, 2N3819 FET (Texas Instruments Ltd.).
TR3, 2N3819 FET or 2N3826 bipolar (see text).
TR4, 5, 6, 2N3826 bipolar silicon planar transistor (Texas Instruments Ltd.).
X1, HC-6/U crystal (and holder) or HC-18/U crystal, as required.
6 Oxley Type SO1 (6B.A.) ceramic stand-off pillars.
Lektrokit S.R.B.P. board and wiring pins, as required.
Belling-Lee coaxial sockets, if required.

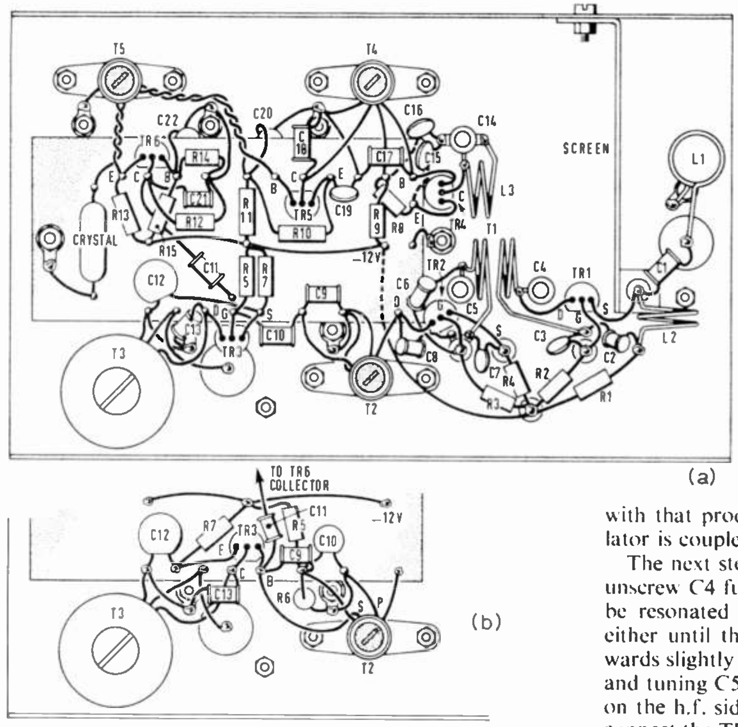


Fig. 5.53. Component layout for the standard converter (a) and alternative mixer (b).

their respective pins on the underside of the wiring board, before the latter is fitted. The negative supply wire to the TR1-TR2 stand-off may also be connected underneath the board in this way.

If an HC-6/U crystal is to be used, the crystal holder should have its lugs bent outwards and then soldered to the wiring pins provided. However, a miniature HC-18/U crystal should be fitted by passing the wires through a convenient pair of holes and then bringing the wires up through the board again for soldering to the pins. Very loose capacitative coupling is required between the crystal oscillator collector and the second mixer gate. This is accomplished by utilizing the capacitance between two adjacent pins on the Lektrokit board in series with the 1 pF ceramicon C11. When the 2N3826 mixer is employed, C11 alone provides the correct oscillator injection level at the mixer base.

TABLE 5.15

Measured Operating Conditions for 430 Mc/s Converter

		Emitter or Source	Base or Gate	
TR1	R.F. Amplifier	0·5 to 1·5	—	
TR2	First Mixer	0·7 to 2·0	—	
TR3	Second Mixer	0·7 to 2·0 0·8	— 1·1	(2N3819 FET) (2N3826)
TR4	Second Mult.	1·2 to 1·8	—	
TR5	First Mult.	1·5 to 2·5	—	
TR6	Crystal Osc.	0·9 to 1·1	1·5	

All voltages are measured with respect to the —12 volt supply line

Aligning the Converter

The local oscillator chain should be aligned first. An r.f. detector and a set of absorption wavemeters are very helpful during this procedure. However, it is almost impossible to find the wrong harmonics using the coils and capacitors suggested. When the oscillator chain is properly adjusted, tune the secondary of T2 until the drain current of TR3 increases slightly, as evidenced by a rise in d.c. voltage across R7. Then detune T2 secondary on the h.f. side until a small noise peak is heard. Next the slug in the primary of T2 may be adjusted to produce a further noise peak, due to the first mixer stage. A check with an absorption wavemeter will be worthwhile at this stage to ensure that T2 is tuned within the 35 to 37 Mc/s range. It will be noticed that the shot-noise output of the FET mixer stage is very low compared with that produced by bipolar mixers, until the local oscillator is coupled in and the input circuit tuned to resonance.

The next step is to disconnect the drain lead of TR1 and unscrew C4 fully. C5 and the secondary of T1 should then be resonated at the oscillator frequency by adjusting C5 either until the d.c. voltage measured across R4 kicks upwards slightly or by coupling an r.f. indicator loosely into L3 and tuning C5 for a dip in indication. Detuning C5 slightly on the h.f. side should give rise to a small noise peak. Reconnect the TR1 drain lead and, with the aerial plugged into the input socket, increase the setting of C4 until a further noise peak is heard, or until local signals are peaked up. The primary of T1 will be found to be more sharply tuned than the secondary. In fact, the over-all bandwidth of T1 may be controlled by adjusting the tightness of coupling between the two coils. The coil spacing should be changed as necessary, retuning C4 and C5, until the response of T1 is fairly flat over a band of from 4 to 6 Mc/s wide, centred at about 433 Mc/s. Insufficient coupling can even result in self-oscillation of the r.f. stage, when using some 2N3819s. It will not be found necessary to alter the component values in the input pi-coupler, but if a noise generator is available, C1 may be temporarily replaced by a 2 to 8 pF Mullard trimmer and L1 and L2 adjusted by bending them to optimize the input coupling circuits for the 2N3819 in use. Increasing the value of C1 will probably be found to reduce the loaded *Q* of the T1 primary, to a certain extent.

A MAST-HEAD PRE-AMPLIFIER FOR 430 MC/S

This unit was originally designed for inclusion in an amateur television system but should prove of interest to all operators concerned with the 70cm band.

There is an inevitable loss of signal strength and degradation of signal-to-noise ratio by even the best quality feeder, and this cannot be recovered by anything which may be done at the receiver. With the availability of transistors capable of both gain and noise figures superior to valves, a mast-head amplifier becomes a practical proposition.

Two feeders are employed, one for transmission and the other, permanently connected to the output of the pre-amplifier, for reception. The latter feeder may be relatively

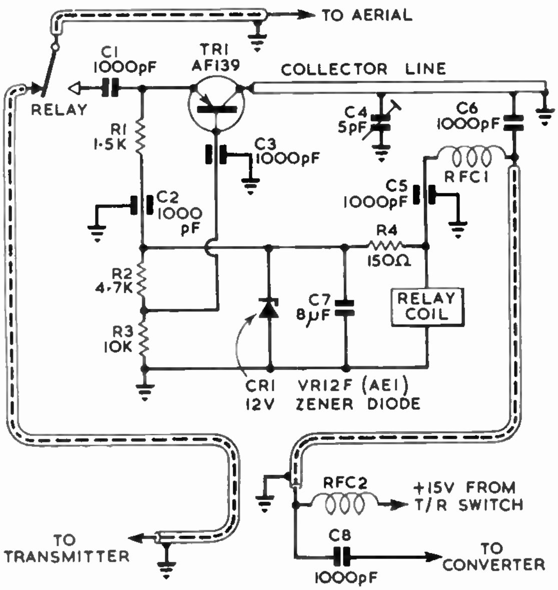


Fig. 5.54. Circuit of the mast-head pre-amplifier.

inexpensive cellular polythene television down-lead without sacrificing signal strength or the ratio of signal-to-noise.

Fig. 5.54 shows the circuit diagram while Fig. 5.55 details the construction of the major portion of the pre-amplifier which comprises a AF139 transistor in a trough line. At the ultra high frequencies layout becomes all important and constructors are urged to follow the illustration as faithfully as possible; if this is done no particular difficulties should be encountered. Only one item requires special mention and that is the collector line tuning capacitor (C4) which is a sub-miniature ceramic barrel trimmer.

Fig. 5.56 illustrates how the pre-amplifier, change-over relay and associated cabling together with the balance of the components are disposed within a polythene box obtainable from Woolworth's or other multiple stores. This box serves not only to house the components but also to protect them from the weather.

It is essential that a coaxial relay should be employed for the switching function. The cost of such an item is normally high, but in the unit described use was made of a surplus relay type 78A which was taken apart and rebuilt to give just the contact sequence required and has been found quite satisfactory. The relay is de-energized on transmit and in this position the transmitter is connected directly to the aerial system via its own feeder. The reason for energizing the relay in the receive position is to ensure that, in the event of a fault developing in the amplifier or in the power supply wiring, the station will not be put off the air. Under such conditions the transmitter coaxial cable becomes the sole

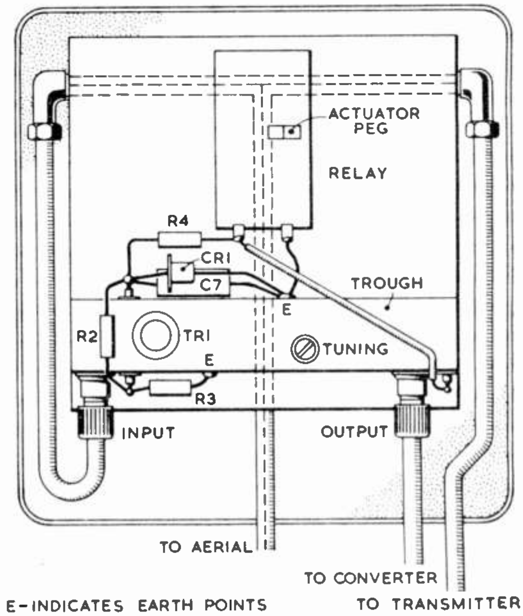


Fig. 5.56. Suitable positions for the transistor amplifier and relay within the polythene box which is fitted to the rear of the aerial reflector. The length of the coaxial lead between the relay and the input of the amplifier is critical—see text.

connection to the aerial, and with suitable send-receive switching at the station end of the feeder, normal operation may be continued. On reception a 15 volt positive supply is fed up to the pre-amplifier via the " receive " coaxial cable to operate the change-over relay. This supply, regulated by a zener diode and decoupled by a suitable capacitor, is also used to power the pre-amplifier. The main requirement of this feed system is that the r.f. chokes at each end of the line shall be highly efficient at the operating frequency. Poor chokes might well cause appreciable loss of signal.

R.F. leakage across the relay while transmitting could cause damage to the transistor, but by careful adjustment of the length of the coaxial cable between the relay and the input socket the current flowing in the emitter-base junction of the transistor from this cause may be reduced to safe proportions. In the original unit this was found to be less than one microamp, when the length of cable was 4 inches. It is strongly advised that such a measurement be carried out starting with low input to the transmitter to ensure that no damage will be caused when using higher power.

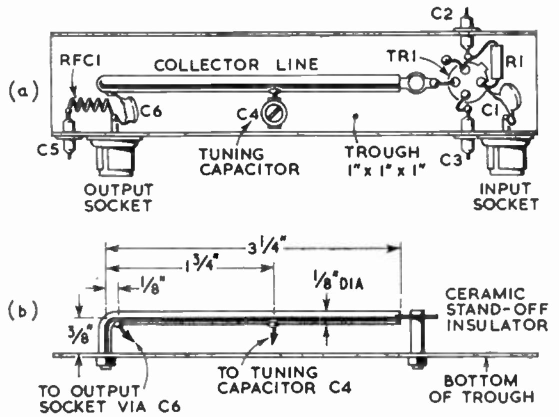


Fig. 5.55. Constructional details of the pre-amplifier chassis.

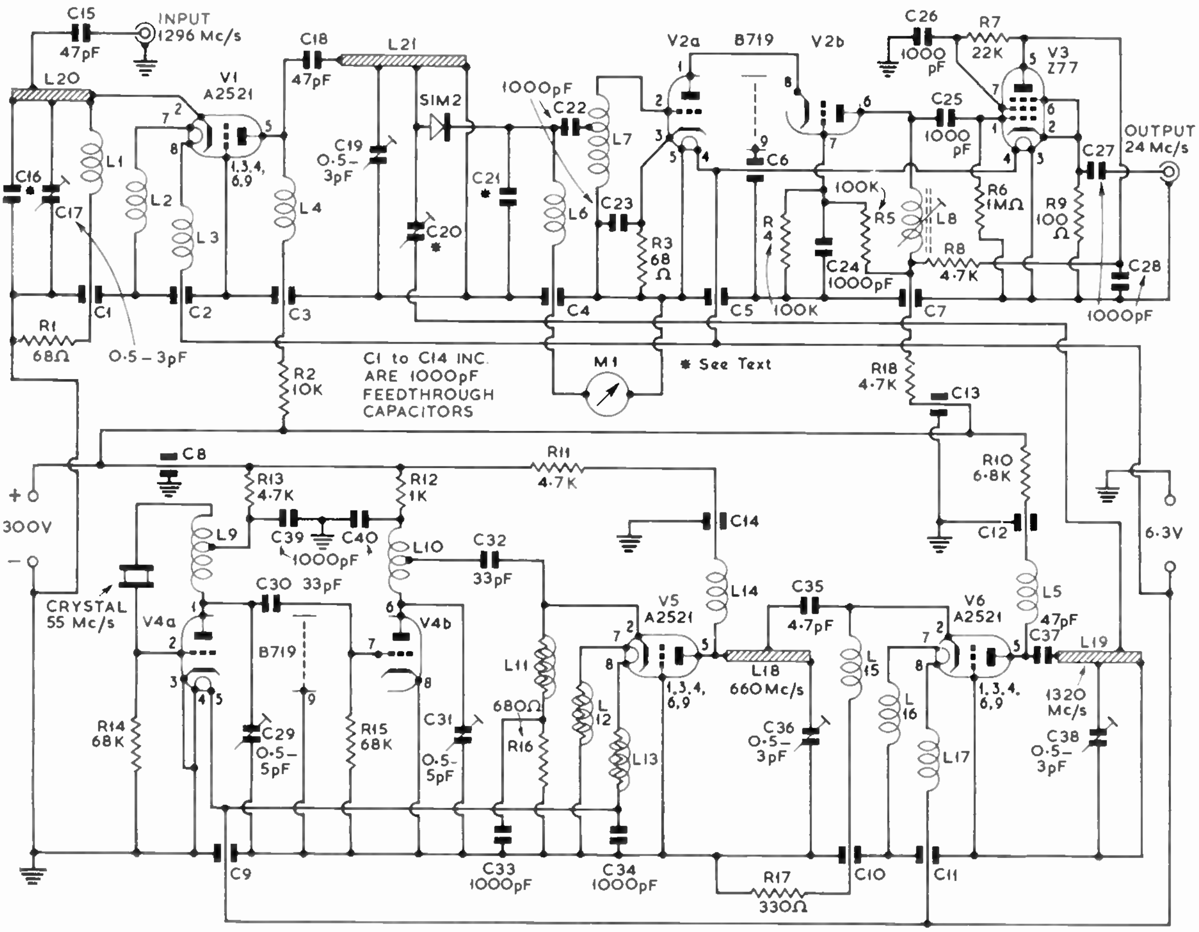


Fig. 5.57. Circuit diagram of the crystal-controlled 1300 Mc/s converter. Details of the inductors are given in table 5.16.

A NARROW BAND CONVERTER FOR 1300 MC/S

This converter employs the A.2521/6CR4 grounded grid triode both as a signal frequency amplifier and in the frequency multiplier chain for the local oscillator. A feature of the design is the simple trough-line circuits in place of the more expensive and difficult to construct coaxial types normally used at this frequency. A crystal mixer is employed and is followed by a series cascode stage as the first i.f. This is intended to be followed by a standard communications receiver providing a tunable i.f. over the bandwidth of the converter.

The complete circuit is shown in Fig. 5.57, the measured specification of the unit being as follows:

Centre signal frequency	1296 Mc/s
R.F. gain before mixer	> 13db
R.F. bandwidth	5 Mc/s
Overall noise factor	11·5db
Local oscillator injection frequency	1320 Mc/s
Image rejection	> 45db
Output frequency	24 Mc/s

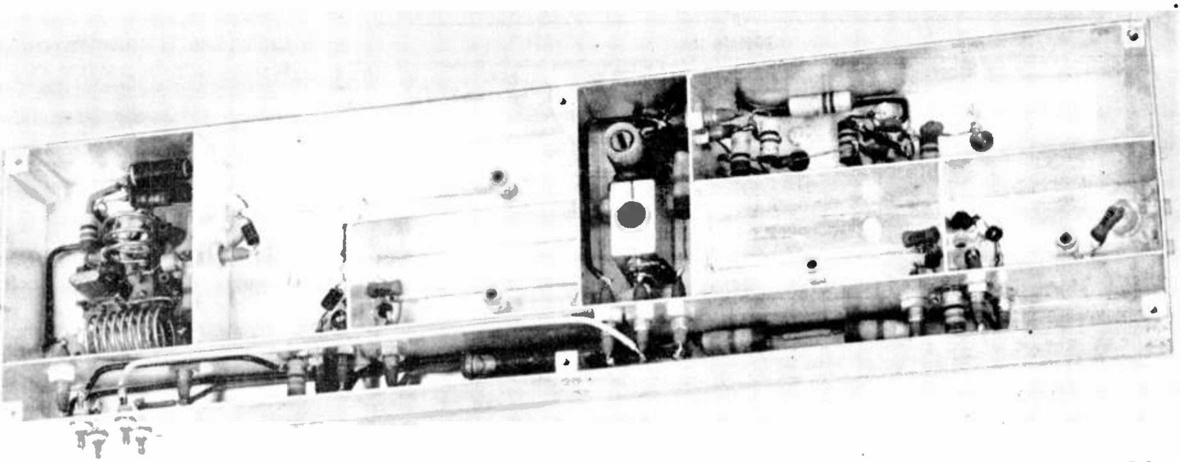
Circuit

The r.f. stage comprises a grounded grid A.2521 troughline amplifier with tuned input and output lines. The input signal is fed into the converter by means of a type BNC socket and the aerial impedance is matched by varying the position of the input tap along the cathode line, and by tuning the line.

The cathode line is a three-quarter wavelength line short-circuited at its extreme end by a capacitor of approximately 50 pF, and tuned by a 0·5 to 3·0 pF Mullard trimmer mounted halfway along the line. The anode line is a three-quarter wavelength line short-circuited and tuned by a 0·5 to 3·0 pF Mullard trimmer, again mounted halfway along the line. This is isolated from the h.t. by a 47 pF capacitor. The output is tapped at the mechanical short-circuit (the electrical r.f. short-circuit being somewhere in the chassis), and fed direct into the crystal mixer.

Crystal Oscillator Multiplier Chain

The crystal oscillator is one half of a ECC85/6AQ8 valve connected in a Squier circuit and controlled by a 55 Mc/s crystal. The second half of this valve is tuned to the third



A view under the chassis of the 1300 Mc/s narrow band crystal-controlled converter.

harmonic of the crystal frequency, thus giving 165 Mc/s output which is fed into a grounded grid A.2521 quadrupler operating in class C, the output circuit being a half wave-length line tuned at its extreme end by a 0·5 to 3·0 pF Mullard trimmer. The output from this stage at 660 Mc/s is tapped off about halfway along the line (this should be adjusted for maximum crystal current) and fed into a grounded grid A.2521 doubler also operating in class C. The anode line of this stage is a three quarter wavelength line short-circuited to the chassis at its extreme end and tuned by a 0·5 to 3·0 pF Mullard trimmer, h.t. isolation being provided by a 47 pF capacitor at the anode pin. The Mullard trimmers are type C004EA/3E. Alternatively, Erie type 3115A may be used.

The output is tapped off at the mechanical short-circuit and fed into a fabricated variable capacitor which couples the output into the crystal mixer.

Mixer and I.F. Stages

The mixer is a SIM2 low-noise diode mounted in a silver-plated brass holder, r.f. decoupling being provided by a 50 pF capacitor formed at the base of the crystal holder. The diode current is 0·3 mA. The output at 24 Mc/s is fed into a tapped input coil, on which the tap should be adjusted for best noise performance; this generally occurs about halfway along the coil. The construction of a holder for the mixer crystal is shown in Fig. 5.58.

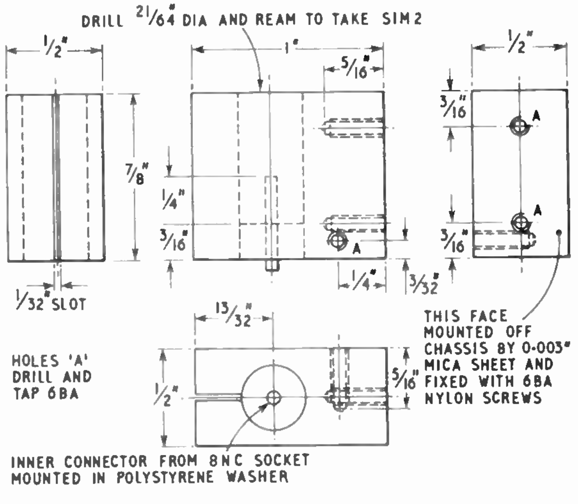


Fig. 5.58. Construction of a suitable holder for the crystal mixer. This is of a slightly different pattern from that shewn in the photograph of the protctype.

TABLE 5.16

C16	≃ 50 pF fabricated with 1·5cm square brass plate separated from chassis by 0·003 in. mica sheet.
C20	≃ 1 pF fabricated by varying spacing between crystal connector and oscillator output tag.
C21	≃ 50 pF fabricated with 7/8 in. × 1/2 in. brass crystal holder separated from chassis by 0·003 in. mica sheet (See Fig. 5.58).
C29, C31	0·5 to 5 pF vane type trimmer.
C33, C34, C39, C40	1000 pF stand-off capacitors.
L1, L2, L3, L4, L5	8cm of 24 s.w.g. enamelled copper wire wound into 1/8 in. diam. air-cored coil.
L6	30 μH choke, Painton type 200154.
L7	30 turns 30 s.w.g. enamelled copper wire, wound on 3/8 in. diam. former, tapped at centre and tuned with ferrite slug.
L8	17 turns 36 s.w.g. enamelled copper wire, wound on 1/4 in. diam. former and tuned with ferrite slug.
L9	9 turns 18 s.w.g. tinned copper wire, wound into 1/2 in. diam. open spaced, air-cored coil, tapped 7 turns from anode end.
L10	3 turns 18 s.w.g. tinned copper wire, wound into 1/2 in. diam. open spaced, air-cored coil, tapped 2 turns from anode end.
L11, L12, L13	30 s.w.g. enamelled copper wire close wound onto 100 K ohms 1/2 watt resistor.
L14, L15, L16, L17	16cm of 24 s.w.g. enamelled copper wire wound into 1/8 in. diam. air-cored coil.
L18	7·5cm of 1/8 in. diam. silver-plated brass rod mounted 1/4 in. above chassis, tuned at extreme end, tapped 3·5cm from open circuit.
L19, L21	6cm of 1/8 in. diam. silver-plated brass rod mounted 1/4 in. above chassis, tuned at 3cm from short-circuit, tapped at short-circuit.
L20	5cm of 1/8 in. diam. silver-plated brass rod mounted 1/4 in. above chassis, tuned 2·5cm from short-circuit, tapped 2cm from short-circuit.

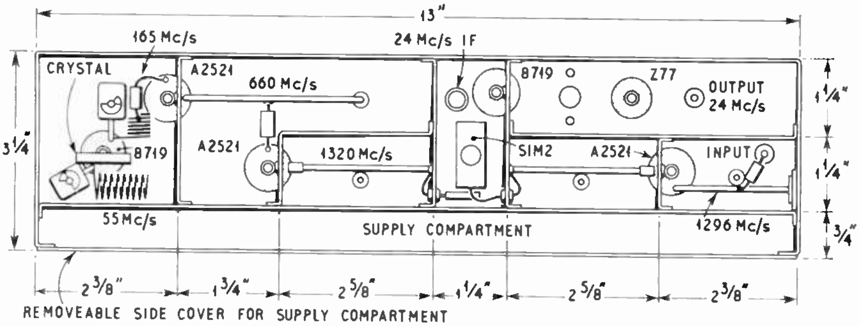


Fig. 5.59. Layout of the principal components in the 1300 Mc/s narrow band converter. The chassis is 1¼ in. deep and fabricated from 20 s.w.g. silver-plated brass sheet and divided into compartments with the shields soldered in position.

The mixer is followed by a B.719 connected in series cascode tuned to 24 Mc/s and followed by a Z77 cathode follower buffer stage.

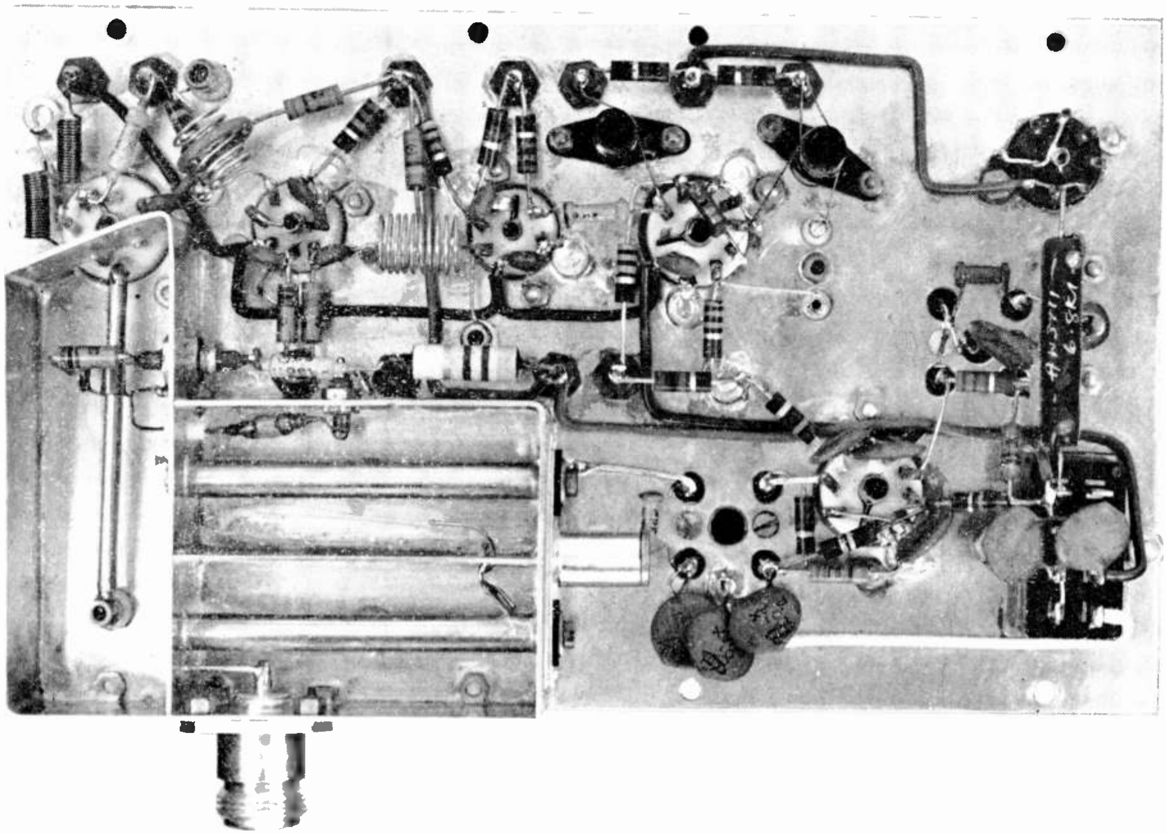
Construction

The chassis is a shallow box 13 in. long by 3¼ in. wide and 1¼ in. deep, constructed of 20 s.w.g. silver-plated brass sheet. It is divided into compartments as shown in **Fig. 5.59** and the walls of the compartments are solidly soldered to the main chassis. One wall of the chassis is made removable to facilitate the making of the h.t. connections. The transverse screens across the valveholders must be made very close fitting and the grid pins solidly earthed to the screen in all cases. A wiring jig or a scrap B9A valve should be plugged into the holder while this is being done to avoid the contacts being pulled out of alignment.

The valveholders are all made of p.t.f.e. and the connections from the u.h.f. stages into the mixer are made by means of p.t.f.e. feed-through connectors. While the output socket may be a normal television coaxial type, the input socket should be a special u.h.f. type. BNC connectors were used in the original, the socket being type UG447/U and the plug type UG88c/U. These are obtainable from Greenpar Engineering Ltd., Cambridge Road, Harlow, Essex.

Details of the components are given in **Table 5.16**.

All power supply filter resistors and feedthrough capacitors are mounted in the power supply compartment. The converter requires an h.t. supply of 300 volts at 67 mA.



Under-chassis view of the experimental 13 cm. converter shown in Fig 5.60.

AN EXPERIMENTAL 13CM (2300 MC/S) CONVERTER

In the experimental 13cm crystal-controlled converter shown in **Fig. 5.60**, signals at 2300 Mc/s are mixed with the output of a crystal oscillator chain at 2280 Mc/s, the i.f. produced being amplified by a wideband head amplifier covering 20–30 Mc/s. The upper half of Fig. 5.60 shows the i.f. head amplifier and the trough-line assembly which comprises the aerial and mixer circuits and the final stages of the multiplier chain. The lower half shows the crystal oscillator and the multiplier chain to 253 Mc/s. Component details are given in **Table 5.17**.

Signals at 2300 Mc/s applied via the aerial connector and the input loop L1 are tuned by L2, C12 and coupled to the mixer diode CR1 by L3. The 2280 Mc/s output of the multiplier diode CR2 is also coupled to CR1 via L3. Intermediate frequency signals produced by CR1 are fed to the i.f. head amplifier V1. This is a cascode stage, the operating conditions of which are stabilized by R3 and R6. Wideband transformers T2 and T1, which cover 20–30 Mc/s, respec-

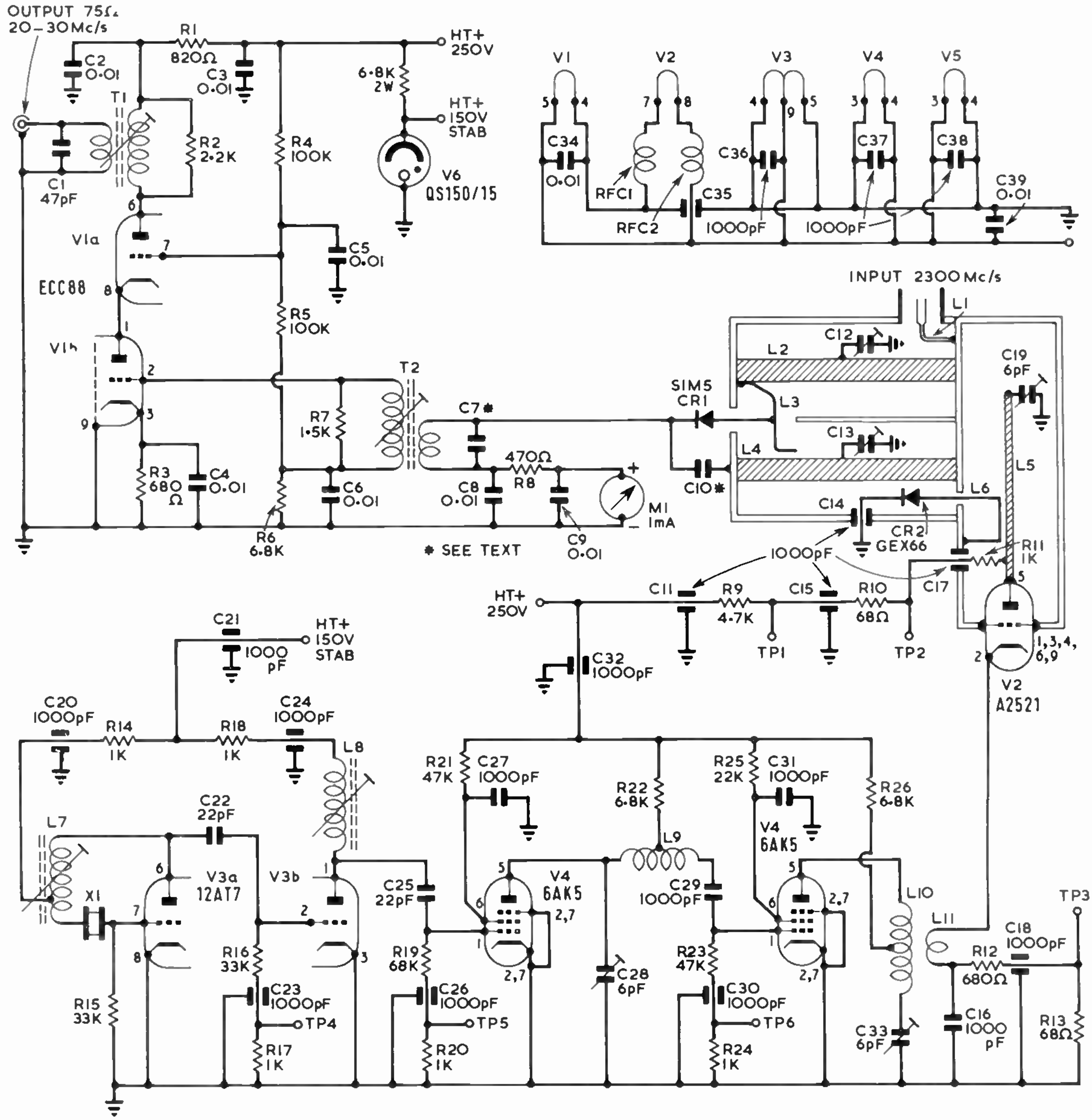


Fig. 5.60. Circuit diagram of the 13cm. converter.

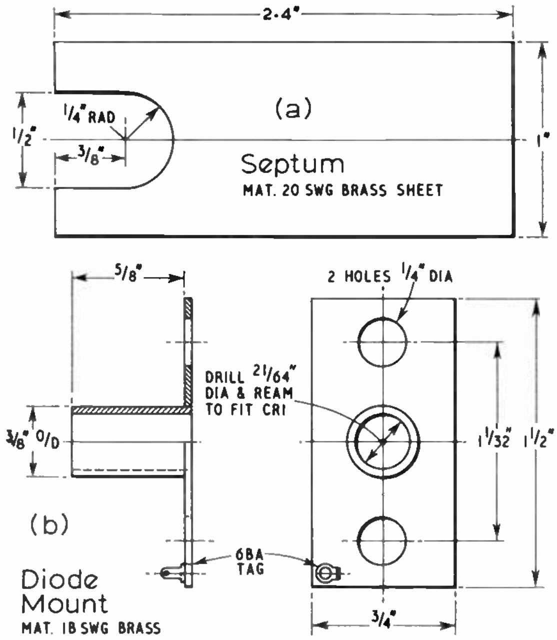


Fig. 5.61. Constructional details of septum and diode mount.

TABLE 5.18

Inductor details for the 2300 Mc/s Converter.

	Tuned to	Winding
L1	—	½ in. 18 s.w.g. tinned copper wire spaced ⅛ in. from L2.
L2	2305 Mc/s	2·40 in. ¼ in. diam. brass rod, ends tapped 6 B.A. Mounted ⅜ in. clear of chassis.
L3	—	see Fig. 5.70.
L4	2280 Mc/s	as L2.
L5	760 Mc/s	2·2 in. ⅛ in. diam. brass rod, mounted ¼ in. clear from chassis, tapped approximately ⅜ in. from anode end.
L6	—	Formed from wire end of CR2. See Fig. 5.71.
L7	31·7 Mc/s	25 turns 30 s.w.g. enam. copper wire close wound on 0·27 in. diam. former.
L8	63 Mc/s	8 turns 30 s.w.g. enam. copper wire close wound on 0·27 in. diam. former.
L9	126 Mc/s	10 turns 22 s.w.g. tinned copper wire wound on 5/16 in. mandrel air-spaced to 9/16 in., centre-tapped.
L10	253 Mc/s	6 turns ditto 7/16 in. long, centre-tapped.
L11	—	1 turn 22 s.w.g. p.v.c. covered wire wound over the centre of L10.
RFC1, RFC2		10 in. 24 s.w.g. enam. copper wire on ⅛ in. diam. mandrel, air-spaced.
T1	20–30 Mc/s	primary: 7½ turns 34 s.w.g. enam. copper wire close wound on 0·3 in. diam. former, 1·0 μH. secondary: 15 turns 34 s.w.g. pile wound to occupy 5/32 in. length, wound adjacent to primary, 2·8 μH.
T2	20–30 Mc/s	primary: 18 turns 40 s.w.g. enam. copper wire close wound on 0·3 in. diam. former 3·9 μH. secondary: 4½ turns ditto immediately adjacent to primary 0·43 μH.

tively match the mixer diode CR1 to V1, and provide a 75 ohm output to the main receiver.

The oscillator and earlier stages of the multiplier chain are shown in the lower part of Fig. 5.60. V3(a) is used as a Squier oscillator, X1 being a 31·6667 Mc/s third overtone crystal. Stages V3(b), V4 and V5 successively double to 253 Mc/s and the output is fed via L10 and L11 to the trough-line assembly. This assembly incorporates two further stages of multiplication, firstly a grounded-grid triode V2 tripling to 760 Mc/s, and secondly a diode multiplier CR2, the 2280 Mc/s output of which is tuned by L4, C13. The positive end of CR2 is decoupled through C14 and earthed with a link which may be removed for metering during aligning. Both V3(a) and V3(b) are fed from the 150 volt stabilized supply obtained from V6.

The general layout of the converter is shown in the photograph. The troughs, 1 in. wide internally and 1 in. deep, are fabricated from 20 s.w.g. brass sheet. Details of the septum and diode mount are given in **Fig. 5.61.** The trough components are temporarily jigged together using chromium-plated instrument screws (which do not solder) screwed into the tapped ends of L2 and L4, and are located by a scrap valveholder in position V2. The assembly may then be soldered directly to the 18 s.w.g. brass chassis using a small gas flame. 0 B.A. brass nuts soldered to the top of the chassis act as threaded bearings for C12 and C13 while 6 B.A. solder tags are soldered directly to the chassis to provide convenient earthing points, and in particular short earthing paths for the cathodes of V4 and V5 (pins 2 and 7). In connecting the valveholder for V2, pins 3, 4 and 6 and the

TABLE 5.17

Component Details

C2, C3, C4, C5, C6, C8, C9, C34, C39	0·01 μF disc ceramics.
C7	value up to 10 pF to pad C10 to 51 pF.
C10	capacitor formed between diode mount and trough wall. Dielectric polythene approximately 0·010 in. thick selected to give capacity 40–51 pF.
C11, C14, C15, C17, C18, C20, C21, C23, C24, C26, C30, C32, C35	1000 pF feedthrough.
C12, C13	capacitor formed between 0 B.A. brass bolts and L2, L4.
C16, C27, C29, C31, C36, C37, C38	1000 pF disc ceramics.
C19, C28, C33	ceramic tubular trimmers 0·5–5 pF Erie 3116A.
CR1	SIM5 or CV2155 (SIM2 and CV2154 similar but reversed polarity).
CR2	GEX66.
R10, R13	metering resistors 10–100 ohms matched ±2 per cent.

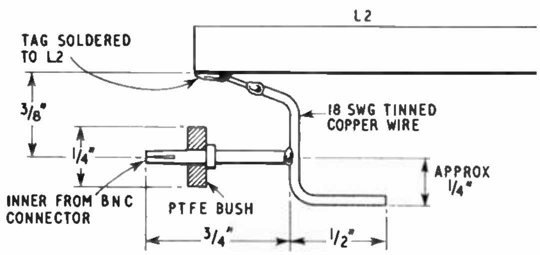
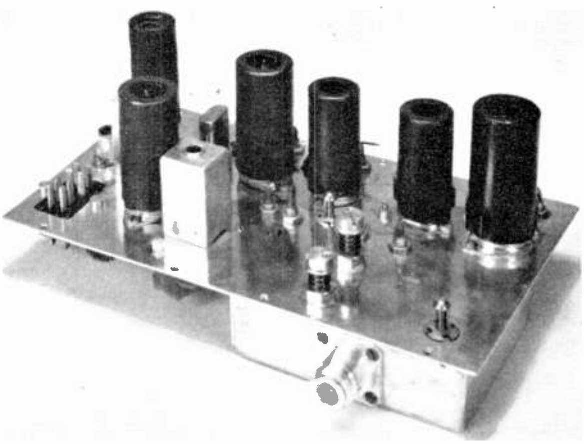


Fig. 5.62. Connector to mixer diode CR1 using inner of BNC socket.



Top view of the converter.

centre screen are soldered directly to the trough wall, and pins 1 and 9 to the solder-tag of the centre-screen of the valve holder. P.T.F.E. valveholders are used for all valves except the stabilizer V6.

The diode mount is held firmly against the trough wall by 6 B.A. bolts screwed into the ends of L2 and L4, the bolts

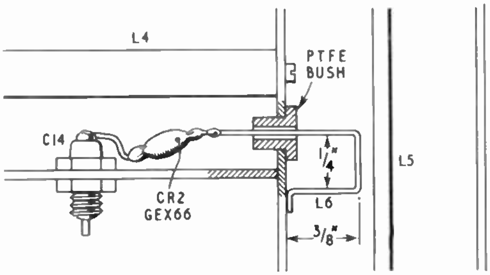


Fig. 5.63. Method of connecting diode multiplier CR2.

being insulated by p.t.f.e. or fibre bushes. Between the mount and the trough is fitted a polythene sheet the thickness of which, approximately 0·010 in., is chosen to produce a capacitor having a value of 40–51 pF. This capacitor, C10, which is both the r.f. bypass at 2300 Mc/s and also the tuning capacitor for the primary of T2, is padded to 51 pF by C7. The connector to the mixer diode CR1 is the inner of a BNC socket, the p.t.f.e. insulation from which also is used in locating the inner within CR2 see **Fig. 5.62**. The adjustable tap on L5 consists of a connector removed from a paxolin octal valveholder.

Alignment of the converter should present few problems provided that the critical dimensions of the trough-circuitry

TABLE 5.19

Meter Between		Current
CR2	chassis	40 mA
CR1	chassis	0·6 mA
TP1	TP2	11·3 mA
TP3	chassis	12·3 mA
TP4	chassis	−0·9 mA
TP5	chassis	−0·5 mA
TP6	chassis	−1·3 mA

i.e. the lengths of L2, L4 and L5 are within $\pm\frac{1}{32}$ in. With the valves plugged in, the anode circuits of V3(a), V3(b), V4 and V5 are tuned to the frequencies given in Table 5.18 using a g.d.o. H.T. may then be applied and the oscillator and multiplier circuits peaked by metering test points, 4, 5, 6 and 3 in that order. The currents measured should be similar to those given in **Table 5.19**. The drive applied to the grid of V3 should be about 1 mA, and is the difference between the cathode and anode currents measured from the voltage drop across the matched resistors R13 and R10 respectively. C19, together with the tap on L5 and the coupling link to CR2 (L6) should then be adjusted to produce the maximum current in CR2 (up to a maximum of 40 mA). The tuning point of C13 is strongly dependent on the position of CR2: that shown in **Fig. 5.63** was found to be the optimum and resulted in the 0 B.A. tuning bolt being unscrewed $2\frac{1}{4}$ turns from being in contact with L4. The small capacitive coupling between L3 and L4 finally is adjusted to give a mixer current of 0·5–1 mA. Final peaking should be done after L2 has been tuned to 2305 Mc/s: an approximate position for C12 is $\frac{3}{4}$ turn from being in contact with L2.

The current through V1 should be set to 10 mA by altering R6, and T1 and T2 adjusted to give a uniform output over the range 20–30 Mc/s.

CHAPTER 6

H.F. TRANSMITTERS

THE purpose of a transmitter is to generate radio frequency power which may be keyed or modulated and thus employed to convey intelligence to one or more receiving stations. This chapter deals with the design and adjustment of that part of the transmitter which produces the r.f. signal, while the methods by which this signal may be keyed or modulated are described separately in other chapters. Transmitters operating on frequencies below 30 Mc/s only are discussed here. Methods of generating r.f. power at frequencies higher than 30 Mc/s are described in Chapter 7—*V.H.F./U.H.F. Transmitters.*

One of the most important requirements of any transmitter is that the desired frequency of transmission shall be stable within close limits to permit reception by a selective receiver, and to avoid interference with other amateurs using the same frequency band. Spurious frequency radiations, capable of causing interference with other services, including television and broadcasting, must also be avoided. These problems are considered in Chapter 18—*Interference.*

The simplest form of transmitter is a single stage self-excited oscillator coupled directly to an aerial system. Such an elementary arrangement has, however, three serious disadvantages:

(a) The limited power obtainable with adequate frequency stability;
(b) The possibility of spurious frequency radiation, and
(c) The difficulty of securing satisfactory modulation or keying characteristics.

In order to overcome these disadvantages, the oscillator must not be called upon to supply power to the following stage. Since frequency multipliers are essentially class C stages (i.e., they require driving power), and since power amplifiers are usually operated in class C these stages must, in the interests of efficiency, be isolated from the oscillator by a buffer stage, which is capable of supplying this power without loading the oscillator.

Since the h.f. amateur bands are harmonically related, the oscillator is generally designed to operate on the lowest frequency band. In order to achieve output on the higher frequencies, it is necessary to employ one or more stages of *frequency multiplication.* An advantage gained from this arrangement is that by operating the oscillator at a relatively low frequency, it is easier to achieve the necessary frequency stability. A single multiplier stage may produce twice the fundamental frequency—*doubler*, three times—*tripler*, or more. Usually, however, the frequency multiplication in one stage is limited to three or four, owing to the rapid diminution in output power as the order of frequency multiplication is increased. It is undesirable to feed the aerial directly from a multiplier stage since harmonics, other than the wanted signal, are produced and may be radiated simultaneously. Where attenuation of unwanted signals is of the utmost importance, the power amplifier must not be driven directly from a frequency multiplier stage. The coupling between one stage and the next may be accomplished in a variety of ways, some of which permit physical separation between the various units. Others have the advantages of simplicity, harmonic rejection and so forth. Methods of coupling are described later.

The power supply for the transmitter may be obtained from conventional circuits such as those described in Chapter 17—*Power Supplies.* It is common practice to use separate power supplies for the low and high power sections of the transmitter, first because the voltages used in the two parts often differ greatly and, secondly, because interaction between the various stages can make the adjustment more difficult when a common supply is used. Unwanted modulation effects may also be introduced. The basic principles of transmitter circuit arrangement are illustrated by the block diagram in **Fig. 6.1.**

THE CRYSTAL OSCILLATOR

The simplest method of achieving the high degree of frequency stability is by using a quartz crystal to control the frequency of the oscillator. Such an oscillator when correctly designed and adjusted remains the most frequency stable device available to the amateur.

The quartz crystal, cut in the form of a suitably dimensioned plate, behaves like a tuned circuit of exceptionally high Q-value and may therefore be connected in the oscillator circuit as would a normal frequency determining tuned circuit.

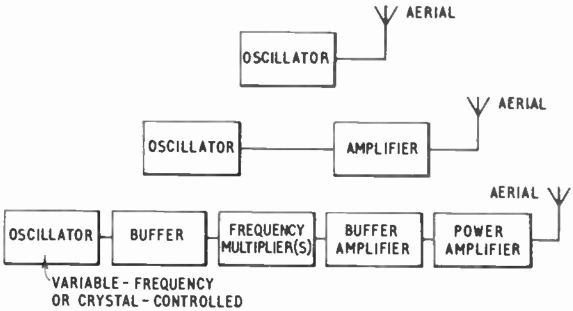


Fig. 6.1. Block diagrams of typical transmitters showing three possible combinations of oscillators, amplifiers and frequency multipliers. Many other arrangements are in common use, depending on the power output required and the number of frequency bands on which the transmitter is to operate.

The action of the quartz crystal depends on the piezo-electric effect explained in Chapter 1—*Principles*.

In the natural vibration of a quartz plate when used in such an oscillator circuit, a small but appreciable amount of heat is generated by the internal frictional effects. The temperature rise may cause the otherwise very stable frequency to drift from its nominal value. This frequency drift can be quite serious, especially when the fundamental frequency is multipled several times in the subsequent stages of the transmitter.

If the r.f. current through the crystal is allowed to become excessive, the vibrations of the quartz plate may become so large that fracture occurs.

In order to limit the r.f. current and the consequent temperature rise and to reduce the danger of crystal breakage, the power output of an oscillator must be kept at a relatively low value. Certain types of crystal have low temperature coefficients which reduce the amount of frequency drift likely to be encountered, but as these crystals are often very thin there is an increased possibility of damage. The power output of an oscillator using such a plate is therefore rather less than with other types of crystal. As described in Chapter 1, there are several types of crystal cuts, some of these, particularly the AT- and BT-cuts, being commonly used for h.f. transmitter control by reason of their low temperature coefficients. The X- and Y-cuts have relatively high coefficients. Some examples of the temperature coefficients for various crystal cuts are given in **Table 6.1.** Although the temperature coefficients of the AT- and BT-cuts are given in Table 6.1 as zero, this is true only at certain temperatures; at other temperatures the coefficient may have a small positive or negative value. Only the GT-cut has a true zero temperature coefficient over a wide range of temperature (0–100°C.) but its use is unfortunately restricted to low-frequency applications. For exceptionally high-precision operation over long periods, the crystal may be installed in a thermostatically controlled oven, although in amateur transmitters sufficient frequency stability can almost always be obtained by isolating the crystal from any part of the transmitter which becomes heated in normal operation, such as valves, resistors and power transformers.

The equivalent circuit of a crystal contains a resistance *R* in series with one of its branches. This represents the mechanical resistance to vibration and is related to the energy which must be supplied to the plate to maintain it in a state of vibration; a low value of resistance implies a high degree of "activity." If this resistance becomes too great, it will be difficult or impossible to maintain oscillation. The crystal is then said to be "sluggish" or inactive. The fault is occasionally found to be due to the contamination of the crystal surfaces. These surfaces must therefore be kept scrupulously clean and dry and without any trace of grease. When necessary crystals can be cleaned with carbon tetrachloride or even with soap and water.

TABLE 6.1

Characteristics of various cuts of quartz crystals.

Type of Cut	Normal Frequency Range (Mc/s)	Temperature Coefficient (cycles/Mc/s/°C)
X	1 – 5	−20
Y	1 –10	+75
AT	0·5– 8	0
BC	1 –20	−20
BT	1 –20	0
GT	0·1– 0·5	0

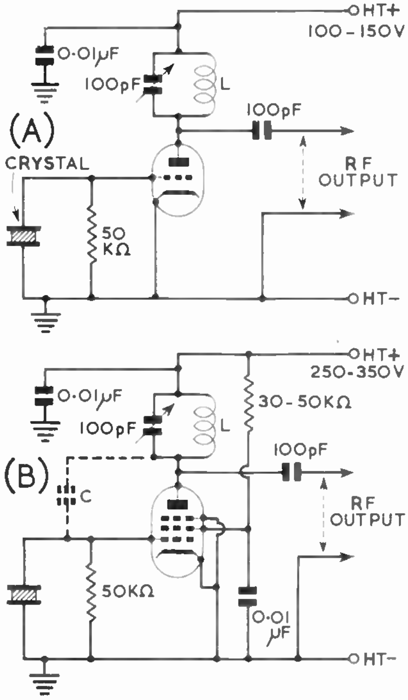


Fig. 6.2. Two examples of simple Pierce-Miller oscillator circuits. For operation on 3·5 Mc/s, L1 should have an inductance of approximately 23 μH; for 7 Mc/s, 7·5 μH.

Crystal Oscillator Circuits

Almost any small receiving triode, tetrode or pentode valve may be used in a crystal oscillator. The use of a small valve is recommended since the power output must be limited in order to reduce the possibility of overheating and crystal fracture.

The simplest possible circuits using a triode or pentode in Pierce-Miller arrangements are shown in **Fig. 6.2.** In these circuits the reactance of the anode circuit causes feedback through the anode-to-grid capacitance with sufficient amplitude and *in the correct phase* to maintain oscillation of the crystal.

As the anode tuning capacitance is reduced from its maximum value a point will be reached at which there is a sharp drop in the anode current. This denotes that the conditions for oscillation are being fulfilled and that the crystal is vibrating. At the lowest part of the "dip" the crystal will be vibrating at maximum amplitude. Continuing towards minimum capacitance, the feed current begins to rise again more gradually and the power output drops. When the oscillator is loaded, for example by the input resistance of a following stage, the dip in anode current is less marked than with the unloaded valve. **Fig. 6.3(a)** shows

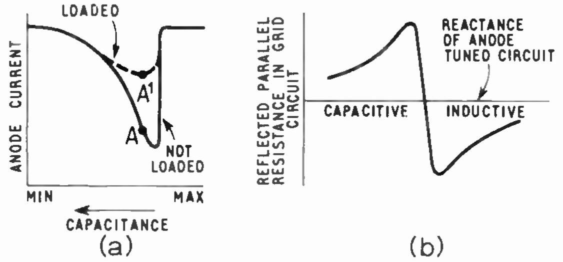


Fig. 6.3. Variation of feed current with tuning adjustment in a crystal oscillator. When the circuit is not loaded the dip in anode current is very pronounced, but it becomes less marked when a load is applied. For stable operation the tuning should be adjusted to correspond approximately to the point A (or A' when loaded).

how the feed current can be expected to vary with the tuning adjustment.

Oscillation is not, however, just a case of the magnitude of anode circuit impedance. Oscillation occurs in the vicinity of the resonant condition of the tuned circuit but only when it is inductively reactive at the operating frequency. The reflected impedance in the grid circuit, due to the Miller effect, then has a negative resistive component, i.e. power is fed back to sustain the losses of the grid circuit. When the anode circuit is capacitively reactive, the reflected impedance has a positive resistive component and oscillation is not possible. At the exact point of resonance the resistance is zero. This explains the sudden " flop " of the crystal oscillator as the anode circuit is tuned from an inductively reactive condition to a capacitively reactive one. The reflected resistive component is shown in Fig. 6.3(b).

With a triode valve (Fig. 6.2(a)) in which the interelectrode capacity between anode and grid is likely to be relatively high, and so gives more than adequate feedback, the anode voltage should be limited to between 100-150 volts for BT and similar low temperature coefficient cuts, with the power output level between 0·5-1 watt. If X cut crystals are employed, it is permissible to increase the applied voltage to 250 volts, and the power output to 2-3 watts. However, when maximum stability is required, the power in the crystal oscillator should be kept to the lowest possible value consistent with adequate drive to the following stage; if the desired conditions cannot be achieved a buffer stage should be used after the crystal oscillator.

With the pentode or tetrode version of the simple crystal oscillator shown at (B) in Fig. 6.2 the h.t. supply can be increased without risk of damage to the crystal and the corresponding power output can be increased to a possible maximum of 5 watts with a suitable valve. This is because a pentode or tetrode has greater gain, allowing a reduction of the r.f. crystal current for a given output power. A further advantage is that the amplitude of the output can be controlled by varying the screen voltage with a suitable potentiometer. If there is very effective screening between the grid and anode circuits, the small capacitor C of 1-2 pF may be required to provide sufficient feedback to maintain oscillation.

The r.f. current through an X-cut crystal may be monitored by means of a small low-current lamp in series with it, such as a 60 mA lamp. Alternatively the current may be measured with a suitable thermocouple meter of about 60 mA full-scale deflection. AT-cut and BT-cut crystals are not suitable for such high currents and no convenient method is available to the amateur for current monitoring purposes. The most satisfactory technique, therefore, is to ensure that the condition whereby excessive current is obtained cannot arise by limiting the power input to the stage. A convenient method of quickly checking the power output of an oscillator, in the absence of better test equipment, is to couple a low-power lamp to the anode-circuit inductor (or "tank" circuit) with a single-turn loop of wire: see **Fig. 6.4**. The loop will absorb sufficient power from the anode coil to cause the lamp to light.

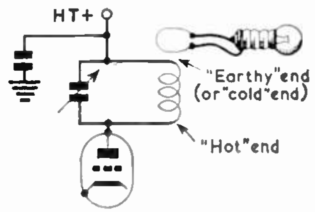


Fig. 6.4. A dial-lamp with a single turn loop of wire soldered to it will glow brightly when coupled to the coil in a tuned circuit in which r.f. power is present.

The Pierce-Colpitts Oscillator

In the Pierce-Colpitts oscillator, shown in **Fig. 6.5**, the crystal operates at a frequency just below the parallel resonance frequency and its equivalent inductance resonates with the anode-to-earth and grid-to-earth capacitances. The simple untuned circuit shown at (A) in Fig. 6.5 will oscillate with crystals of any frequency and is therefore convenient where widely different frequencies are to be used for various channels. However, if the crystal happens to have a spurious

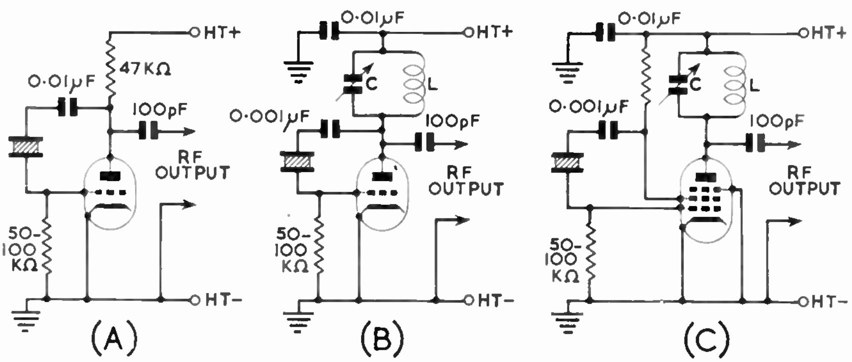


Fig. 6.5. The Pierce-Colpitts crystal oscillator. (A) shows the simplest form of untuned oscillator. In (B) the resonant circuit enables the fundamental frequency or harmonic output to be selected. The electron coupled circuit shown at (C) is especially useful for developing a large harmonic output.

TABLE 6.2

Harmonic output voltage compared to crystal frequency and the order of the harmonic when employing a 6AG7 valve in the circuit of Fig. 5(C).

Harmonic	Fundamental Crystal Frequency 3·7 Mc/s	5 Mc/s	11 Mc/s	14 Mc/s
Second	117	142	43	41
Third	58	71	26	25
Fourth	27	21	7	15
Fifth	21	16	2·7	9
Sixth	12	12	2·9	4·3
Seventh	7	7	3·3	—

These voltages were obtained using an anode voltage of 250 volts.

mode of high activity at a frequency differing from the desired frequency, this may be accidentally excited. The arrangement shown at (B) avoids this possibility by the use of a tuned circuit *LC* to select the desired frequency and at the same time gives appreciably greater output. An electron-coupled circuit based on the untuned Pierce-Colpitts oscillator is shown at (C). Here the screen grid acts as the "anode" of the oscillator and electron coupling takes place through the earthed suppressor grid from the screen to the actual anode. It is dangerous to tune the output tank circuit *LC* to the crystal frequency, since the crystal current may then be excessive. Any desired harmonic can, however, be selected by suitably tuning the anode circuit. Measurements taken on a typical pentode valve indicate that harmonics up to the sixth are of useful magnitude and **Table 6.2** shows the voltages that can be expected.

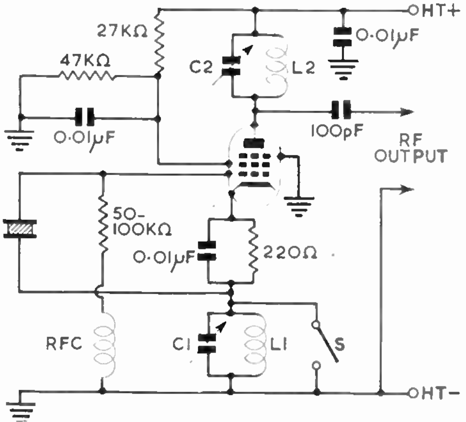


Fig. 6.6. The tritet oscillator. The oscillator feedback circuit L1C1 is tuned to approximately the crystal frequency, while the anode tank circuit L2C2 selects the desired harmonic frequency. Care is needed to avoid tuning L2C2 to the crystal frequency. If output is required at the fundamental the switch S must be closed to short-circuit L1C1. Many different types of valve may be used, e.g. 6AG7 or 5763.

The Tritet Oscillator

The triode-tetrode (tritet) circuit was intended mainly for the production of harmonics of the crystal frequency, but is now really only of classical interest. A complete practical circuit is shown in **Fig. 6.6**; it must not be used with modern miniature crystals owing to the danger of fracture.

An analysis shows it to be a form of tuned-anode/tuned-grid oscillator. In this case the crystal constitutes the resonant grid circuit and L1C1 is the resonant feedback circuit, the effective anode being the screen grid (as in the electron-coupled oscillator). Besides the feedback which occurs in the oscillator section, coupling takes place from the main anode circuit since current through this circuit passes also through the tank circuit L1C1 in the cathode lead. Feedback is therefore least when the stage is not loaded.

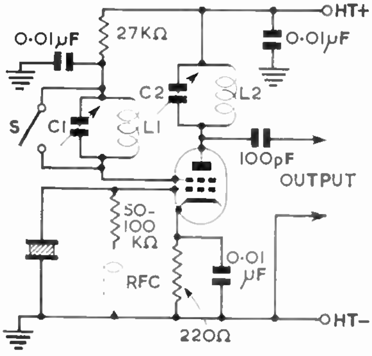
The circuit is adjusted by first tuning the cathode capacitor C1 from its maximum value downwards; oscillation begins when L1C1 is tuned to the fundamental crystal frequency and is indicated by a drop in anode current. The anode circuit L2C2 is then tuned for minimum feed current at the chosen harmonic frequency (which should be verified with an absorption wavemeter); at this setting the power output is at its maximum. By increasing C1 slightly above the setting which gives minimum anode feed current, the crystal current will be reduced and the frequency stability improved.

The value of L1 is important; it must be so chosen that C1 is large enough at the required resonance frequency to act as a low-reactance bypass for the harmonic frequencies which may be generated in the anode tank circuit L2C2. A recommended arrangement is a 200 pF variable capacitor in parallel with at least a 100 pF fixed capacitor. The addition of this fixed capacitance has the advantage that it will limit the tuning range and prevent the selection of a harmonic frequency with its associated heavy crystal current and the possibility of overloading caused by oscillation between the two tuned circuits. Output at the fundamental frequency is obtainable by short-circuiting L1C1. The circuit is then identical with that shown at (B) in Fig. 6.2. An important point to note is that the anode circuit must never be tuned to the fundamental frequency of the crystal without L1C1 being short-circuited; otherwise there will be excessive crystal current owing to self-oscillation between the anode and cathode circuits.

Greater power output can often be obtained by the use of a beam tetrode rather than a pentode. However, if the latter is preferred the suppressor grid should have an independent connection which must be directly earthed: a pentode in which the suppressor grid is internally joined to the cathode will not be satisfactory. Valves with good screening between anode and grid circuits are preferable.

If it is inconvenient to allow the cathode of a tritet oscillator to be at a radio-frequency potential with respect to earth, the circuit may be rearranged as shown in **Fig. 6.7**.

Fig. 6.7. Alternative circuit arrangement of the tritet oscillator shown in Fig. 6.6. Here the cathode is maintained at zero r.f. potential.



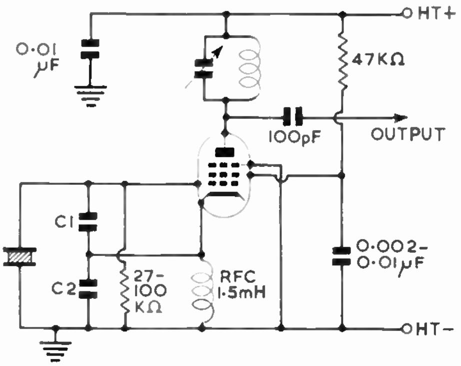


Fig. 6.8. Harmonic crystal oscillator. Ample power output is available on the fundamental and harmonic frequencies. Recommended valves: EF91, 6AM6, EL91, 6AM5 or similar.

Crystal Frequency	C1	C2
2-4 Mc/s	10pF	100pF
4-16 Mc/s	4·7pF	47pF

The circuit then operates in exactly the same way as before, except that the electron coupling to the true anode is accomplished without permitting the output current to pass through the oscillator tuned circuit L1C1. At the harmonic frequencies, since the output current is not passed through the oscillator tuned circuit L1C1, slightly greater output can be expected than with the first arrangement. The usual precautions to avoid danger to the crystal should be observed.

Colpitts Oscillator

The circuit of the " harmonic " crystal oscillator, shown in Fig. 6.8, appears to resemble that of the tritet oscillator. In fact, however, it is a modification of the Colpitts oscillator described on page 6.7; the " anode " of the oscillator section proper (i.e. the screen) is at zero r.f. potential and feedback takes place through the capacitors C1 and C2. The anode circuit is electron-coupled to the oscillator section, and good power output at either the fundamental or a harmonic frequency can be obtained. No cathode tuning circuit is required, the purpose of the r.f. choke being to provide a d.c. return path for the cathode current while at the same time allowing the cathode to take up the correct r.f. operating potentials.

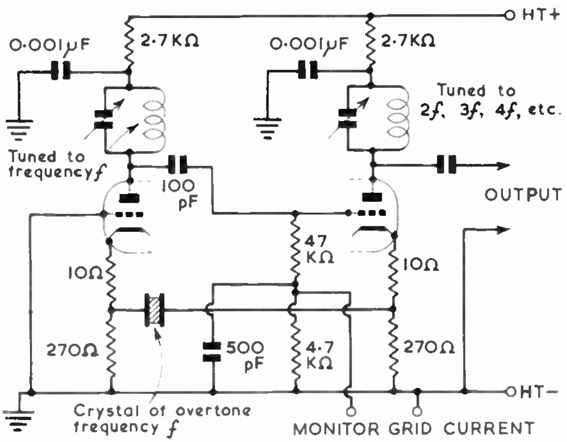


Fig. 6.9. The Butler overtone oscillator. This circuit is primarily intended for use with crystals designed for overtone operation.

Overtone Oscillators

The oscillator circuits described above produce vibration of the crystal plate in its fundamental mode. Any required harmonic output is then obtained by the non-linear operation of the valve, the harmonics being selected in the tuned-anode tank circuit. Harmonic output can alternatively be obtained by making a crystal oscillate at an " overtone " of its fundamental. Theoretically the foregoing circuits could be adjusted to give overtone oscillation of the crystal itself, but it would in general be necessary to inhibit the tendency of the crystal to vibrate at the fundamental frequency; this may be difficult to accomplish with the circuits described.

Usually, circuits which assist the overtone mode of operation cause oscillation at the series-resonance frequency (of the overtone mode) and are typified by the Butler, Squier and Robert Dollar circuits: these are shown in Figs. 6.9, 6.10 and 6.11 respectively. Overtone oscillation is more often required for v.h.f. applications where a large frequency-multiplication factor is wanted, and further information on these circuit types is therefore given in Chapter 7—*V.H.F./U.H.F. Transmitters*. Nevertheless, the possibility of using overtone modes to assist in the avoidance of interference (for example, with television) at certain frequencies is suggested. The separation of the harmonics produced (compared with operation of the crystal at its fundamental frequency) is increased; the harmonics which are produced are multiples of the overtone frequency.

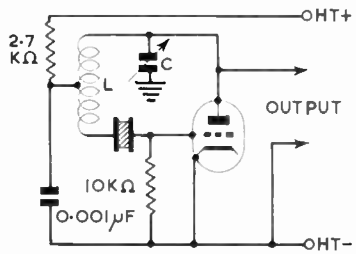


Fig. 6.10. The Squier overtone oscillator. The circuit LC should resonate at the overtone frequency and most small triodes are suitable. The valve may be one half of a twin triode such as a 12AT7.

When such an oscillator is adopted, care has to be taken in the design of the transmitter as a whole to ensure that none of the unwanted harmonics are eventually radiated.

Varying the Crystal Frequency

There are two methods of varying the frequency of a crystal oscillator. One is mechanical and the other is electrical.

The mechanical method requires an alteration of the dimensions of the crystal plate or its mass in some way. This generally involves the grinding of the surface to *increase* its resonant frequency or the loading of the vibrating part to cause a *reduction* of the frequency. Such changes to the plate must of course be regarded in general as permanent

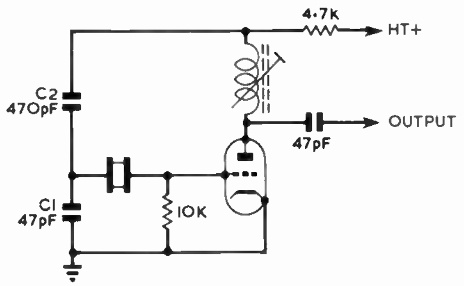


Fig. 6.11. The Robert Dollar overtone oscillator. This circuit does not require the use of a tapped coil, and may be used with the same valves as the Squier oscillator.

and irreversible. Crystal grinding, mentioned in Chapter 1 *Principles*, requires great care and mistakes made through lack of skill can easily have disastrous results. The limit to the mechanical loading of a crystal is usually set by a gradual decrease of activity as material is added; for this reason a maximum reduction of about 25 kc/s is possible with a 3·5 Mc/s plate. Suitable loading material for a crystal is cold soft solder. It is gently and gradually rubbed into the two surfaces as evenly as possible near the plate centre; the activity should be checked at regular intervals, and in the event of a decrease in activity being noted no further loading should be attempted. Another material found suitable for this purpose is ordinary pencil-lead.

In the electrical method, a reactance is connected in series with the crystal when the circuit is operated at the series-resonance crystal frequency, or in shunt with it for the parallel mode. The effects of such added reactances are summarized in Table 6.3. This electrical method is often useful for making slight frequency changes such as may be required to avoid temporary interference. The maximum adjustments possible are very small; for instance in the 7 Mc/s band the change will not be more than a few kilocycles. The limits of variation are set either by instability or by a complete lack of oscillation.

Mixer Crystal VXO Circuits

An increase in the frequency deviation obtained by varying crystal frequencies can be achieved by employing two separate crystal oscillators, one of which is tuned by a reactance which makes its frequency go l.f.; the other is tuned by a reactance which moves the frequency in an h.f. direction. Such oscillators used in a circuit similar to that of Fig. 6.12 will permit greater deviation to be secured, compared with a basic frequency crystal.

If two oscillators are used, one on 8·5 Mc/s, the other on 12 Mc/s, and assuming that the 8·5 Mc/s crystal is changed l.f. by 1 kc/s per Mc/s, and the 12 Mc/s crystal h.f. by 1 kc/s per Mc/s, the maximum alteration in frequency will be $12 + 8{\cdot}5 \sim 20$ kc/s.

In many of the oscillator circuits described, the grid is connected to earth through a simple resistor. In some cases, an increase in output may be obtained by fitting an r.f. choke between the grid of the valve and the resistor but this must be done with care however because, in certain circumstances and particularly with a valve of high mutual conductance, self-oscillation may occur if the choke is of high *Q*.

If the crystal fails to oscillate in some circuits it is possible for the anode current to rise to such a high level that the valve is destroyed. This can be avoided by fitting a resistor in the cathode circuit to provide sufficient bias to limit the anode current to a safe value in case of crystal failure. It is essential to bypass such a resistor for r.f.

When using pentode or tetrode valves as oscillators, the screen voltage should be derived from a potential divider rather than the usual series dropping resistor. For best

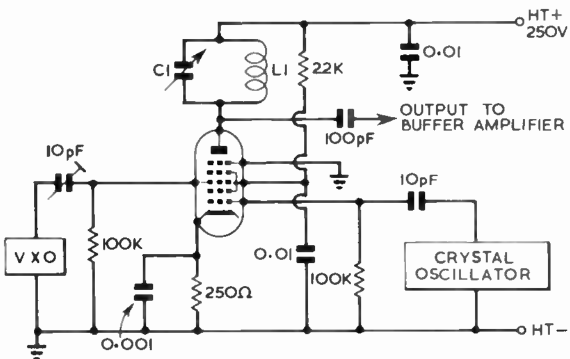


Fig. 6.12. Low-level mixing circuit for a VXO showing typical values for a 6SA7. The values of L1 and C1 are similar to those for a normal tank circuit.

stability, however, both anode and screen voltages for oscillator valves should be obtained from a stabilized source. Electronically regulated supplies are preferable to those depending on simple voltage regulator tubes.

Keying of any oscillator is seldom satisfactory but when it is necessary the crystal oscillator must receive special attention. First, the crystal itself must be of high activity: an inactive crystal is likely to produce chirp due to changes in frequency as oscillation commences. Secondly, the oscillator must be lightly loaded and the crystal current kept as low as possible. D.c. keying was popular at one time, the usual method being to do so in the cathode lead. This is a dangerous practice from the point of view of the lethal voltage that may exist across the key as well as overstressing the heater/cathode insulation. A better method is to employ grid block keying in which a high negative voltage sufficient to bias the valve to cut-off is applied under key-up conditions. When the key is closed, the negative supply is short-circuited to earth, thus allowing the oscillator to function normally. A high resistance in series with the negative supply limits the

TABLE 6.3

Effect on Crystal Frequency of Adding External Reactances

Reactance Added	Series-resonance Frequency	Parallel-resonance Frequency
Capacitance in series	Increased	Unaltered
Capacitance in parallel	Unaltered	Lowered
Inductance in series	Lowered	Unaltered
Inductance in parallel	Unaltered	Increased

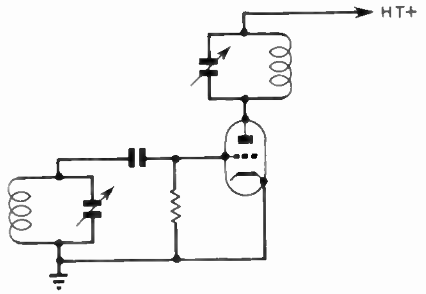


Fig. 6.13. Tuned anode tuned grid oscillator.

current under key-down conditions and protects the operator.

Information on keying and control methods is given in Chapter 8 *Keying and Break-in.*

VARIABLE FREQUENCY OSCILLATORS

Although the crystal oscillator is the most simple, stable and satisfactory device for providing accurate frequency control, it suffers from the obvious disadvantages of fixed-frequency operation, particularly on congested amateur bands where the operating conditions change from minute to minute. A *variable frequency oscillator (v.f.o.)* is now essential for most if not all of the h.f. amateur bands. The frequency-determining circuits of such an oscillator consist of ordinary inductances and capacitances, either or both of which may be variable. Special care must be taken to eliminate all possible cause of unwanted frequency variation.

A v.f.o. of good design should give a frequency stability comparable with that of a crystal oscillator. Some idea of the stability desirable may be obtained by considering the following example. A sudden frequency variation of 50 c/s when operating in a congested c.w. band can cause the signal to be lost or its intelligibility to be severely reduced if the narrowest receiver passband is in use: on the 21 Mc/s band this variation would correspond to only one part in 400,000 of the transmitted frequency.

Basic Types of V.F.O.

Many different v.f.o. arrangements have been developed, but when analysed they are found to be modifications of a few basic circuits. The modifications are introduced with practical aspects in view; e.g. to allow the earthing of one end of the tuned circuit or to improve the stability.

An oscillator is essentially a tuned amplifier in which some of the output voltage is fed back to the grid 180 degrees out of phase with respect to the anode voltage. The appropriate feedback may be obtained either by mutual inductance between the anode and grid circuits or by capacitive coupling between parts of the circuit at which the correct phase and voltage relationships exist for positive feedback to occur. Four basic oscillators are shown in Figs. **6.13** (the tuned anode tuned grid), **6.14** (the Hartley), **6.15** (the Colpitts) and **6.16(a)** (the Franklin).

The tuned anode tuned grid oscillator, Fig. 6.13, is similar to the Pierce-Miller crystal oscillator of Fig. 6.2(A). When the anode and grid circuits of the t.a.t.g. (sometimes known as a t.p.t.g.) are tuned to the same frequency, feedback occurs through the internal capacity of the valve from anode to grid. This simple oscillator is no longer used in practice but its principle of operation should be borne in mind.

In the Hartley oscillator, Fig. 6.14, feedback is achieved by connecting the cathode of the valve to a tap on the tuned grid circuit. As the anode current also flows through the cathode of the valve, this arrangement effectively allows part of the r.f. output current to flow through the tuned circuit in phase with the grid circuit current. The degree of feedback is determined by the position of the tapping point on the coil. The larger the number of turns through which the cathode current flows the greater the feedback. Alternatively, the oscillator can be considered as having the grid and anode connected to opposite ends of a tapped tuned circuit, the voltages at the two ends being in antiphase with respect to the tapping point. The excitation ratio is the ratio of anode-tap : grid-tap turns.

The cathode of the valve in the Colpitts oscillator, Fig. 6.15, is connected to a capacitive tap across the grid tuned circuit provided by C1 and C2. The feedback arrangement is therefore similar to the Hartley. C2 is bypassed for d.c. by the r.f. choke.

The basic Franklin oscillator shown in Fig. 6.16(a) is very

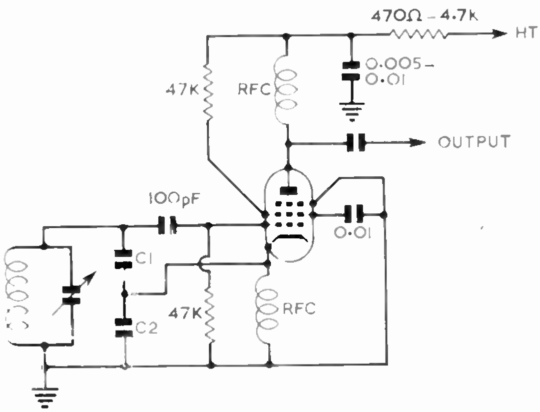


Fig. 6.15. Basic electron-coupled Colpitts oscillator.

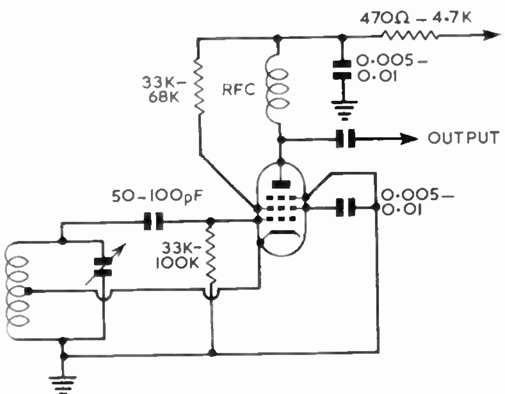


Fig. 6.14. Basic electron-coupled Hartley oscillator.

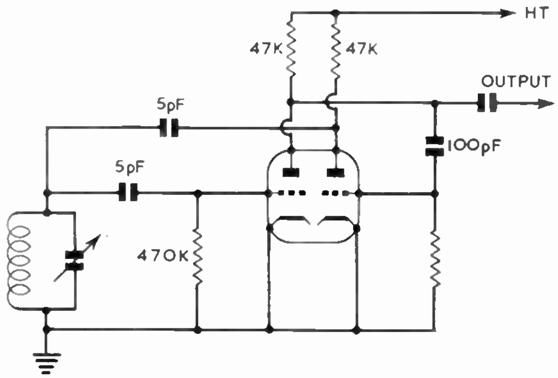


Fig. 6.16(a) Basic Franklin oscillator.

similar to a conventional resistance-capacitance coupled amplifier with the addition of two capacitors to provide feedback. Disregarding the tuned circuit, the feedback path is formed by the two 5pF capacitors in series connected between the input of the left-hand triode and the output of the right-hand valve. A parallel tuned circuit is connected between these two capacitors and earth. At frequencies other than the resonant frequency of this tuned circuit there is a short-circuit to earth across the feedback path but at resonance there is a high impedance to earth. Oscillation will therefore only occur at the resonant frequency of the tuned circuit.

The circuit of the so-called cathode coupled oscillator is shown in Fig. 6.16(b) which first appeared in *CQ Magazine* in April 1960. The circuit is very tolerant of high *C* operation, a useful feature where a wide range oscillator is required. However, under certain conditions the circuit may squeg, particularly if the grid coupling capacitor is increased much beyond the suggested 10pF. A small non-inductive resistor of 10 to 50 ohms will prevent this and enable a nearly constant output to be obtained over several bands without altering the grid capacitor.

The h.t. bypass C3 should be connected directly between the "free" anode of the valve and the earthing point of the oscillator to ensure low harmonic output. For harmonic suppression, the series inductance of the capacitor is more important than its actual value and it is worth trying different physical types. For example, a cylindrical paper capacitor, particularly if in an earthed metal casing, may be found

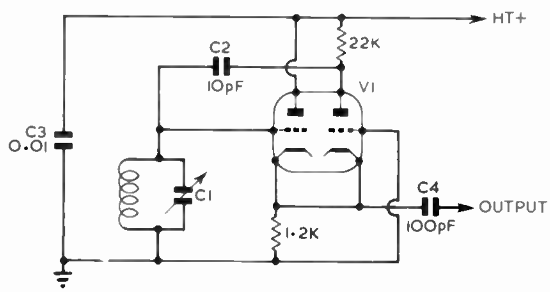


Fig. 6.16(b). The so-called cathode coupled oscillator. V1 may be a 6J6 or 6SL7GT. It will be seen that the circuit is very similar to the standard Franklin oscillator.

superior to a silvered mica type whereas with different makes the opposite might be true. Hi-*K* ceramic types seem to be unsuitable for this application. A combination of different types sometimes proves satisfactory. A 5K ohms resistor in the "free" anode circuit may also prove helpful in eliminating harmonics.

In these basic types of oscillator or in any of their modifications the frequency stability is limited primarily by the stability of the frequency-determining tuned circuit and also by the effect of the valve and other components associated with it: other factors such as the operating voltages and currents and the loading of the oscillator are likewise important. All these design problems require close and careful study if satisfactory performance is to be achieved.

Tuned-circuit Stability

To ensure the highest stability over both long and short periods the primary requirement is that the frequency-determining tuned circuit shall be in every way of a high standard. Mechanically, it must be perfectly rigid since any vibration or displacement of the various parts of the tuned-circuit assembly will cause variations in the effective inductance and capacitance and therefore in the oscillator frequency. To overcome these effects, the whole oscillator (in particular the items associated with the tuned circuit) must be made as mechanically rigid as possible. It is preferable to construct the oscillator in a die-cast box, which gives sufficient strength for most amateur purposes and at the same time screens the v.f.o. from external fields. The wiring should be of stout gauge (16 s.w.g. or thicker).

The tuning capacitor must be of low-loss construction, preferably with ceramic insulation for the stator supports and with bearings for the rotor which prevent any tendency towards wobble or backlash. The electrical connection to the rotor should be made separately through a spring contact—or a pig tail—in order to avoid the introduction of spurious resistance paths through the bearings, which can cause frequency scintillation—small random changes in frequency. Since the capacity of a variable capacitor depends not only on the area of the fixed and moving plates exposed to each other, but also on the spacing between them, it is advisable to employ a tuning capacitor with widely spaced plates. Like any other item, heat will cause expansion in the tuning capacitor, and the resulting capacity change will be far less in a unit with widely spaced plates compared to one in which the plates are physically close. The capacitor must also be selected with regard to the nature of the dielectric supporting the fixed plates from the main frame, for consideration will show that this dielectric is directly across the tuned circuit when this is a parallel configuration, and across the capacitor itself in a series tuned arrangement. High grade ceramic or preferably p.t.f.e. must be used for these supports. Materials such as paxolin and Bakelite are quite unsuitable. To reduce r.f. losses to the minimum, it is desirable that the plates themselves should be silver plated and soldered together.

The coil should be wound in such a manner as to make any movement of its turns virtually impossible. Apart from the use of grooves in the former, sufficient rigidity can be achieved by winding the wire under tension.

There is an increasing tendency to employ physically small coils in oscillator circuits, and to fit these with ferrite cores in order to achieve the desired inductance. Such a procedure is satisfactory so long as the ferrite material is not subject to a high coefficient of expansion, and the core itself is firmly locked in place. If such coils are used and frequency drift is noted above that which may reasonably be expected, a change of ferrite core type may well result in a substantial improvement in frequency stability.

Changes in the values of the tuned circuit components are not the only source of frequency drift. Instability in the values of other components or changes in the interelectrode capacities of the valve may produce capacity changes across the tuned circuit or result in an alteration in the working point of the valve, both of which can cause the operating frequency to be unstable.

Circuit Q

An important factor in the maintenance of frequency stability is the loaded Q of the frequency controlling circuit, the stability being improved with increase of Q. This is because in a high Q circuit the phase change of the voltage existing during oscillation across the circuit varies with frequency more rapidly than in low Q circuits. Thus, in the high Q case, random phase changes in the oscillator feedback path are cancelled out by smaller changes of frequency than in the low Q case.

The loaded Q value of the tuned circuit may be improved by reducing the losses in its components, the chief loss usually occurring in the coil. Therefore coil formers, which may be ribbed, should be made of low loss material (e.g. p.t.f.e., polystyrene or ceramic) and the wire must be copper, preferably silver plated; Litz wire is useful at frequencies below 1 Mc/s. A specific shape can be found which gives highest *coil Q*.

The coil should be situated at least 1-1·5 coil diameters away from metal screening or other components to prevent loss of energy and subsequent lowering of Q due to induced eddy currents and dielectric losses. Other items associated directly or indirectly with the tuned circuit such as padding capacitors, insulators and so forth must be of high grade material, as any dielectric losses will be reflected in a reduced circuit Q. Capacitors should be of air or mica dielectric except for any necessary temperature compensating ceramic capacitor.

In the practical oscillator the frequency controlling circuit must be coupled to the oscillator valve. The effect of this is to reduce the overall Q of the circuit by an amount depending on the degree of coupling. The coupling should therefore be

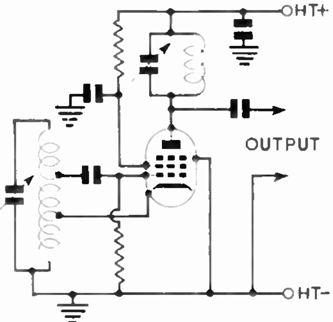


Fig. 6.17. Method of increasing the stability of the Hartley oscillator by tapping the grid connection down the tuned circuit.

small. For a given amount of loading, however, the Q of a tuned circuit is proportional to the C/L ratio, which should consequently be as great as possible. In fact the C/L ratio is usually chosen as a compromise between greatest ratio and the ability to tune over the required frequency range using components of convenient value.

Apart from the improvement in loaded Q value obtained through the use of high C/L ratio in the oscillator tuned circuit, the effect of changes in the valve interelectrode capacities due to extraneous influences is considerably reduced, as the ratio of valve capacity to total circuit capacity is very small and as a result is less capable of varying the resonant frequency.

Reduction of oscillator *loading* as a means of improving the circuit Q is exemplified by the Franklin oscillator (see Fig. 6.16). In this oscillator the two valves are extremely loosely coupled to the tuned circuit through two very small capacitors.

In another type of oscillator the connections from the valve electrodes are tapped down the coil (example: **Fig. 6.17**) thus reducing the loading effect of the valve on the circuit and the influence of the interelectrode capacity on the resonant frequency.

All other things being equal, the most stable oscillator will be that in which the loosest coupling exists between the tuned circuit and the valve, or a circuit in which the capacity across the tuned circuit is extremely large in proportion to any likely external capacity changes.

The valve, because of its non-linearity, will generate harmonics of the oscillator frequency. These can intermodulate with each other producing fundamental components in random phase relationship with the original fundamental component. The presence of these spurious components will produce instability of the frequency but their effect can be considerably reduced by employing high Q circuits which by virtue of their selectivity prevent them being fed back to the grid circuit.

It is also possible for the frequency to change if fundamental components are fed back to the oscillator from high power stages of the transmitter. Since these will change in amplitude with keying or modulation they will combine with the existing fundamental component and the resultant will change in phase. The oscillator will then alter frequency so that the loop phase shift is zero. Such frequency changes can sometimes be of the order of 100 c/s or more. The solution is effectively to screen the complete oscillator unit and bypass and filter all supply leads including the heater.

Avoiding Temperature Effects

Temperature variations can produce a considerable change in the resonant frequency of a tuned circuit. The effect of a temperature change upon the coil in such a circuit is to cause an expansion or contraction of the former and wire with a consequent change in inductance and resistance values. Both of these values normally increase with the temperature.

The various parts of the tuning capacitor assembly will likewise contract and expand and the amount of the resultant change in capacitance will depend on the type of construction. In general it is found that the resonant frequency of an ordinary tuned circuit comprising a parallel-connected coil and capacitor has an effective negative temperature coefficient of 50-100 parts/million/degree C.

The most obvious method of avoiding the frequency drift caused by temperature variation is to remove the tuned circuit from all sources of heat. The tuned circuit should be located in a part of the equipment well away from valves (which can radiate a large proportion of their wasted power) and high-wattage resistors. It is preferable to place the circuit within its own compartment, and if this is brightly polished on the outside any unavoidable heat will be reflected away from the surface.

The second method, which is in common use owing to the difficulty of preventing all heat from affecting the tuned circuit, is to compensate the frequency variation by means of capacitors having a suitable negative temperature coefficient. When these are connected across the tuned circuit in their correct capacitance proportions, any temperature rise will tend to cause them to increase the resonant frequency, whereas in the remainder of the circuit the effect of the temperature rise is to lower the frequency. The result is that the frequency stays almost constant, i.e. the overall temperature coefficient of frequency is reduced almost to zero. To obtain a required tuning range with a given variable tuning capacitor it is necessary to connect a fixed capacitance of suitable size across the tuned circuit, and a selected proportion of this capacitance may conveniently be of the negative temperature coefficient type. The compensating capacitors themselves are usually of the tubular wire-ended ceramic type and are available in almost any nominal value of capacitance with several different values of temperature coefficient.

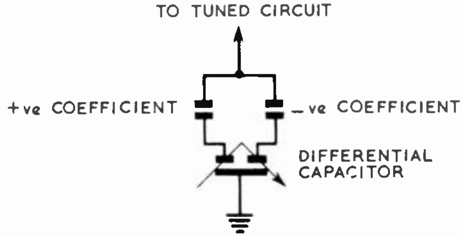


Fig. 6.18. Temperature compensating circuit employing a differential capacitor.

Selection of the correct value of compensating capacitor can be a laborious task. An ingenious solution to the problem is shown in the circuit of **Fig. 6.18** in which a positive temperature coefficient capacitor and a negative coefficient capacitor of the same value are connected to the two arms of a differential capacitor. It is convenient to make both fixed capacitors the same value and by varying the differential capacitor the temperature coefficient may be adjusted precisely. The local capacitance " seen " by the tuned circuit will vary as the differential capacitor is adjusted (max. in mid-position) but this can be compensated by adjusting the band-set capacitor. The " Tempatrimmer " manufactured by Oxley Industries is a commercially available component performing the same function as the arrangement of Fig. 6.18. Furthermore, if the oscillator is operated at a low power level there will be less difficulty in preventing a temperature rise in the tuning circuit due to heat from the valve. There will also be less heat developed in the tuning circuit itself (arising from r.f. losses), but in any case this is likely to be quite small compared with the amount of heat which reaches it from the valve.

Power-supply Variations

Variations in the supply voltages to the oscillator valve can alter its interelectrode capacities and other parameters. Details of suitable methods of stabilizing the power supply voltages are given in Chapter 17—*Power Supplies*. Changes in heater voltage generally produce negligible variation of frequency, but their effects should not be ignored.

Loading Effects of Following Stages

The loading effect on the oscillator of the buffer or frequency-multiplying stages which follow it causes reactive and resistive components to be reflected into the frequency-determining circuit of the oscillator. The reactive components cause direct frequency variations while the effect of resistance across the resonant circuit is to decrease its Q, thereby reducing the inherent frequency stability. Moreover, r.f. voltages may be fed back from the succeeding stage. These voltages, the amplitudes of which depend upon the power output, and the tuning adjustments of other parts of the transmitter, will affect the calibration and the stability of the v.f.o.

For maximum stability the coupling to the following stage must therefore be as loose as possible consistent with the required drive being obtained.

The only way in which frequency variations due to changes in loading can be overcome is by the use of adequate buffer amplifier stages after the v.f.o. It is not possible to remove all loading from the v.f.o. since it must supply some power to the following stage, but it is possible to arrange the loading on the v.f.o. so that the frequency stays sensibly constant. The following stages then supply the required drive power.

It is strongly recommended that the v.f.o. should be followed by an untuned class A buffer amplifier or a cathode follower, as shown in **Fig. 6.19.** Both of the circuits shown

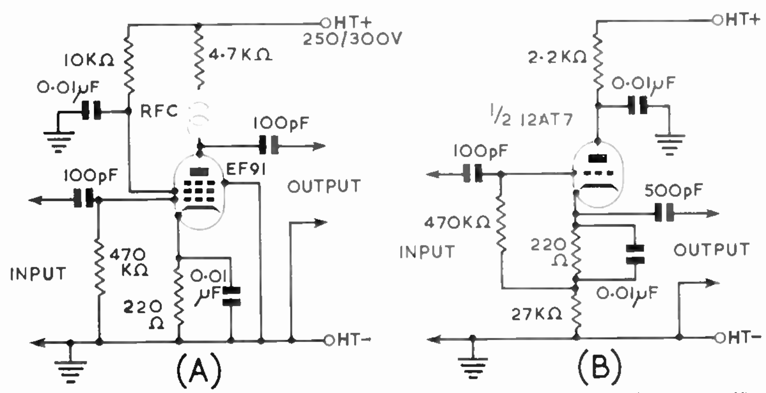


Fig. 6.19. Methods of isolating the v.f.o. from loading effects. The untuned class A amplifier shown at (A) provides a useful amount of gain besides affording adequate isolation. In the cathode follower (B), the impedance presented to the previous stage is very high while the output impedance is very low (e.g. about 100 ohms): the voltage gain is, however, less than 1. In view of the low output impedance, the cathode follower is well suited for feeding into a coaxial line.

in Fig. 6.19 present a very high impedance to the oscillator stage, and feedback from subsequent stages is at a satisfactorily low level.

The class A amplifier shown at Fig. 6.19(A) is suitable for feeding another stage (the coupling lead from the output capacitor should be kept short). The following stage may in most cases be a small tuned class C amplifier.

The circuit is that of a pentode amplifier operating in class A giving a high degree of isolation. For maximum efficiency, the output from the anode to the following stage should use the shortest possible lead length, and for preference be no longer than the wires on the 100pF coupling capacitor. To ensure that the stage operates under class A conditions when connected to the v.f.o., the connection between the earthy end of the 470Kohm grid resistor and the chassis should be broken, and a 0-500 μA meter inserted. With both the v.f.o. and the buffer stage operating, the meter should be checked for any signs of grid current. If current is present, then the 100 pF coupling capacitor in the grid circuit must be reduced in value to the point where grid current ceases.

Although the gain of the cathode-follower circuit shown at Fig. 6.19(B) is less than unity, the degree of isolation between input and output circuits is extremely high. The output impedance is quite low (approximately equal to $1/g_m$, where g_m is the mutual conductance at the operating point of the valve used); a typical value is 100 ohms. It is therefore possible to feed the output directly and without serious mismatch into a low-impedance co-axial cable which can then supply the main transmitter at a point remote from the v.f.o. unit.

V.F.O. Design Recommendations

The rules for the avoidance of frequency drift and other variations in a v.f.o. may be summarized as follows:

(a) Ensure complete rigidity and strength of the mechanical structure and wiring.

(b) Protect the tuned circuit from the effect of heat sources such as valves and resistors.

(c) Construct the tuned circuit and associated oscillator components of material having the lowest r.f. losses.

(d) Avoid draughts across the tuned circuits.

(e) Operate the oscillator at the lowest possible power level.

(f) Provide stabilized h.t. to the valve.

(g) Use a buffer amplifier and/or cathode follower to minimize the loading effect of the subsequent stages.

(h) Screen and decouple all the circuits to prevent the pick-up of r.f. energy from higher-power stages.

(i) Use heat dissipating valve screens.

(j) Operate the oscillator or tuned circuit at a fixed temperature (above maximum ambient) by means of an oven and thermostat.

Selecting the Frequency of the V.F.O.

It is usually advantageous to operate a v.f.o. on a relatively low frequency since, first, the stability in terms of frequency is improved as a consequence of the diminished effect of the various factors causing frequency variations and, secondly, it can facilitate multi-band operation. Owing to the harmonic relationship between the various amateur bands the output of an oscillator working on a suitable low frequency can be frequency-multiplied to produce the required drive voltage in one or more of the higher-frequency bands. The multiplication is usually effected in separate stages, but often it can be arranged within the v.f.o. unit itself. In many amateur stations a v.f.o. covering the range 1·75-2·0 Mc/s is found to be a very convenient means of providing multi-band frequency control. It is not generally satisfactory to operate the oscillator on the final carrier frequency.

V.F.O. Bandspread

It is desirable that the whole of the amateur band to be covered by the oscillator should be spread across the full width of the tuning dial. In this way the operating frequency can be quickly read, adjustment made easier, and the danger of out-of-band operation reduced. To produce the required bandspread, the tuning capacitance is usually made in the form of fixed (or pre-set) capacitance plus a small variable capacitance just sufficient to vary the frequency over the desired range. The necessary values may be determined by experiment or from simple calculations.

If for example a transmitter were being constructed for the 28 Mc/s-29·7 Mc/s band, and the v.f.o. was to operate in the 3·5 Mc/s region, the frequency multiplication required in the transmitter would amount to eight times (3·5 Mc/s $\times$ 8 = 28 Mc/s). Thus to cover 28 Mc/s-29·7 Mc/s the v.f.o. would need to tune from 3·5 Mc/s to 3·7128 Mc/s. In practice this would probably be adjusted to 3·49 Mc/s to 3·72 Mc/s to give about 10 kc/s overswing at each end. Similar calculations can be made for other bands.

One of the most satisfactory types of slow-motion drive for a bandspread tuning capacitor is the friction type with a ratio between 6 : 1 and 9 : 1 or the gear type. Frequency changes can then be made swiftly and the setting accuracy should be adequate, provided that a suitably marked dial is used. The dial may show ordinary scale divisions (0-100 or 0-180) but it is far more convenient to have a dial which has been directly calibrated. The frequency can then be read instantly and the risk of error that is always present in the use of a separate calibration chart is avoided. A particularly suitable slow motion drive and dial for the de-luxe transmitter is the Eddystone type 898.

Appraisal of a Newly Constructed V.F.O.

As the v.f.o. is the heart of the transmitter it is important that it should be thoroughly tested before the transmitter is put on the air. The following is intended as a guide to the procedure to be adopted when testing a v.f.o. for the first time.

(a) The unit should be installed in position in the transmitter or chassis of which it is intended to form part.

(b) The buffer stage should be connected and suitably dummy loaded.

(c) It may be useful to monitor h.t. current and this may be conveniently done by measuring the voltage across a series resistor in the h.t. lead. In the absence of such a resistor it will be necessary to break into the h.t. lead and measure the current directly.

(d) Apply power and observe that the h.t. current is of a reasonable value.

(e) A check may be made that the unit is oscillating by touching the grid lead of the valve and observing that the anode current increases.

(f) It is now necessary to determine the approximate fundamental frequency of oscillation; for this purpose a calibrated absorption wavemeter should be used. Find the strongest signal using the absorption wavemeter and tune well to either side of this signal in order to ensure that it is the fundamental output of the oscillator.

(g) If necessary, adjust the oscillator frequency to the required value determined by the absorption wavemeter, using the band-set capacitor. The position of the tuning capacitor should be appropriate to the chosen frequency, i.e. it should be fully meshed if the chosen frequency is maximum l.f. and vice versa.

(h) Depending on the accuracy of the absorption wavemeter used, it may be possible to use it to determine the bandspread to a first approximation; otherwise a crystal calibrator must be employed.

(i) It may now be convenient to determine the stability of the oscillator; the oscillator should be tuned to zero beat with the signal from the crystal calibrator and the drift measured over a period.

(j) Should the circuit include adjustable temperature compensation based on the Oxley " Tempatrimmer " capacitor then adjustments may be made to reduce the drift to acceptable limits. A hairdryer is a useful tool at this juncture since accurate setting of the temperature compensation may be a long and tedious process without accelerated warm-up.

(k) The components normally employed in v.f.o. circuits, for instance air spaced capacitors for tuning, band-set and band-spread, silvered-mica capacitors for the fixed values and valves all have positive temperature co-efficients. A small amount of negative temperature co-efficient compensation is therefore usually required. If capacitors employing other dielectrics are used note should be taken of their temperature co-efficients so that appropriate compensation can be arranged.

(l) The purity of note should next be checked by listening to the v.f.o. on a receiver. The presence of supply mains frequency modulation would normally indicate cathode heater leakage in the oscillator or buffer valve and X2 mains frequency modulation would indicate inadequate smoothing of the h.t. supply. Modulation at other frequencies is probably attributable to the presence of parasitic oscillations.

(m) A useful check for parasitic oscillations, apart from a wide range sensitive absorption wavemeter, is to monitor the h.t. current to the v.f.o. while tuning over its full tuning range. Sudden changes in h.t. current usually indicate the presence of parasitics.

(n) Should evidence of parasitic oscillations be found, reference should be made to page 6.36.

(o) Having achieved satisfactory operation of the v.f.o. to this stage of testing, accurate calibration may be carried out using the crystal calibrator.

As a practical criterion of stability, a fundamental 5 Mc/s v.f.o. which drifts no more than 50 c/s/hour under average shack conditions may be considered to be good.

Measuring Beat Notes

While it is not essential to measure the frequency of the beat note between the v.f.o. and the reference standard, it does permit some assessment of the effects of alterations.

If an oscilloscope and a variable frequency audio-oscillator are available they can be used to determine the frequency of the beat notes. (See Chapter 19—*Measurements*.) Not all workshops have a variable frequency audio oscillator and an oscilloscope. In such cases a somewhat less accurate method can be adopted. This involves the use of a meter arranged in a circuit in such a manner that its deflection is proportional to the applied frequency, so allowing the beat frequency to be read directly.

Although not of extremely high accuracy, a simple direct reading frequency meter does have the advantage that the frequency change taking place over a given period can be instantly seen without the manipulation of controls. The circuit of such a frequency indicator is given in **Fig. 6.20.** This instrument is provided with three ranges, the full scale deflections of which are 0–100 c/s, 0–1 kc/s, and 0–10 kc/s.

The transistor functions as a limiter amplifier, the output of which is differentiated by one of the three capacitors associated with the range switch, resulting in a train of positive and negative going pulses. The negative pulses are clipped by the diode CR1, the remaining positive pulses being fed to the metering circuit where they are integrated by the capacitor Cx. The voltage thus developed across Cx is linearly proportional to frequency and the meter measuring the voltage may therefore be calibrated directly in cycles per second. It is essential that the input to this circuit be sufficient to allow the transistor to be driven into limits, as otherwise the meter readings may be incorrect. Approximately 2 volts peak-to-peak is needed. A minor disadvantage of this device is that the indications are not accurate below about 5 per cent of the frequency range in use and thus while the lowest range claims 0–100 c/s, frequencies lower than 5 c/s

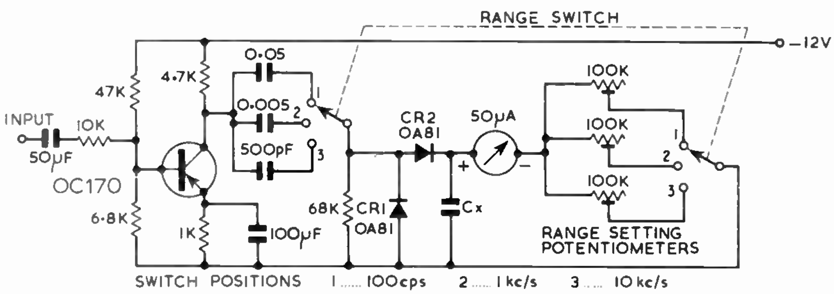


Fig. 6.20. A direct reading frequency indicator suitable for the determination of the frequency of beat notes in the range 5c/s to 10 kc/s. The 50μA meter may be replaced by a 100 μA meter; if this is done the 100K ohm range setting potentiometers must be reduced in value to 50K ohms each.

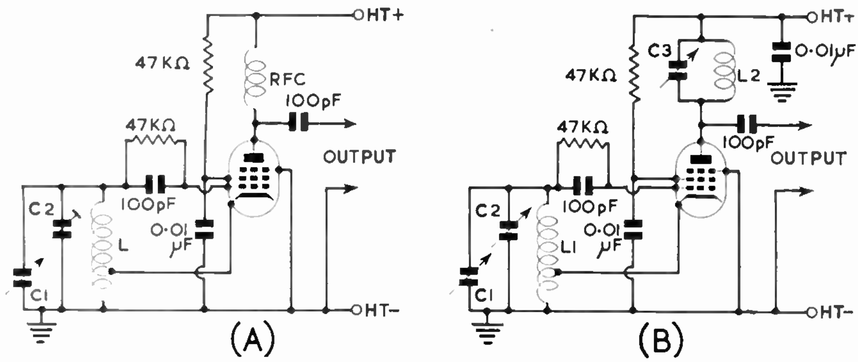


Fig. 6.21. Two versions of the Hartley circuit. In (A) the output works into an r.f. choke, while in (B) a tuned circuit is employed resulting in a higher output level. For operation in the 3·5 Mc/s range, L1 will be about 4 μH, C2 (composed of a combination of fixed and variable capacitors in parallel to achieve the desired value), will total about 500 pF, and C1, the tuning capacitor, 150 pF. An EF91 (6AM6) or similar small pentode with separate suppressor grid is suitable. The h.t. voltage must be stabilized.

should not be taken seriously. Consistent accuracy is maintained on all readings which are above 10 per cent of full scale deflection.

To calibrate the instrument the range potentiometers should first be set to maximum resistance. On Range 1 the output from the 6·3 volt winding of a mains transformer should be applied to the input and the range potentiometer adjusted so that the meter reads half scale. Range 2 can be calibrated against either the BBC tuning signal of 440 c/s or the 1000 c/s, tone duration transmitted by MSF. An audio generator is needed to calibrate Range 3.

V.F.O. CIRCUITS

Many different types of circuit are in common use, each one having its own particular merits. Many of the arrangements described here will be found satisfactory for use in amateur equipment.

The Electron-coupled Oscillator

In this oscillator a pentode valve is used in such a way that the screen and control grids correspond to the anode and grid respectively of a triode. Electron flow through the valve is controlled by the action of this triode oscillator and the anode current, i.e. the current in the output circuit is therefore caused to vary in an oscillatory manner, although there is no direct coupling between the output circuit and the oscillator circuit. A marked reduction in the loading effect on the frequency determining circuit is thus obtained. A popular version of this circuit is shown in **Fig. 6.21.** The anode circuit may be untuned as shown at (A) or alternatively the desired output frequency, which may be the fundamental or a harmonic, may be selected by means of the tuned circuit as shown at (B).

It is possible to tune the anode of the oscillator of Fig. 6.21(B) to the second harmonic of the basic oscillator frequency but this is not advised. Under such conditions there will be a substantial increase in the flow of harmonic currents through the cathode of the valve and hence the fundamental tuned circuit. Such currents are likely to have a detrimental effect on the stability of the basic oscillator. A valuable property of the electron-coupled oscillator circuit is that the frequency variation for a given change in anode voltage is found to be of opposite sign to that caused by a similar change of screen voltage as shown in **Fig. 6.22.**

If the relative values of these two voltages are carefully chosen the effect of one electrode can be made to balance that of the other over quite a large range when the two voltages are supplied from a common source. An important feature to note is that a pentode which has its suppressor grid internally connected to the cathode is unsuitable for the e.c.o. type of circuit since the cathode is at r.f. potential to earth. If the same r.f. potential is allowed to appear on the suppressor grid, direct capacitive coupling between the oscillator section and the output anode circuit will exist. A valve with a separate connection to the suppressor grid must therefore be chosen so that it can be earthed.

The presence of oscillation is indicated by a rise in the anode current when the grid terminal is touched. When the circuit is oscillating, the feed current should be not less than half the non-oscillating value in the case of an untuned anode circuit. It can be altered by varying the gridleak value and the cathode tap position. The latter should be placed as near to the earth end of the coil as possible; usually about one-third of the total number of turns from this end is found to be a suitable position. A further refinement to this circuit is to tap the grid connection down the tuning inductance as shown in Fig. 6.17, thus reducing the loading thereon and improving the Q-value. The electron-coupled oscillator gives probably the highest output of all the v.f.o. circuits consistent with high stability and good tone quality. It can therefore be recommended safely to the newcomer to v.f.o. techniques.

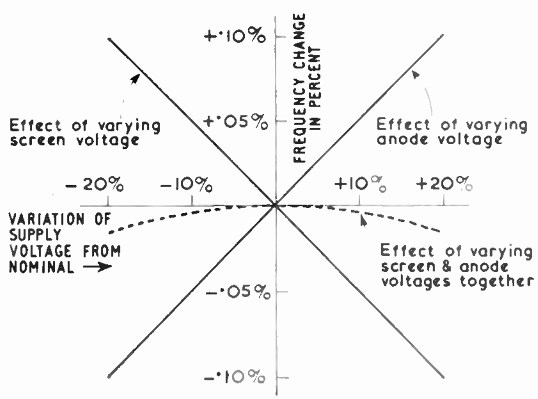


Fig. 6.22. Variation of frequency of an electron-coupled oscillator with electrode voltages. When the screen and anode voltages are varied simultaneously the effect on the frequency is very much less than that caused by varying either of them alone.

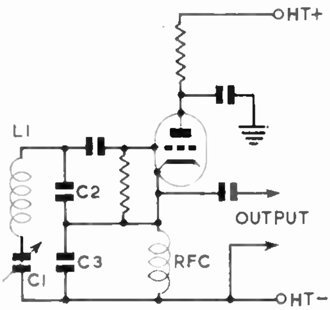


Fig. 6.23. Fundamental circuit of the Gouriet-Clapp oscillator, from which its similarity to the Colpitts oscillator may be seen.

The Gouriet Clapp Oscillator

The circuit of the Clapp oscillator, which was originally devised by G. C. Gouriet of the BBC, is shown in Fig. 6.23. It closely resembles that of the Colpitts oscillator. The important difference is that a small capacitor C1 is inserted in series with the tuning inductance L1 (which is made correspondingly larger so as to resonate at the required frequency), while the capacitors C2 and C3 constitute the feedback system and are of the order of 1000 pF each. Because they are connected across the valve electrodes, any capacitance variations within the valve itself become negligible in comparison. The whole arrangement may be compared with a crystal oscillator, since as previously explained the crystal has an equivalent circuit consisting of a very high inductance in series with a small capacitance in one arm in parallel with a second arm comprising a relatively large capacitance. The inductive reactance of the first arm in combination with the capacitive reactance of the second arm then causes the circuit to oscillate in the parallel-resonant mode. In a Clapp oscillator the first arm of the resonant circuit consists of L1 and C1 in series, while the second arm is the series combination of C2 and C3.

The tuning control is effected by the small variable capacitor C1. Satisfactory performance can only be obtained over a frequency range of about 1·2 : 1 because the power output diminishes rapidly as the series capacitance is reduced. This range is, however, more than enough to cover any of the amateur bands.

A practical circuit for use on the amateur bands is shown in Fig. 6.24. The triode oscillator portion comprises the cathode, grid and screen of a pentode and the output is derived from the tuned circuit connected to the anode proper which is electron-coupled to the oscillator section. For the highest frequency stability, the valve selected for use in this circuit should have the largest possible ratio between mutual conductance and inter-electrode capacity.

The ratio of the normally equal capacitances C1 and C2 to C3 plus C4 should be as high as possible consistent with reliable oscillation. If oscillation is not maintained over the desired range, C1 and C2 must be reduced in value or the valve should be substituted by one of higher mutual conductance: it may also help to increase the Q of the coil if possible.

The oscillator can be adapted for frequency modulation by connecting a reactance valve to the cathode or across C3.

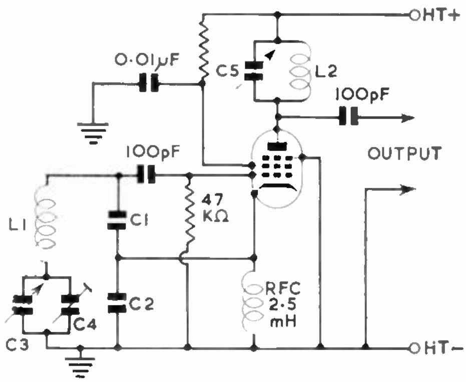


Fig. 6.24. A practical Gouriet-Clapp oscillator. For the 1·8 Mc/s band, typical values are: C1, C2, 2200 pF; C3, tuning, 50 pF; C4, bandset, 80 pF; C5, output tuning, 200 pF; L1, 70 μH; L2, 11 μH. If the oscillator is required to operate in the 3·5 Mc/s band, these values should be halved. Suitable valves are the EF91, Z77 or 6AM6. The h.t. voltage must be stabilized.

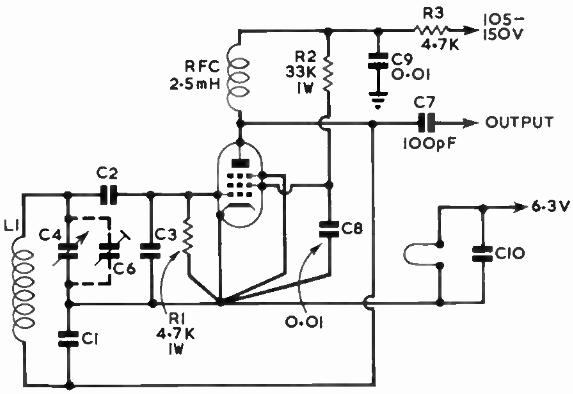


Fig. 6.25. Practical Vackar oscillator for frequencies in the range 1·5 Mc/s to 15 Mc/s. See Tables 6.4 and 6.5 for component details not given on the diagram. The h.t. voltage must be stabilized.

The Vackar Oscillator

Whereas the Clapp oscillator is limited in use to a relatively small frequency range, the Vackar (Tesla) oscillator which is a development of it, provides a fairly constant power output over a much wider range (about 2·5 : 1). It has similar

TABLE 6.4

Tuning Range	L1 S.w.g. enam.	L1 Turns close wound	C1	C2	C3	C4
Amateur Bands						
1·8– 2 Mc/s	34	70	556 pF	556 pF	4700 pF	15–250 pF
3·5– 3·8 Mc/s	28	45	500 pF	300 pF	2700 pF	10–100 pF
7·0– 7·1 Mc/s	26	30	200 pF	200 pF	1800 pF	10–25 pF
14 –14·35 Mc/s	24	15	100 pF	100 pF	1000 pF	10–35 pF
Special Frequencies						
8 Mc/s	26	25	200 pF	200 pF	1800 pF	+
9 Mc/s	26	20	200 pF	200 pF	1800 pF	+
10 Mc/s	24	25	140 pF	140 pF	1800 pF	+
11 Mc/s	24	20	140 pF	140 pF	1000 pF	+

All coils are wound on 3/16 in. diameter formers fitted with ferrite cores, the winding starting at the foot and progressing towards the top of the former. In the Special Frequencies section, no value is specified for the tuning capacitor C4 as this depends on the desired frequency coverage. A 1/4 in. former may be substituted if the number of turns is increased by 10 per cent.

TABLE 6.5

Tuning Range given by full excursion of core	L1 S.w.g. enam.	L1 Turns close wound	C1	C2	C3	C4
1·5– 2·5 Mc/s	34	70	556 pF	556 pF	4700 pF	
2·3– 3·3 Mc/s	34	45	556 pF	556 pF	4700 pF	
3·2– 4·5 Mc/s	28	45	500 pF	400 pF	2700 pF	
4·3– 6·3 Mc/s	28	35	300 pF	300 pF	2700 pF	
6·1– 8·8 Mc/s	26	30	200 pF	200 pF	1800 pF	
7·8–11·0 Mc/s	26	20	200 pF	200 pF	1800 pF	
10·5–15·0 Mc/s	24	20	100 pF	100 pF	1000 pF	

All coils are wound on $\frac{3}{16}$ in. diameter formers fitted with dust iron cores. No value is quoted for C4 as this will depend on the frequency coverage required. Since adding capacity at C4 will decrease the frequency, a coil is chosen which covers the highest frequency required within the range of its core, and the circuit tuned in a low frequency direction by a suitable capacitor at C4.

good frequency stability and low harmonic content.

The circuit adapted for amateur use is shown in **Fig. 6.25.** Operation at the fundamental frequency on all the h.f. bands is practicable. In the simplest arrangement tuning is effected by the variable capacitor C4, but a greatly improved performance is achieved if a split-stator capacitor is used, the other section being connected across C1. The capacitor C6 serves as a band-setting control. Typical coil and capacitor values are given in **Tables 6.4** and **6.5.**

As in the Clapp oscillator, maximum frequency stability is obtained by the selection of a valve with the highest possible ratio of mutual conductance to inter-electrode capacitances. A reactance modulator may be added for narrow-band frequency modulation. Cathode keying is found to be satisfactory although screen keying is usually superior.

The Franklin Oscillator

The basic circuit principle of the Franklin oscillator is shown in **Fig. 6.26.** Here two triodes V1 and V2 are connected as a conventional resistance-capacitance coupled amplifier. A rise of potential applied at the grid of V1 will appear on the grid of V2 as a fall in potential, and again at the anode of V2 as a rise in potential though now of course considerably greater than its original magnitude. If sufficient feedback is provided through C_B to the grid of V1 the whole amplifier will go into oscillation at a frequency determined by the natural time constant of the circuit which will depend mainly on the values of the resistances and the capacitances C_A and C_B. This is called a multi-vibrator.

To convert the arrangement into a constant-frequency

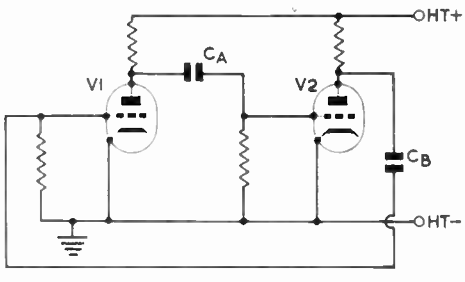


Fig. 6.26. Basic circuit to explain the operation of the Franklin oscillator. See text.

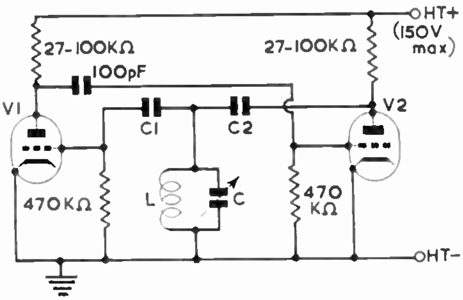


Fig. 6.27. A practical Franklin variable frequency oscillator. The tuned circuit is coupled to the valves by the capacitors C1 and C2, both of which have very small values, typically 1-5 pF depending on frequency. For optimum stability, C1 and C2 should be the lowest values needed to maintain adequate oscillation. Although separate triodes are shown, twin triodes such as the 12AT7 and the 12AU7 are satisfactory. The h.t. supply must be stabilized.

oscillator of controllable frequency, some means must be provided whereby the feedback or the loop gain is insufficient to maintain oscillation except at the desired frequency. This is done by inserting a parallel resonant circuit LC between the junction of C1 and C2 and earth in the feedback path from V2 to V1: see **Fig. 6.27.** This circuit has a low impedance except at its resonant frequency, and at any other frequency the feedback path is virtually short-circuited.

If C1 and C2 are reduced to a minimum possible value for maintaining oscillation, the frequency of oscillation will depend almost entirely on the natural frequency of the LC circuit. In practice suitable values are found to be about 1 pF, thus giving a high degree of isolation of the tuned circuit from the remainder of the circuit. Variations in valve capacitance can only slighly affect the tuning; further, the *Q* of the tuned circuit can be maintained at a high value since the load presented to it by the valve is small.

No tappings are required on the coil, and the tuned circuit is earthed at one end. Both of these features are of great constructional advantage.

The circuit may be used also as a crystal-controlled oscillator by substituting a crystal for the LC circuit. Owing to the high amplification available, oscillation is easily obtained with relatively inactive crystals at the frequency of parallel resonance.

The power output obtainable is quite low and extra amplification may be needed to provide adequate drive. Usually this type of oscillator will be found to have poor keying characteristics and the keying should therefore take place in one of the later stages.

To achieve the maximum frequency stability the values of C1 and C2 should be reduced as far as possible consistent with stable oscillation. The two capacitors should be varied together so as to be always approximately equal.

Mixer Oscillators

In a mixer type of v.f.o. the output voltage of a crystal oscillator is mixed with the output voltage of a variable-frequency oscillator of much lower frequency in a suitable valve mixer circuit. The sum or difference of these frequencies may be employed to control the transmitter frequency. The degree of frequency stability attainable by this method on any particular band is superior to that normally achieved by the use of the straightforward type of v.f.o. but the additional complexity of the mixer arrangement has prevented it from becoming widely adopted. The principle of its operation is shown in block diagram form in **Fig. 6.28.**

The reason for the improved stability is explained by considering the following example. Suppose that the crystal frequency (f_2) is 4·0 Mc/s and that the v.f.o. tunes from 200 to 500 kc/s (f_1). By extracting the difference frequency ($f_2 - f_1$) from the mixer by means of a frequency selector the output would cover the amateur band 3·5–3·8 Mc/s. If the v.f.o. is reasonably well designed its frequency should be constant within the limits of ± 6 c/s at 300 kc/s, which is only one part in 25,000; at the output from the mixer, i.e. in the range 3·5–3·8 Mc/s, this represents a stability of the order of one part in 300,000. For comparison, a typical v.f.o. of straightforward design having an output in the same band would probably have a stability of about one part in 50,000.

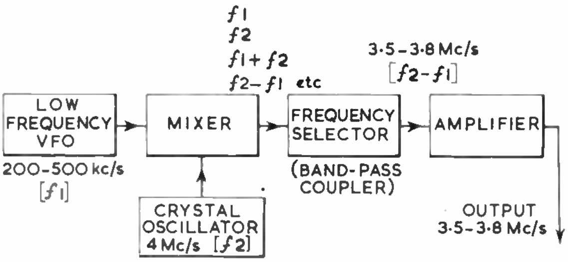


Fig. 6.28. Block diagram of a mixer-type v.f.o. The frequencies shown would be suitable for producing an output in the 3·5 Mc/s band.

In the example given, the stability multiplication factor is 12. This figure could be increased by using an even lower frequency for the v.f.o. but this would necessitate a crystal frequency nearer to the required output frequency, and a point is eventually reached at which it is difficult to separate them in the succeeding amplifier circuits. The best compromise must therefore be found. The figures given in the example should prove satisfactory for the 3·5 Mc/s band: a crystal frequency of 7·8 Mc/s in conjunction with a v.f.o. operating over a range of 500–800 kc/s should be satisfactory for the 7 Mc/s band.

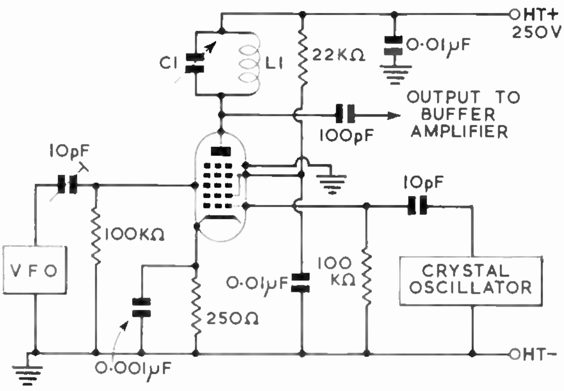


Fig. 6.29. Low-level mixing circuit for a v.f.o. showing typical values for a 6SA7. The values of L1 and C1 are selected as for a normal tank circuit.

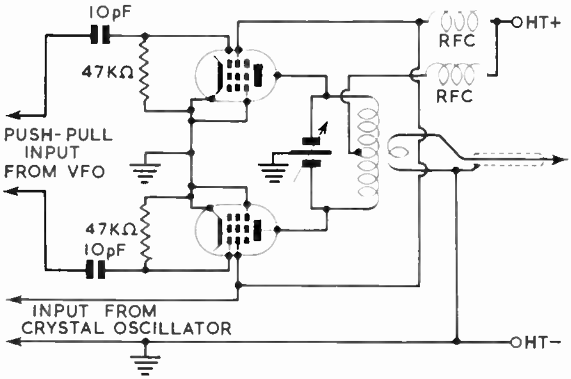


Fig. 6.30. Push-pull mixing circuit for a v.f.o. using screen injection of the crystal-oscillator voltage. The v.f.o. voltage is applied in push-pull either through a suitably tapped transformer arrangement or from a push-pull oscillator direct. The crystal-oscillator voltage is balanced out in the centre-tapped output tank circuit. Suitable valves are 6AK5 and 6BH6.

A low-level mixer circuit similar to that used in superheterodyne receivers is recommended: either a triode-hexode arrangement or a heptode such as the 6SA7 shown in **Fig. 6.29** is suitable. The anode circuit L1C1 is tuned to the centre of the required frequency band and is coupled to a tuned buffer amplifier. Any of the orthodox coupling methods may be used instead of the capacitive coupling indicated.

In another mixing arrangement, the crystal-oscillator voltage is applied in the same phase to the screen grids of a push-pull mixer while the v.f.o. voltage is applied in opposite phases to the control grids, as shown in **Fig. 6.30.** Alternatively heptode or hexode mixer valves may be used with the v.f.o. voltage applied in push-pull to the control grids and the crystal-oscillator voltage in parallel to the oscillator grids. The crystal-oscillator voltages are then balanced out in the output circuit, leaving only the v.f.o. frequency and the sum and difference frequencies. These are relatively easy to separate in the tuned anode circuit and in the following buffer amplifiers.

Ideal keying characteristics may be more easily obtained with this type of oscillator unit. Neither of the two frequencies initially generated falls within the amateur bands and thus the two oscillators may be left running continuously without fear of causing interference; it is only when mixing takes place that other frequencies are produced. If the mixer itself is suitably keyed, the output in the required band will be interrupted at a low power level and chirp can be avoided, provided of course that the effect of the changing load on the variable-frequency oscillator is negligible.

Phase Locked V.F.O.

The phase locked v.f.o. is relatively more complicated than those already described, but it has advantages in transmitters where the carrier frequency is high and large excursions in frequency are required. In the phase locked system, the basic arrangement of which is shown in block form in **Fig. 6.31**, a low frequency oscillator directly governs the frequency of a high frequency variable oscillator in such a manner that the stability of the high frequency variable oscillator is equal to the stability of the low frequency oscillator.

In many ways the control circuit of the phase locked v.f.o.

resembles the automatic frequency control system employed in some v.h.f. receivers to overcome drift in the local oscillator. There are of course differences since, in a receiver, the object is to maintain the oscillator locked on to a particular frequency, while in the phase locked v.f.o. the control system has to respond to intentional changes in the frequency of the carrier oscillator and then assume a new locked condition.

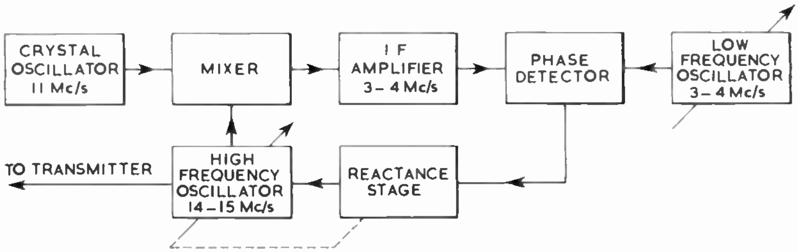


Fig. 6.31. Block diagram of a phase locked v.f.o. As the frequency of the l.f. oscillator is varied it affects the reactance valve which in turn tunes the h.f. oscillator. When the i.f. is equal to the l.f. oscillator, the system is said to be locked; under these conditions any change in the h.f. oscillator produces an output from the phase detector due to the difference between i.f. and the l.f. oscillator frequencies which, via the reactance stage, automatically corrects the h.f. oscillator.

In Fig. 6.31, a relatively high frequency variable frequency oscillator (usually operating at half the carrier frequency) has its actual operating frequency controlled by a reactance valve, while its output, in addition to being fed to the transmitter, is also fed to a mixer. A frequency from a crystal controlled source is also fed to this mixer with the result that an intermediate frequency is produced, the frequency range of which will be either the sum or the difference between the upper and lower limits of the high frequency oscillator and the crystal oscillator. This i.f. is fed to a phase detector, as is the output from a low frequency oscillator covering the same range as the i.f.

The basis of the system is the phase detector. Only when the two frequencies are precisely the same will the phase of their voltages be identical; under other conditions there will be a phase difference. The voltage from the i.f. amplifier and the low frequency oscillator are fed to the phase detector in such a manner that the output from the phase detector fed to the high frequency oscillator via a reactance stage adjusts the frequency of that oscillator until the output from the i.f. amplifier is exactly in phase with that of the low frequency oscillator, i.e. until the i.f. is of precisely the same frequency as that of the low frequency oscillator. Any drift in the high frequency oscillator will immediately reflect as a change in the intermediate frequency, and hence as a change in phase in the phase detector; this will in turn apply a correction signal to the reactance modulator to restore the h.f. oscillator frequency to that which produces an i.f. equal to the low frequency oscillator. Not only does the low frequency oscillator physically control the frequency of the high frequency oscillator, but in addition, the stability of the high frequency oscillator must be virtually that of the low frequency oscillator.

At the moment of switching on, there is almost bound to be a very considerable difference between the intermediate frequency and that of the low frequency v.f.o. In order to lock the frequency, the phase loop must have a wide bandwidth. Once the frequency is locked, however, a wide bandwidth is not needed. One way of overcoming the difficulties caused by the wide bandwidth is to arrange the circuit in such a manner than upon switching on the bandwidth is wide enough to ensure capture of a lock, and then, after a predetermined time, for the circuit automatically to reduce its bandwidth.

A particularly suitable valve for use as the phase detector is the 6BN6 gated beam valve, a circuit for which is given in **Fig. 6.32.**

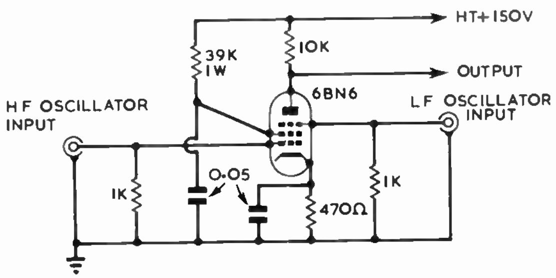


Fig. 6.32. Using the 6BN6 gated beam valve as a phase detector. Provided that the inputs are in excess of 2 volts, the stage will function correctly. A wideband isolating amplifier should be interposed between this phase detector and the reactance control valve.

SOLID STATE OSCILLATORS

Solid state oscillators possess some obvious advantages over their valve counterparts: smaller size, absence of heaters, good mechanical stability and long-term electrical stability. The transistor possesses certain characteristics, however, which must be taken into account when using it to maintain an oscillator circuit:

(a) The low input and output impedances.
(b) The gain characteristics are frequency dependant.
(c) The gain characteristics are also temperature dependant.
(d) The crystal or tuned circuit may be excessively damped due to the low values of biasing resistors which are necessary.
(e) The considerable spread in the characteristics of any given type of transistor makes individual bias adjustment advisable.
(f) Because of the very low thermal capacity of the transistor, it is subject to relatively rapid changes in temperature and must therefore be provided with an adequate heat sink.
(g) Transistors are electrically fragile and care must be taken not to exceed the maximum permitted voltages, currents, and temperatures.

The criteria applied to the oscillatory circuit, described

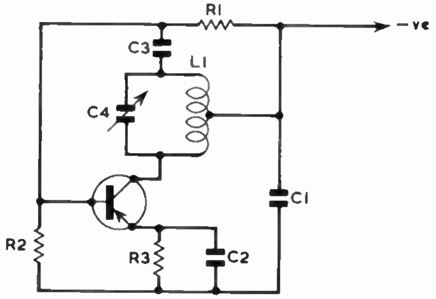


Fig. 6.33. Hartley transistor oscillator.

under valve circuits, apply equally to the solid state oscillator.

Most of the well-known valve oscillator circuits may be adapted to transistors, taking into account the difference in the parameters of the two devices. Examples of such circuits are shown in Figs. 6.33, 6.34, 6.35 and 6.36.

The circuit of a transistor bridge type v.f.o. is shown

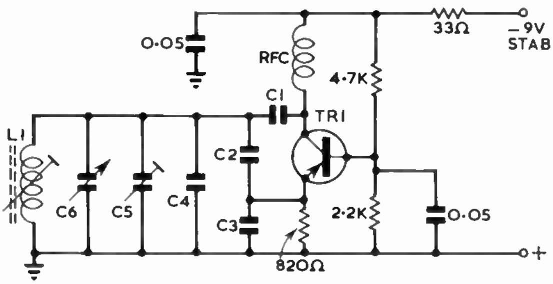


Fig. 6.34. High stability transistor high C Colpitts oscillator.

in Fig. 6.36. The attraction of this circuit is that external conditions do not affect the tuned circuit and, by carefully balancing the bridge, the effects of temperature and voltage changes on the transistors can be reduced to the point where they practically have no effect. Under such conditions, it is the reaction of the tuned circuit itself to temperature changes which will determine the ultimate stability.

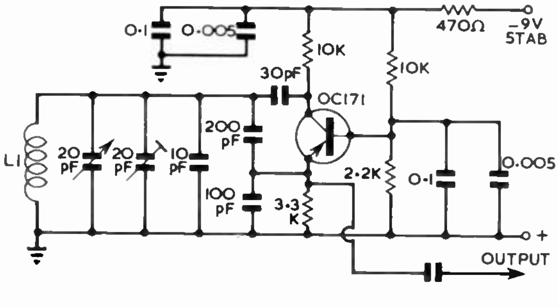


Fig. 6.35. Colpitts v.f.o. for 30 Mc/s. L1, 4½ turns of 16 s.w.g. enam. wire close wound on ⅜ in. diam. former fitted with ferrite core.

The resistors R3, in parallel with which is the series tuned circuit, R5, R6 and R4 form the bridge circuit which is connected between the input and output of a two stage amplifier. At the resonant frequency of the tuned circuit, R3 is virtually shorted out, and the bridge changes from a negative feedback condition to a positive feedback circuit, and thus oscillation occurs at the resonant frequency of the series tuned circuit. The feedback level may be adjusted by changes in the value of R5 or R6. Increasing R5, or decreasing R6, will result in a higher level of feedback.

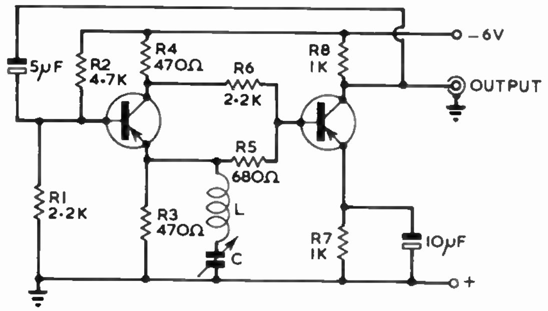


Fig. 6.36. Typical example of a bridge type transistor h.f. oscillator suitable for use as a v.f.o.

The "synthetic rock" circuit due to W3JHR (Fig. 6.37) illustrates the principle of employing a capactive divider network to match the transistor to the oscillatory circuit in order to prevent the low input impedance of the transistor damping the circuit excessively.

Adequate buffering is essential in the transistor oscillator, as in its valve counterpart, but the transistor buffer stage is limited in its effectiveness by virtue of its essentially low

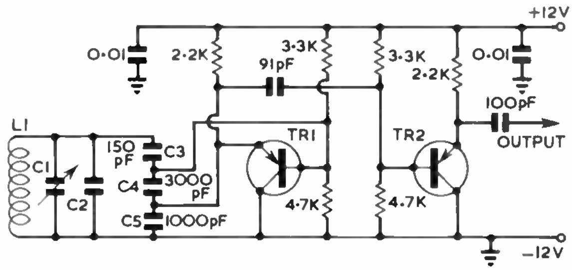


Fig. 6.37. The "synthetic rock" transistor v.f.o. designed by W3JHR.

input impedance and also by its "transparency"; that is, its input characteristics are effected by changes in output load.

With the advent of the field effect transistor, all the advantages of a solid state device, combined with characteristics directly comparable with the valve are available. The FET is therefore the ideal device for oscillator applications: it has high input impedance, it is much less transparent than the bipolar transistor and it may therefore replace valves in conventional oscillator circuits, taking account of the operating voltages. See Fig. 6.38. For the same reasons FETs make excellent buffer amplifiers.

The FET, although exhibiting pentode drain/gate (anode/

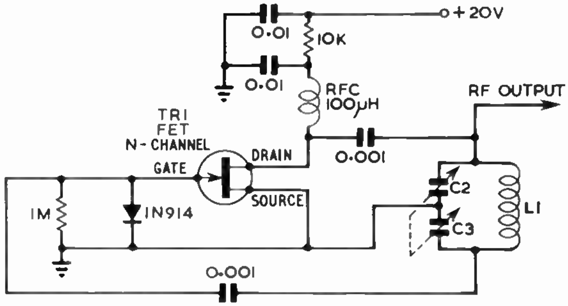


Fig. 6.38. A Solid state oscillator employing an FET.

grid) characteristics, does not have the effect of the screen grid for input/output capacitive isolation and thus the drain/gate (anode/grid) capacitance is comparable with that of a triode valve. In all straight amplifier applications the FET must be neutralized (Chapter 3—*Semiconductors*).

The FET possesses a convenient square law relationship between gate voltage and drain current, thus giving the device a linear relationship between gate voltage and g_m. This permits amplitude control of the oscillator by rectifying the output and feeding it back to control the gate bias voltage, thus allowing class A operation with increased stability due to reduction in the harmonic currents flowing in the device. It should be noted that an IGFET cannot draw gate current corresponding to grid current in a valve. The gate voltage must be limited by a diode suitably connected between gate and source as shown in the above diagram.

INTERSTAGE COUPLING METHODS

Correct coupling between two stages is obtained when the input impedance of the driven stage is equal to the output impedance of the driver stage, resulting in the maximum transfer of power. The required condition can be secured in several ways, each of which has its own special advantages that make it more suitable for particular applications.

Capacitive Coupling

Various arrangements of capacitive coupling are shown in **Fig. 6.39.** In its simplest form, shown in (A), a capacitor C3 is connected from the anode of the driver valve V1 to the grid of the next stage V2 in order to transfer the r.f. voltage from the driver tank circuit L1C1. The capacitor C4 shown dotted, represents the grid-to-earth capacitance of V2 while R2 represents the equivalent resistance of the driven grid circuit. Both C4 and R2 are effectively in parallel with the driver tank circuit L1C1 through the r.f. bypass capacitor C2 across the h.t. supply and if the coupling capacitor C3 is made too great, V1 may be incorrectly loaded. As the load is increased, the Q of the tank circuit falls and a point is eventually reached at which the efficiency of the driver is lowered; moreover harmonics are then produced and may give rise to interference.

The effect of C4 is likewise to reduce the L/C ratio in the anode tank circuit of V1 and if it is large it must be allowed for in the design. A method of improving the L/C ratio of the tank circuit is to tap the grid of V2 into the coil as shown in (B); the Q of the tank circuit is also increased and often the impedance matching between stages can be achieved more accurately with consequent greater power transfer. Besides adding to the mechanical complexity, however, the tapped-coil arrangement shows a tendency to generate parasitic oscillations.

The resistor R1, which is the d.c. grid return for V2, is in effect also in parallel with the tank circuit of V1 and if it were not for the r.f. choke (RFC) in series with it this resistance would present an extra (though small) r.f. load. The r.f. choke reduces the power loss in the resistor by reducing the r.f. current flowing through it, while the d.c. path remains unchanged.

As shown in (C), a single-ended driver valve can be

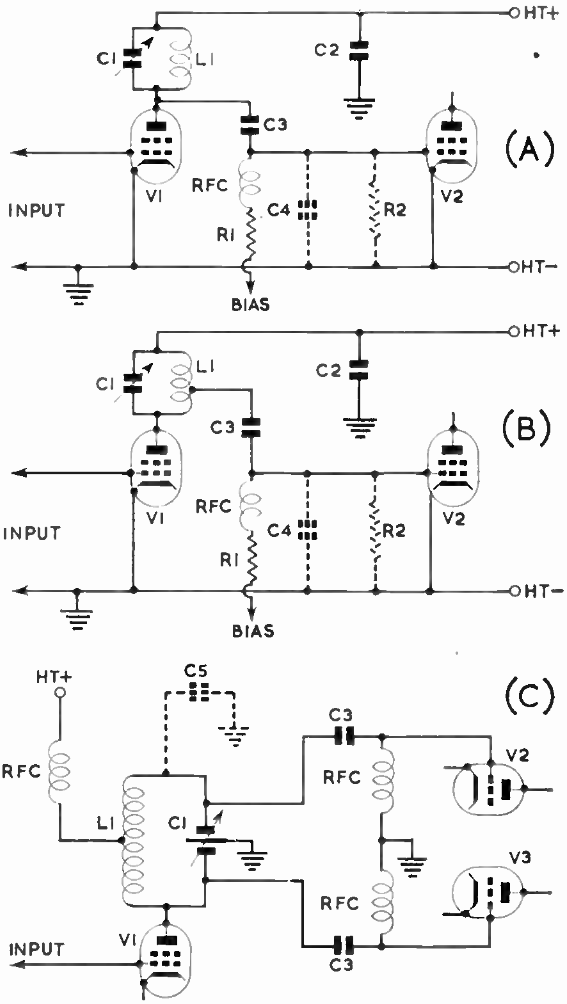


Fig. 6.39. (A) Simplest capacity coupling method which usually results in a high degree of harmonic transfer from V1 to V2. With this circuit accurate matching between V1 and V2 is difficult to obtain. (B) Provided the impedance of the anode of V1 is higher than that of the grid of V2, tapping the coupling capacitor down L1 may permit an acceptable match to be secured. (C) The circuit of (A) re-arranged to drive valves in push-pull. This circuit has the same general characteristics as that shown in (A). To obtain equal drive, the capacitor C5 may be needed to balance the anode to earth capacity of V1. The grid circuits of V2 and V3 are symbolic only, no biasing arrangements being shown.

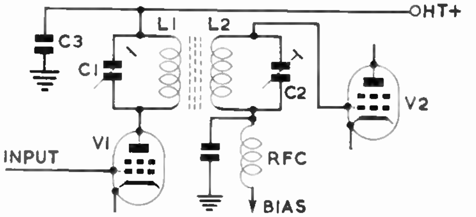


Fig. 6.40. Inductive coupling. A band-pass effect can be obtained by suitably close coupling between L1 and L2. L1C1—anode tank circuit of driver stage; L2C2—grid tank circuit of driven stage; C3—r.f. by-pass capacitor.

coupled to a push-pull stage V2 and V3. In this case, the tank coil of V1 is centre-tapped and is tuned by the split-stator capacitor C1. The r.f. voltages at the ends of the coil are equal in amplitude and 180 degrees out-of-phase and are therefore suitable for feeding directly to the grids of V2 and V3. The rotor of the split-stator capacitor must be earthed by the shortest route possible, preferably direct on to the chassis. If this is not done, the inductance of the connection may be sufficiently great to prevent harmonics from being adequately bypassed to earth and interference through the inadvertent radiation of these harmonics may result.

Inductive Coupling

Inductive coupling between stages has several distinct advantages as compared with capacitive coupling. The basic circuit arrangement is shown in Fig. 6.40.

To understand these advantages, consider the two tuned circuits L1C1 (primary) and L2C2 (secondary), loosely coupled by a small degree of mutual induction between the coils, and suppose that each of the circuits resonates separately at a frequency *f*. If an r.f. voltage of variable frequency is applied to the primary of this arrangement and the resulting output voltage across the secondary is measured, the variation of output voltage with the frequency of the applied voltage will be in the form shown in Fig. 6.41, curve A: i.e., the circuit would possess considerable selectivity with a single peak at the frequency *f*. If the coils are now brought closer together, thus increasing the coupling between windings, the curve obtained on repeating the measurement will become flatter and the output voltage at the peak will be slightly higher, as depicted in curve B. Further tightening of the coupling causes the response to become still flatter until, after a certain critical value of the coupling has been reached (curve C), two peaks separated by equal amounts on opposite sides of the central frequency appear, with a pronounced dip at the centre (curve D). The value of coupling just below that which causes the appearance of the two peaks is known as the *critical coupling*. When the circuits are in this condition, it is found that the output voltage remains virtually constant over quite a wide frequency range, often as much as 5 per cent of the central frequency, and also that on either side of the flat top of the curve the voltage falls away somewhat more rapidly than in the single-peaked curve obtained with looser coupling (such as curve A or curve B). A pair of resonant circuits having critical coupling, and therefore exhibiting a flat-topped characteristic, are referred to as a *bandpass* or *wideband* coupling.

The bandwidth (i.e. the frequency range over which the output voltage remains within specified limits, usually 3db) of such a critically coupled arrangement is about 1·4 times that of a single tuned-circuit coupling. The overall bandwidth of a series of stages coupled by double-tuned circuits, as might be used in a drive unit, narrows much less rapidly as the number of stages is increased, compared with single tuned-circuit coupling. The same applies to double-tuned circuits irrespective of whether they are coupled by mutual inductance, common inductance or capacitance, or by link coupling as described in the following section.

The practical advantage gained therefore by the use of a properly adjusted bandpass system is that after the initial setting up of the circuits, the drive to the final stage can be made to remain virtually constant over the width of an amateur band without the necessity of any tuning adjustments.

The circuit shown in Fig. 6.40 is a simple practical bandpass coupling arrangement. To avoid capacitive coupling effects which would modify the bandpass characteristic in an undesirable and unpredictable way, the grid end of the secondary winding should be situated at the opposite end from the anode end of the primary. It is quite a simple matter to measure the frequency-response characteristics (as typified by Fig. 6.41) merely by using the v.f.o. as the source of variable-frequency input voltage and observing the variations in the grid current of the driven stage. The voltage produced by the v.f.o. (or by an untuned buffer stage following it) can be assumed to be reasonably constant over a limited frequency range. The correct degree of inductive coupling can then be found from a progressive series of measurements. If the bandwidth obtained with critical coupling is found to be too small, it is usually possible to broaden it by increasing the L/C ratio of the tuned circuits.

Sometimes it can happen that at first only one voltage peak is observed, although in fact the over-coupled case shown in curve D is being obtained. The other peak may be lying outside the range of measurement.

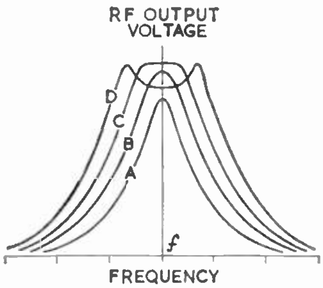


Fig. 6.41. Variation of output voltage with frequency for the inductively coupled circuits shown in Fig. 6.40. The curve C indicates the conditions known as "critical coupling" the output voltage being constant over approximately a 5 per cent variation in frequency.

Bandpass Couplers

The use of bandpass couplers in a transmitter not only reduces the possibility of radiating spurious signals, but brings about a worthwhile reduction in the number of controls required for the r.f. driver section. While such couplers have these advantages, their use does invariably require somewhat higher power to be developed in the driver stages than that needed in the case of single tuned circuits. This however is a small price to pay for the increase in operating convenience.

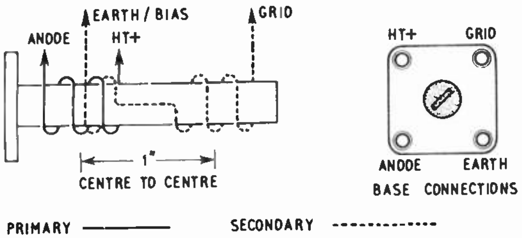


Fig. 6.42. General construction of wideband couplers for the amateur bands 1·8 Mc/s-28 Mc/s. Standard $\frac{7}{16}$ in. diameter formers are employed complete with screening cans. Both windings are in the same sense. Further details are given in Table 6.6.

The construction of a bandpass coupler is shown in **Fig. 6.42** and winding details in **Table 6.6.** No additional capacity is required across the windings except on 1·8 and 3·5 Mc/s. The capacity provided by the valves is sufficient.

The couplers are constructed on standard $\frac{7}{16}$ in. diameter i.f. former assemblies fitted with screening cans. To assist in the winding, a length of adhesive paper has fitted to it a narrower piece in such a manner that part of the length of the narrower strip is in contact with the adhesive of the wider strip. The combined pieces are then wrapped round the former with the adhesive of the narrower piece in contact with the former, and the adhesive of the wider section exposed. The winding is then made over the exposed adhesive. On the three lower frequency bands the windings are layer wound. Once one layer has been completed, a layer of tape should be wound over it before starting the next layer.

The actual winding should commence with the primary. When this is completed, the secondary should be started with the link over the primary first, and then progress on to the main secondary winding. Four 18 s.w.g. tinned copper spills should be affixed, one to each terminal eyelet, and the windings terminated as indicated in the diagram.

As it may be necessary to adjust the position of the link winding on the 28–30 Mc/s coupler in order to achieve a level response, it should not be too tightly wound over the primary in the first instance.

While these couplers are primarily intended for use with 6BW6 valves, there is no reason why they should not be employed with similar types so long as the mutual conductance and the grid anode capacities of the substitute do not differ very greatly.

In the actual alignment, the primary (lower) core should be adjusted to give the required drive at the lower end of the band, and the secondary (upper) core for the same value of drive at the high frequency end of the band. The adjustments will tend to interact to some extent, and it will be necessary to go back and forth to arrive at a condition where the drive is reasonably level from one end of the band to the other. If it is suspected that a particular winding cannot be tuned sufficiently in an h.f. direction a brass core can be inserted in place of the iron core to confirm that there is excessive inductance. In some cases, it may be found that the use of a brass core will provide the required tuning range, and that modification to the winding is not necessary.

These couplers will provide more than adequate drive from a 6BW6 driver stage to a 6146 p.a. either single ended or push-pull. On the lower frequencies it may be necessary to fit a variable potentiometer in the screen supply of the driver valve in order to reduce the grid drive to the correct value.

TABLE 6.6

Amateur Band	Number of Turns Primary	Link	Secondary	Wire	Capacity
1·8 Mc/s	65 + 60 + 55	16	60 + 55 + 50	40 s.w.g.	10/15 pF
3·5 Mc/s	33 + 32 + 31	9	30 + 29 + 28	32 s.w.g.	5 pF
7 Mc/s	25 + 25	5	24 + 22	32 s.w.g.	Nil
14 Mc/s	24	3	24	32 s.w.g.	Nil
21 Mc/s	18	1	18	28 s.w.g.	Nil
28 Mc/s	13	2	12	28 s.w.g.	Nil

The primaries and secondaries of the three lower frequency ranges are layer wound. All wires are enamelled. Each winding is fitted with a dust iron core.

Link Coupling

The practical disadvantage of simple bandpass inductive coupling between a pair of tuned circuits is that it may be difficult to arrange for the necessary mechanical adjustment. This can be overcome by using a variable link coupling between the two coils.

A small coil having a very few turns of wire is coupled to the driver tank circuit, thus transforming the output from a high impedance to a low impedance. A flexible transmission

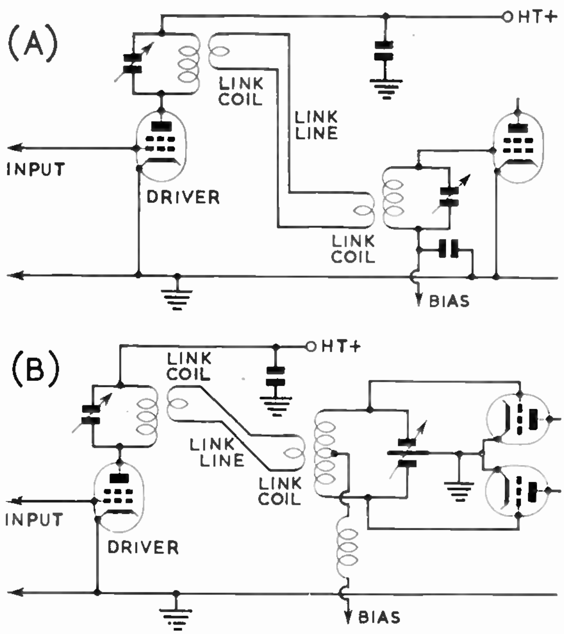


Fig. 6.43. Link coupling. The arrangement at (A) is used for coupling two single-ended stages, while (B) shows how a single-ended driver can be coupled to a push-pull amplifier. The link coils must always be placed at the "cold" end of the tuned-circuit inductances to avoid unwanted capacitive coupling. Suitable numbers of turns for the link coils are as follows:

Frequency (Mc/s)	1·8	3·5	7	14	21	28
Number of turns	6-8	4-5	3	2	1	1

In certain circumstances it may be desirable to earth one side of the link line.

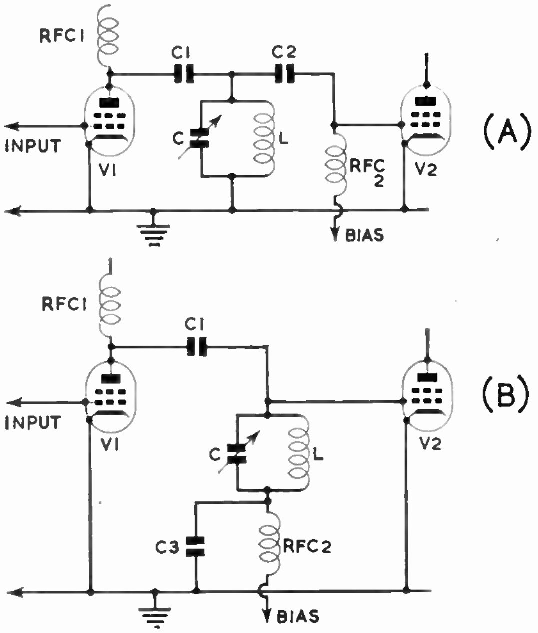


Fig. 6.44. Parallel anode feed in capacitance-coupled amplifiers. In these arrangements the resonant circuit LC is the anode load. In (A) the bias for V2 is applied through the grid choke RFC2 whereas in (B) it is applied in series with the tuned circuit. C1, isolating coupling capacitor (0·001-0·01 μF); C2, inter-stage coupling capacitor (100 pF); C3, bypass capacitor (0·001-0·01 μF); LC, tank circuit of driver stage V1.

line transfers the energy at this low impedance to another similar coil coupled to the tuned input circuit of the driven valve where it is again transformed to the high impedance value required by its grid: see Fig. 6.43. The degree of coupling is adjusted by altering the position of either of the link coils or by altering the number of turns in them.

The link coils should be situated at the " cold " or " earthy " ends (i.e. the ends having zero r.f. potential) of the tank coils since the capacitance coupling which exists between the link and the tuned circuits will otherwise result in spurious tuning effects and the transference of unwanted harmonics. It is also beneficial to earth one side of the link circuit because this eliminates the possibility of stray capacitive coupling between the tuned circuits through the link.

The link line may in theory be of any length, but in practice, unless it is limited to about 2 ft. the resulting reactance presented by the link may make adequate coupling difficult. However, if the line is initially adjusted to have a " flat " characteristic, this difficulty should not arise. The link reactance may also be tuned out by making the link circuit resonant; this is usually accomplished by the addition of capacitance in parallel or in series with one of the link coils. The method is not always satisfactory since the correct adjustment varies with the operating frequency and the losses in the line and the radiation from it are increased. A further advantage of link coupling is that it simplifies the problem of coupling a single-ended stage to a push-pull (or push-push) amplifier, as shown at (B) in Fig. 6.43. Balanced low-impedance line, such as flat-twin cable of 75-ohms impedance, is suitable for this purpose; co-axial cable (outer braid earthed) can also be used and may be desirable in order to reduce radiation from the line.

Parallel Anode Feed

If desired, the d.c. anode voltage of the driver valve can be removed from the associated coupling components in any of the coupling arrangements so far described. The basic principles are shown in Figs. 6.44 and 6.45. The choke RFC1 presents a high impedance over a reasonably wide range of frequencies including the operating frequencies, and the necessary r.f. voltage is developed across the tuned circuit to which it is coupled through the isolating capacitor C1.

Apart from the increased factor of safety achieved by removing the anode voltage from the tuned circuits, the peak voltage across the tuning capacitor is reduced and one of smaller voltage rating may therefore be used—a matter of some importance in high-power driver stages or power amplifiers. The capacitor C1 should be rated to withstand the d.c. anode voltage plus the peak r.f. voltage and should preferably have a mica dielectric.

R.F. Chokes

As mentioned above, the r.f. choke must present a high impedance at the operating frequency. If the latter is variable over a very wide range as in an multi-band transmitter covering 1·8–29·7 Mc/s appreciable losses may occur in the choke owing to the effects of self-capacitance and perhaps resonance. The difficulty may be partially overcome by the use of two or more different types of chokes connected

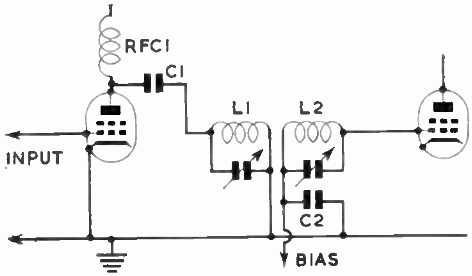


Fig. 6.45. Parallel anode feed in an inductively-coupled amplifier. C1, isolating capacitor (0·001-0·01 μF); C2, bypass capacitor (0·001-0·01 μF); L1, L2, mutually coupled coils.

in series. A good compromise for an all-band choke is one having an inductance in the region of 2·5 mH, but its self-capacitance should be low.

Where r.f. chokes are present in both the grid and the anode circuits of a valve amplifier, it is advisable to make them of different inductance values since otherwise there is a possibility of low-frequency parasitic oscillations being generated by feedback in the tuned-anode/tuned-grid circuit thus formed.

Interstage Pi-Network Coupler

The pi-network coupler is often used for delivering power to an aerial or an aerial feeder but it is equally suitable for coupling together two intermediate stages as shown in

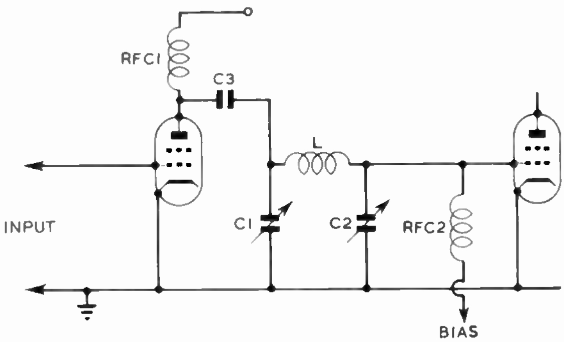


Fig. 6.46. Inter-stage pi-network coupler. By varying C1 and C2 the impedances of the two stages can be matched: see text for design data. The blocking capacitor may be 0·001-0·01 μF.

Fig. 6.46. The impedances of the two stages can be matched by adjusting the tuning capacitors C1 and C2. The properties of pi-network couplers are further discussed on page 6.41 and the method of finding the coil and capacitor values is given. When used as a matching device between two class C stages the following values may be taken:

Load required by driven stage—

$$R_1 = \frac{(0{\cdot}57 \times V_{ht})^2}{P}$$

where V_{ht} = d.c. anode voltage
P = output power in watts (input power to driven stage).

Input resistance of the driven stage

$$R_2 = \frac{6{\cdot}22 \times P}{I_g^2} \times 10^5$$

where P = total power to grid circuit of the driven stage
I_g = d.c. grid current in milliamps.

The total power P to the grid circuit depends upon the biasing arrangement used. For example, if the bias is derived by current through a grid resistor and this resistor is not isolated from the grid by an r.f. choke then it must be added in parallel with the resistance R_2 derived above. In addition, the value of P used in the formula is not only the driving power required for the valve alone since the driver has to supply power for the bias circuit itself. Where the bias is derived from the flow of grid current through a grid resistor the power loss in the circuit will be I_g^2R watts where I_g is the d.c. current through the resistor and R_g the value of the resistor in ohms. If the grid resistor is isolated from the grid by an r.f. choke the shunting effect can be ignored but the bias power $I_g^2 R_g$ must still be included in P in the formula.

In the case of fixed bias obtained from a source isolated from the grid by a choke or tuned circuit, the power loss in the bias circuit is $I_g V_g$ where I_g is the d.c. grid current and V_g the bias voltage. Again there is no shunting effect.

The formula for R_1 may be expressed in terms of anode supply voltage and anode current. If the efficiency is assumed to be 70 per cent when operating at the fundamental, 50 per cent when doubling and 30 per cent when tripling the expressions are:

Fundamental

$$\frac{(0{\cdot}57 \times V_{ht})^2}{0{\cdot}7\dfrac{(V_{ht} \times I_a)}{1000}} = 460\frac{V_{ht}}{I_a}\text{ ohms}$$

Second Harmonic

$$\frac{(0{\cdot}57 \times V_{ht})^2}{0{\cdot}5\dfrac{(V_{ht} \times I_a)}{1000}} = 650\frac{V_{ht}}{I_a}\text{ ohms}$$

Third Harmonic

$$\frac{(0{\cdot}57 \times V_{ht})^2}{0{\cdot}3\dfrac{(V_{ht} \times I_a)}{1000}} = 1080\frac{V_{ht}}{I_a}\text{ ohms}$$

The pi-network coupler also acts as a low-pass filter and therefore helps to prevent the transfer of undesirable harmonics from the driver to the following stages. In fact, the harmonic output from a transmitter stage using such a coupler is less than that produced by a normal tank circuit of the same Q by a factor of $1/n^2$ where n is the order of the harmonic.

Transistor Interstage Coupling

The methods of coupling employed in valve circuits may also be used with transistors, although greater care must be taken in matching. For example the collector impedance of a transistor working in a common emitter mode is substantially lower than that of a thermionic valve anode. To attempt to connect it directly to a parallel tuned circuit would load the circuit so heavily that no appreciable Q could be realized. When a transistor is employed in a common emitter configuration, the collector has therefore to be connected to a tap on the tuned circuit at a point corresponding to the output impedance of the transistor—see Fig. 6.47(A). When a transistor is operated in the common base mode (signal applied to the emitter) the collector impedance is high, and adequate Q can be achieved without the need for a tap on the coil of the tuned circuit—see Fig. 6.47(B).

From a functional point of view, there are two differences between common emitter and common base modes of operation. The common emitter configuration is a current amplifier, whereas the common base configuration is a voltage amplifier. In addition, the degree of isolation between the input and output is considerably higher in the common base mode, from which it follows that it will perform satisfactorily at higher frequencies. The common

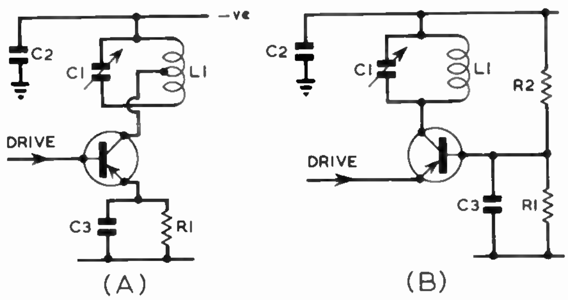


Fig. 6.47. Collector to tuned circuit matching. In the common emitter mode of A, the output impedance is low and the collector is tapped down the tuned circuit at a point equal to its impedance. In the common base mode shown in B the output impedance is high and no tap is required on the coil of the tuned circuit.

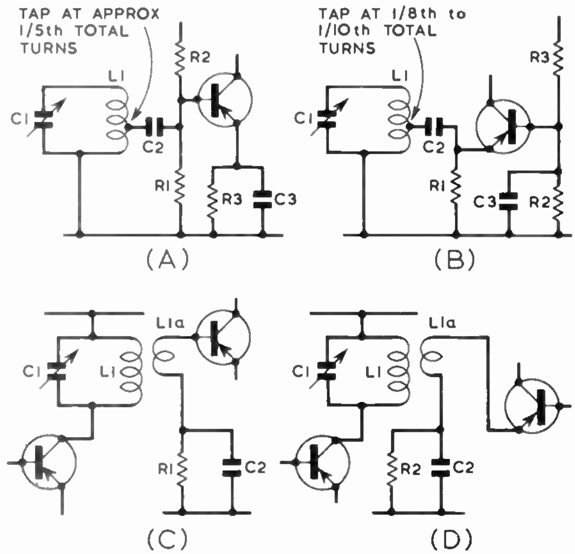


Fig. 6.48. Transistor input coupling. Diagrams A and B show the base and emitter fed from taps on the tuned circuit. The preferred methods are those shown in C and D since they both permit easier adjustment of the drive and the matching.

emitter arrangement is similar to a conventional triode amplifier, while the common base arrangement is analogous to a grounded grid triode stage. As the common base mode has a high output impedance and will produce reasonable levels of peak-to-peak voltage, it is commonly used in hybrid designs in which it is required to couple from transistors to valves since it permits either link or direct capacity coupling to be employed.

The input of the common emitter circuit has a somewhat higher impedance than the input of the common base configuration, although both are low and so require suitable tapping points on the circuits feeding them—see **Fig. 6.48(A)** and **(B)**. In place of a direct tap on a tuned circuit, it is more usual in transmitter practice to feed the base or the emitter from a suitably proportioned link winding coupled to the circuit in the collector of the preceding transistor (**Fig. 6.48(C)** and **(D)**). Even with mutually coupled circuits including wideband couplers, this has certain advantages. In the power stages of transistor transmitters it is usual to use pi-coupling.

Transistor interstage amplifiers in transmitters normally function in class B or class C, the actual class of operation depending on the drive and the voltage developed across the resistor R1 in Fig. 6.48. Some forward bias may be used as shown in Fig. 6.48(A) and (B) but this is not usual unless class A operation is required when the level of the drive has to be reduced accordingly.

FREQUENCY MULTIPLIERS

A power output at a multiple of the generated frequency is often required in amateur transmitters to obtain, for example, multi-band operation or merely to gain the advantage of the higher frequency stability which results from operating the oscillator on a relatively low frequency.

Frequency multiplication is made possible by the distortion of the anode-current r.f. waveform which occurs in a class C amplifier stage. As described on page 6.25, the current in such an amplifier flows only for a short time during the cycle in the form of pulses. It can be shown that such a series of pulses is composed of a number of sine waves of frequencies, f, $2f$, $3f$ and so on (where f is the fundamental or input frequency), the amplitude of these *harmonics* as they are called gradually diminishing with increasing order. Their relative amplitudes depend upon the shape of the pulse, which is determined by the operating conditions of the valve. A typical frequency multiplier circuit is shown in **Fig. 6.49.**

By inserting a parallel-tuned circuit in the anode circuit of the valve, resonant at the required harmonic frequency and therefore offering a high impedance at this frequency, the passage of the harmonic current produces a corresponding voltage-drop across the circuit. The tuned circuit in this way acts as a selector for the appropriate frequency.

In order to increase the energy of the harmonic current contained within the pulse, and therefore the efficiency of the valve as a frequency-multiplying device, it is desirable to reduce the angle of flow of current (defined on page 6.25) to values somewhat below the normal class C amplifier levels, e.g. 90 degrees or less. This is accomplished by increasing the grid bias to beyond normal class C operation and driving the valve with a high r.f. grid voltage.

The maximum power output obtainable at various harmonic frequencies in a frequency-multiplier stage, relative to the output at the fundamental frequency are—

Second harmonic	..	55 per cent
Third harmonic	..	35 per cent
Fourth harmonic	..	25 per cent

The fourth harmonic is the maximum degree of multiplication usually attempted in one stage, although for low output-power requirements a frequency multiplication of up to seven or eight is occasionally used.

The increase of harmonic output power obtained by this method may be found to aggravate the problem of avoiding interference. In general a frequency-multiplying stage should be operated at the lowest possible power level and preferably it should be followed by a further buffer amplifying stage before being used to drive the final power amplifier.

Neutralization of a frequency-multiplier is not required

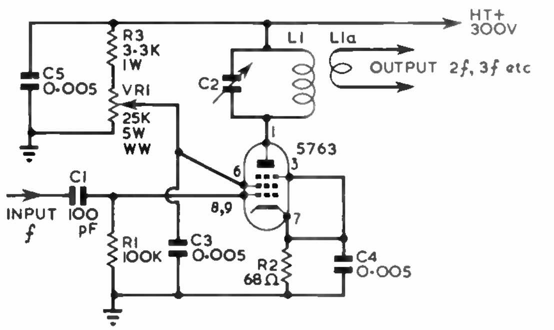


Fig. 6.49. Typical single ended frequency multiplier. The anode circuit should resonate at the desired multiple of the input frequency f. This circuit illustrates the method by which varying the screen potential varies the drive to the following stage. Primary bias is developed by grid current through R1. R2 is fitted to give protective bias in the event of drive failure.

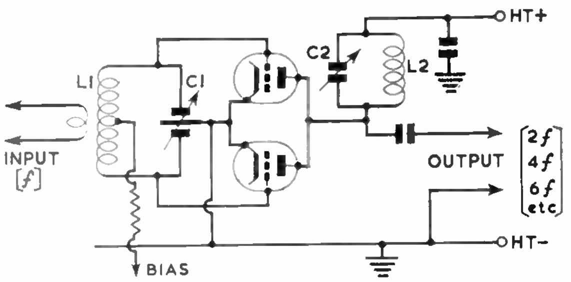


Fig. 6.50. (a) A push-push requency multiplier. Output from this circuit is only obtained on even order harmonics, i.e. second, fourth, etc.

even with triodes since the grid and anode circuits are tuned to different frequencies, and self-oscillation is not possible.

Push-push Doubler

To improve the efficiency of frequency multiplication at the *even* harmonic frequencies a push-push doubler is often used: see **Fig. 6.50 (a)**. The anodes of the two valves are connected in parallel to the normal tank circuit, whereas the grids are driven in push-pull. The tuned circuit thus receives pulses of current alternately from the two anodes and the harmonic energy supplied to the tuned circuit is thereby doubled. When operated as a doubler, such a stage is capable of giving efficiencies approaching that of a straight class C amplifier; at the higher *even* harmonics the output is correspondingly increased.

Push-pull Frequency-Multipliers

To increase the efficiency of a multiplier stage at *odd* harmonics a true push-pull circuit such as is shown in **Fig. 6.50 (b)** may be used, both the grids and the anodes being connected in push-pull. In this case, the even harmonics are self-cancelling in the anode circuit and the odd harmonics are reinforced.

Design Features and Adjustments

Although triodes may be suitable for frequency multiplication, the drive-power requirement for tetrodes or pentodes is much lower. In many amateur transmitters a drive unit consisting of a number of successive pentode frequency-multiplying stages preceded by a v.f.o. or crystal oscillator are used. All these stages are operated at low power to reduce the possibility of TVI and other forms of interference. The output of any of these multipler stages may be selected at will to provide drive for the power amplifier on the desired frequency band.

In modern designs, the coupling between stages is in the form of wide-band couplers as described in the previous section which give a notable reduction in the transference of undesired harmonics. Furthermore, once the circuits have been initially tuned no further adjustments are necessary.

The first step in setting up a frequency-multiplying stage is to reduce its h.t. supply voltage. The drive can then be applied and the grid circuit adjusted to resonance (as indicated by a maximum reading on the grid-current meter). On varying the tuning of the anode circuit one or more dips in the value of the anode current will be noticed, corresponding to resonance at the various harmonic frequencies. The dips become progressively less deep as the order of harmonic selected is increased.

The possibility of tuning to an incorrect harmonic should be borne in mind and it is advisable to check the frequency with a suitable absorption wavemeter coupled to the tank circuit; alternatively the effect of the absorption wavemeter on the anode current of the multiplier at resonance can be observed. The current should vary as the wavemeter is tuned through the harmonic frequency.

POWER AMPLIFIERS

As already explained, the generation of r.f. power must be at a low level to ensure complete stability. The power level is raised to the desired value by means of r.f. amplifiers. The amplification must take place without the introduction of any spurious or harmonic frequencies which if radiated might cause undue interference to other services. According to the terms of the Amateur (Sound) Licences the maximum d.c. power input to the anode circuit of the final amplifier must be limited to certain specified values, depending on the frequency band in use and other factors, and therefore in order to radiate the greatest possible power for the limited input the anode-circuit efficiency of the amplifier (i.e. conversion of d.c. power to r.f. power) must be as high as possible. For this reason and also on the grounds of economy, the two classes of r.f. power amplifier in common use for c.w. and a.m. operation are the class B and class C types described below.

Class C Operation

In a class C amplifier the standing grid bias voltage is increased to approximately twice the value required to cut off the anode current in the absence of r.f. drive to the grid. The r.f. drive voltage must be of sufficient amplitude to swing the grid potential so far that it becomes positive with respect to the cathode during the positive half-cycle. The resultant anode current is thus in the form of a succession of pulses, flowing only during a part of the r.f. drive cycle; the " angle of flow " (which measures its duration) is usually about 120 electrical degrees in a class C amplifier, i.e. about one-third of a full cycle (360 degrees), whereas for comparison the angle of flow in a class B amplifier is 180 degrees.

Fig. 6.51 shows the conditions existing in a class C amplifier. The broken horizontal line AB at about one-third of the peak pulse height represents the mean anode current during

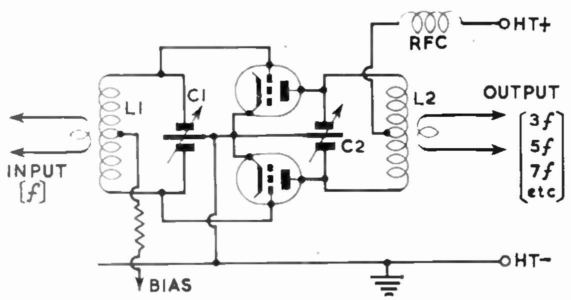


Fig. 6.50(b). A push-pull frequency multiplier. Output from this circuit is only obtained on odd order harmonics, i.e., third, fifth, etc.

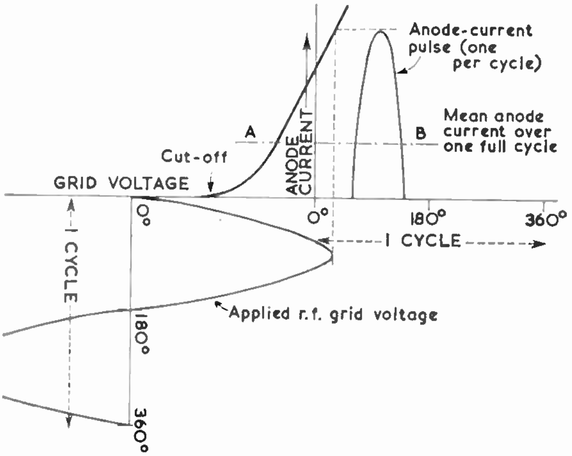


Fig. 6.51. Class C operation. The valve is biased by a voltage which is not less than twice that required to give anode current cut-off under d.c. conditions. The signal voltage applied to the grid must be of sufficient amplitude to swing the grid positive with respect to the cathode, causing grid current to flow. The resulting anode current flows in the form of a single pulse for each grid cycle.

one cycle, as would be indicated by a moving-coil milliammeter.

Although the anode current flows in the form of pulses, the r.f. voltage developed across a resonant circuit connected to the anode is approximately sinusoidal. This is due to the " flywheel " action of the resonant circuit, i.e. its tendency to allow the oscillations to continue after the removal of the current pulse.

As the angle of flow of anode current is reduced the efficiency increases. This is accomplished by increasing the negative grid bias and increasing the drive, which results in increased peak amplitude of the anode current pulses and increased peak anode current. Because the pulses are narrowed the power output will fall, but due to the increased efficiency it will fall proportionately less than the anode input power. The increase in efficiency achieved by narrowing the angle of flow to less than 120 degrees is small and is rarely worthwhile. Taking into consideration the increased drive and the higher peak current which the valve must provide, the efficiency to be expected at 120 degrees is about 75 per cent.

Class B Operation

In the class B amplifier, the anode-current pulses flow for approximately 180 degrees, i.e. half of the r.f. cycle as indicated in Fig. 6.52. The negative grid bias is sufficient to reduce the anode current in the absence of drive only to the cut-off point, and the applied grid voltage swing drives the grid slightly positive once in each cycle, as represented by the shaded areas.

Owing to the larger angle of flow the anode efficiency is less than in class C operation, but even so it may reach 60–65 per cent for c.w. operation. The r.f. voltage in the output circuit is proportional to the r.f. grid excitation voltage; i.e. the power output is proportional to the square of the grid excitation voltage. This class of amplifier may therefore be used for the amplification of amplitude-modulated signals (including single sideband signals). Since the anode efficiency varies with the grid excitation level, the mean efficiency for the amplification of modulated signals is considerably less than the value of 60–65 per cent quoted above and is dependent on the nature of the amplified signal.

Linear amplifiers operating in class B are considered in detail in Chapter 10—*Single Sideband Transmission.*

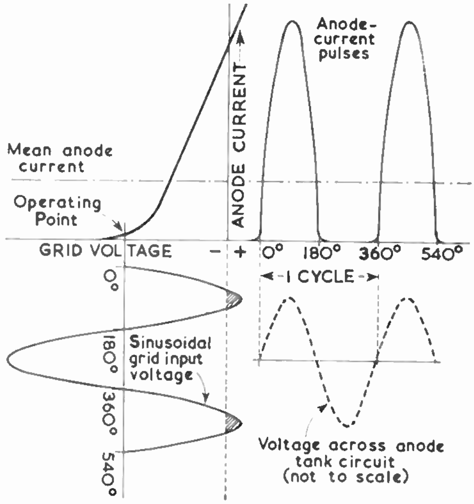


Fig. 6.52. Class B operation. The anode-current pulses are approximately of half-cycle duration and appreciable grid current flows owing to the positive incursions of the grid drive voltage.

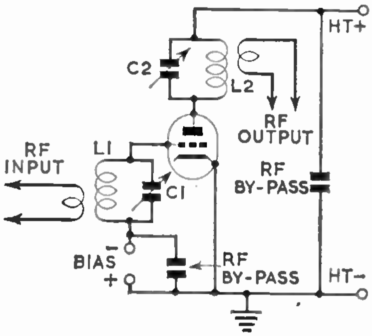


Fig. 6.53. Basic triode amplifier circuit.

Triode R.F. Amplifiers

The fundamental circuit of an r.f. amplifier using a triode is shown in Fig. 6.53. The grid is connected to a tuned circuit which is coupled to the oscillator or driver valve by any of the means previously suggested, while the output is derived from a resonant anode circuit, which is tuned to the same frequency as the grid.

It will be noticed that this basic circuit corresponds to that of a tuned anode/tuned grid oscillator. Unless something is done to prevent it, the amplifier will oscillate at a frequency determined by the tuned circuits due to the feedback from anode to grid through the inter-electrode capacitance. Only in pentode or tetrode valves designed specifically for r.f. work will the anode/grid capacitance be small enough to prevent the occurrence of self-oscillation. An 807 for example has an anode/grid capacitance of about 0·25 pF and will not generally oscillate on any amateur h.f. band unless the layout is poor. Poor layout can cause an increase in effective anode-grid capacitance due to wiring, use of poor quality valveholders, etc., or may permit sufficient inductive coupling to exist between anode and grid tank circuits to cause self-oscillation.

While it is considered advisable to neutralize pentodes and tetrodes, all triodes, without exception, must be neutralized when employed as straight amplifiers in the grounded cathode mode as Fig. 6.53. The subject of neutralization is dealt with in more detail later in this chapter.

When adjusting a triode amplifier for the first time, the following is a guide to the procedure which should be adopted:

(a) Connect the stage to a suitable dummy load.
(b) Reduce the neutralizing feedback to minimum, i.e. set the capacitor NC to minimum or reduce inductive coupling to minimum as appropriate.
(c) Without applying h.t. to this power stage, apply drive to the grid circuit and tune for maximum grid current. Reduce the drive coupling as necessary to maintain the grid current at the required value.
(d) While observing the grid current tune the anode tuning condenser over its full range, as it passes through resonance a sharp dip will be observed in the grid current. Increase the neutralization until this dip is just reduced to minimum. The stage is now neutralized.
(e) Reduced h.t. may now be applied and the tuned circuits and the drive level readjusted.
(f) If operation is satisfactory under reduced h.t. then full h.t. may be applied.
(g) Check the stability of the stage by removing the drive: key-up in a c.w. transmitter and noting that the anode current falls to the desired quiescent value and remains so for all settings of the anode tuning capacitor.

It will be found that the r.f. power output increases as the drive is increased. The power output continues to rise to a point beyond which no further increase is achieved: the amplifier is then said to be saturated. The drive power should be set slightly higher than the level at which saturation commences: at this point, the anode current can be adjusted to the value corresponding to the required power input by increasing the load coupling. Slight readjustment of the tank tuning capacitor may then be required. Finally, the excitation should again be varied to ensure saturation. It is inadvisable to allow the grid excitation to rise much higher than the value at which saturation occurs since not only is the grid dissipation thereby increased but the production of harmonics is enhanced and there is no significant increase in efficiency.

Tetrode or Pentode R.F. Amplifiers

In a tetrode or pentode amplifier the screen grid is held at a constant potential and it therefore reduces the capacitance between the control grid and the anode to an extremely low value. If suitable auxiliary external screening is provided, the coupling between the input and output circuits can be reduced to a low value. At the same time the electric field of the anode is prevented from influencing the ability of the grid to control the current flowing through the valve, and consequently much less power is required to drive a tetrode or pentode than a triode; i.e. a greater power gain is available.

Because the characteristics of a tetrode or pentode valve tend to make it operate like a constant-current device, very high voltages may occur in the anode tank circuit if it is tuned to resonance in the absence of an output load or if it is inadequately loaded. No attempt should be made therefore to explore the anode-current dip with such valves except with greatly reduced power input.

It is more satisfactory to supply the screen through a series voltage-dropping resistor from the anode supply than directly from a fixed source, since in this way the screen is protected from the danger of excessive current if the control grid should be overdriven. The series dropping resistor also protects the screen from excessive dissipation in the event of the load being removed from the anode circuit. Moreover, by feeding the screen from the anode h.t. supply the possibility of damage owing to the application of screen voltage without the normal anode voltage is diminished.

If, however, it is considered that overdriving and underloading are not likely to occur, it will be found easier to obtain full output from the valve when the screen voltage is fixed at its correct value by the use of a potential divider across the h.t. supply or a separate constant-voltage source. The high power-sensitivity of pentodes and tetrodes and the fact that neutralization is not always required are the main reasons for their popularity as r.f. amplifiers. The small drive requirements enable a complete multiband transmitter to be built within a single screened container, and the possibility of TVI resulting from the leakage of r.f. power is thus greatly reduced.

Push-pull Operation

A pair of identical valves in a push-pull circuit will produce double the power output obtainable from a single valve. Although this is also true of a pair of valves connected in parallel there are several theoretical advantages to be gained from the inherent balance which exists in the push-pull arrangement. If the valves are similar in characteristics the second harmonic and other even harmonics are balanced out in the common tank circuit. Similarly, the currents set up by one valve in the common bypass circuits are cancelled by those of opposite phase set up by the other. Since these currents can become quite large, particularly at the higher frequencies, increased stability is attainable by the use of push-pull operation. In practice, however, it is unlikely that a perfect balance will be found to exist.

A further advantage is that the L/C ratio of the tank circuits may be made higher than with single-ended stages, since the capacitances of the valves are presented in series across the coils. This feature is likely to be of greater value at frequencies of 21 Mc/s and above where the desired L/C ratio is difficult to achieve.

Parallel Operation

The output power obtainable from a push-pull or parallel arrangement is the same, provided that the correct impedances are presented to the anode and the grid circuits by use of correct L/C ratios in the tank circuits. It may prove

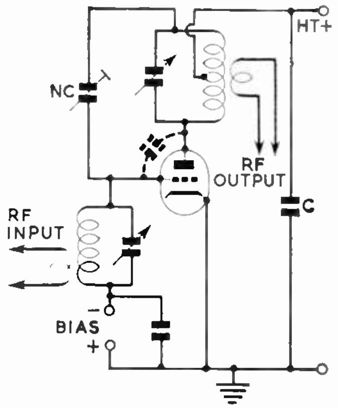


Fig. 6.54. Anode neutralization. In this arrangement the centre tap of the tank coil is earthed through the r.f. bypass capacitor C. A typical value for the neutralizing capacitor NC is 3-30 pF. When the circuit is correctly balanced, the peak voltage across NC is approximately the same as that present at the anode of the valve.

impossible to satisfy this condition when certain valves of high anode-to-earth capacitances are used in the parallel arrangement above about 21 Mc/s, but nevertheless this circuit is often considered preferable to the push-pull arrangement, owing to the simpler construction involved and in view of the fact that the theoretical advantages of the push-pull circuit mentioned earlier are not easily obtained in practice. The parallel circuit also lends itself more easily to use with a pi-section aerial coupling network.

NEUTRALIZATION

The anode-to-grid capacitance of a valve provides a path for the feedback of energy from the anode to grid; if this feedback is of sufficient amplitude oscillation will occur. This can be overcome by deliberately feeding back a voltage of equal amplitude but opposite phase. This process is known as *neutralization*.

Although it is often considered that pentodes and tetrodes do not require neutralizing for operation in the h.f. range (up to 30 Mc/s), it is usually preferable to arrange for neutralization, even when good layouts are employed. A valve may appear to be stable under constant carrier conditions but when it is amplitude modulated the rise in feedback through the anode-to-grid capacitance may be sufficient to cause either momentary or continuous self-oscillation. The former is the more difficult to diagnose because the distortion of the signal may be confused with a fault in the modulator. It is even more difficult to find when self-oscillation occurs only on the peaks of modulation.

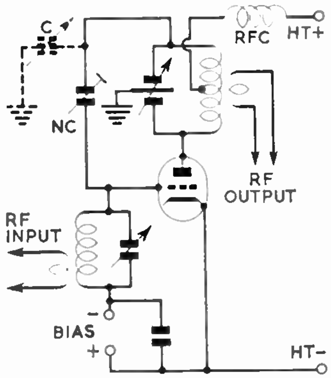


Fig. 6.55. Anode neutralization using a split-stator capacitor to provide the zero-potential centre tap in the tank circuit. The circuit also shows a capacitor C for balancing the anode-to-earth capacitance of the valves.

Anode Neutralization

In this method, a centre-tapped anode tank circuit is used in place of the simple inductor-and-capacitor so that an out-of-phase voltage is made available for neutralizing the feedback from the anode. **Fig. 6.54** shows such an arrangement in which the anode tuning inductor is centre-tapped and is fed at this point from the h.t. line thereby producing an r.f. voltage at the end of the coil opposite to the anode equal in value to the anode voltage but 180 degrees out-of-phase with it. To neutralize the internal feedback due to the anode-to-grid capacitance (shown dotted), the neutralizing capacitor NC should therefore be equal in value to this inter-electrode capacitance; in most valves this is quite small—less than 10 pF—but its effect is important and an accurate balance is required.

With this circuit a certain amount of undesirable feedback is usually present when the operating frequency is higher than about 7 Mc/s owing partly to the difficulty of tapping the coil at its exact radio-frequency centre and partly to the inevitable inductance of the leads connected to the neutralizing capacitor.

An improvement on this arrangement is shown in **Fig. 6.55.** Here a split-stator capacitor is used to provide the zero-potential centre-tap in the tank circuit. The voltages to earth of the opposite ends of the coil are determined by the relative capacitance values of the two sections of the capacitor and an accurate balancing of the two voltages is thus more easily obtained. The balancing of the circuit may be further improved by the addition of the small variable capacitor C shown dotted in Fig. 6.55. This serves to balance the anode-to-*earth* capacitance of the valve. It may be found especially beneficial whenever the neutralizing adjustment shows a tendency to vary from one band to another although with the split-stator circuit this does not normally occur.

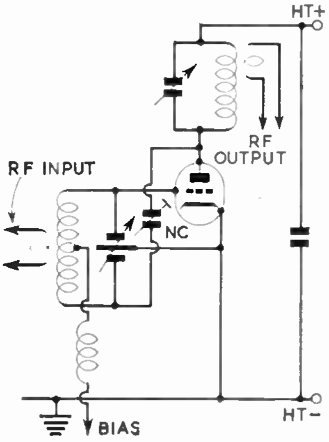


Fig. 6.56. Grid neutralization.

Grid Neutralization

A less commonly used circuit is shown in **Fig. 6.56.** This system is analogous to anode neutralization, the only important difference being that the voltage-phase reversal is achieved in the grid circuit. The method of adjustment is otherwise similar to that required for anode neutralization. Its chief advantage is that it allows a normal anode tank circuit to be retained.

Link Neutralization

The required feedback voltage may be coupled from the anode tank circuit to the grid tank circuit by means of a link

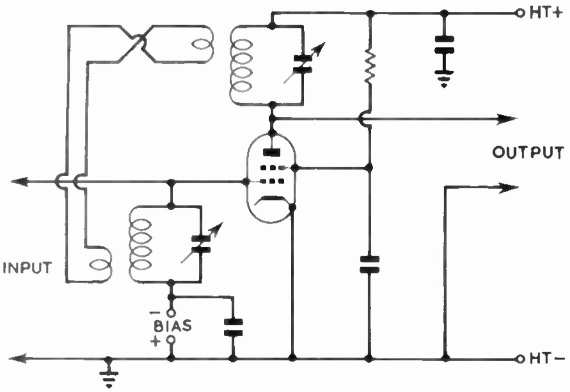


Fig. 6.57. Link neutralization. This arrangement is particularly suitable for pentode or tetrode circuits where the required amount of neutralization is relatively small.

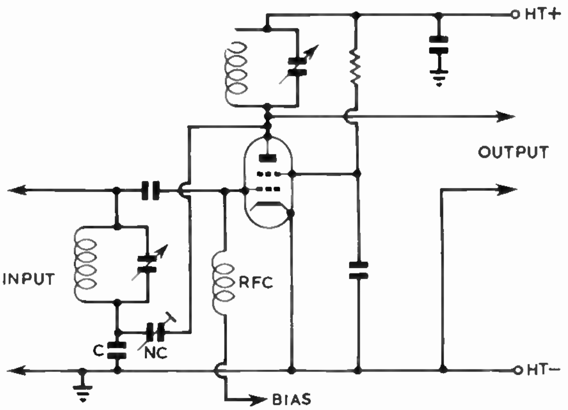


Fig. 6.58. Series-capacitance neutralization. The balancing voltage is derived from the potential divider comprising the capacitors NC and C. Typical values for an 807 are NC = 5 pF and C = 1000 pF.

line connecting the two, as shown in **Fig. 6.57**. The phase is determined by the polarization of the two coils, and if incorrect the connections to one or other of the coils should be reversed. The method is useful when a suitable neutralizing capacitor is not available or cannot easily be fitted into the layout. It is also suitable for neutralizing a tetrode or pentode amplifier which has been built without provision for neutralization and subsequently has been found slightly unstable.

As the method depends on magnetic coupling rather than capacitive feedback, it is only suitable for use on a single band unless separately positioned links are switched for each frequency range.

Series-capacitance Neutralization

Another neutralizing circuit using a capacity potential divider is shown in **Fig. 6.58**. A small capacitor NC is connected between the valve anode and a large capacitor in series with the grid tank circuit or the anode tank circuit of the preceding valve. The relative values of NC and C may be selected from the following expression:

$$\frac{NC}{C} = \frac{C_{ga}}{C_{in}}$$

where C_{ga} = anode-to-grid capacitance
C_{in} = input capacitance of the valve plus stray capacitance across the input circuit.

It is advisable not to use a value greater than 5 pF for NC because this capacitance is effectively in parallel with the output of the valve and the LC ratio of the tank circuit may be reduced undesirably. The capacitor NC in all these neutralizing arrangements is subjected to considerable voltage stress, and to avoid breakdown it must be adequately rated, especially if the stage is anode modulated.

Push-pull Neutralization

In a push-pull amplifier, neutralization can be easily applied, as shown in **Fig. 6.59** since a centre-tapped coil is already an essential feature of the circuit. Perfect balance can be obtained; if the connecting leads are kept short and equal in length the neutralization setting remains the same over a wide range of frequency including perhaps several of the amateur bands. It is common practice to use split-stator capacitors in both the grid and the anode circuits and to earth the rotors, preferably by the shortest and thickest possible connections to ensure that harmonics are effectively bypassed.

Methods of adjusting neutralizing circuits are described on page 6.43.

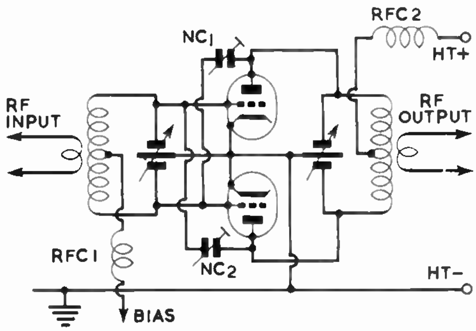


Fig. 6.59. Neutralization of a push-pull r.f. amplifier. Due to the symmetry of the arrangement, excellent balance is usually easy to obtain: one setting of the neutralizing capacitor will hold over several amateur bands provided that the connecting leads are kept short.

GRID BIASING METHODS

There are three main methods by which the required grid bias voltage may be developed for a class C stage, and these are illustrated in **Fig. 6.60**. Some methods offer a higher degree of protection to a valve than others, and for this reason, it is not unusual to find that a valve may be operated under conditions which employ more than one of the basic methods simultaneously. The methods are:

(a) *Fixed bias*, applied directly to the grid through an r.f. choke from a battery or suitable power supply with the positive terminal earthed.
(b) *Cathode bias* (sometimes known as *automatic bias*) obtained from the voltage drop across a resistor in the

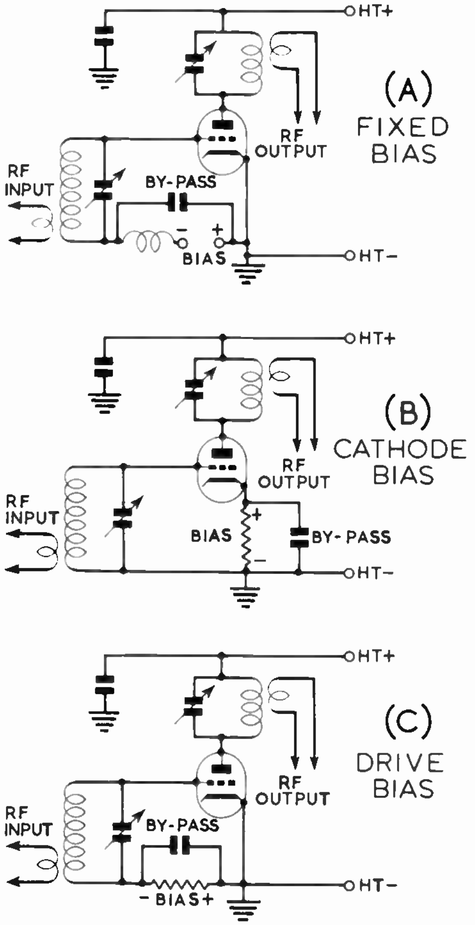


Fig. 6.60. Methods of providing grid bias in a class C amplifier.

cathode lead. Since the voltage between the anode and cathode is reduced by an amount equal to that developed across the cathode bias resistor, the supply voltage has to be increased accordingly to arrive at the optimum operating conditions. This method is therefore generally limited to valves requiring a low anode to cathode voltage.

(c) *Grid resistor* or *drive bias* in which the cathode and grid of the valve function as a diode. The applied r.f. voltage developed across the tuned circuit causes a d.c. current to flow through a resistor between grid and ground, the voltage drop across which is equal to the desired bias.

Clamp Valve

If the grid bias is derived solely from the flow of r.f. current through the grid resistor, the valve will take excessive anode current when the drive is removed. To prevent this occurring in a tetrode or pentode amplifier, a clamp valve connected as shown in Fig. 6.61 may be used. During normal operation the p.a. grid resistor or drive bias also provides bias for the clamp valve and thus renders it ineffective. On removal of the excitation, the bias on the grid of the clamp valve becomes zero and its anode/cathode resistance falls to a relatively low value. The resulting increased voltage-drop across the screen resistor *R* reduces the screen voltage of the p.a. stage sufficiently to limit the anode current to a safe value.

For effective clamp operation, a valve should be selected which is capable of passing almost the whole current available through the screen resistor. Thus, for a p.a. valve operating with a series screen dropper of 30 K ohms from an h.t. supply of 750 volts, the clamp valve should be capable of passing a current of—

$$\frac{750 \text{ volts}}{30{,}000 \text{ ohms}} = 0{\cdot}025 \text{ amperes} = 25 \text{ mA}$$

The clamp valve chosen must, however, have a grid base sufficiently short to ensure cut-off under normal drive conditions, a requirement not difficult to meet with the 6BW6, EL84 or a similar type if the bias derived across the grid resistor is greater than about 25 volts.

This method will not reduce the screen grid voltage to zero since, for current to flow through the clamp valve, there must always be a positive potential on its anode, and hence on the screen of the power amplifier. Successful operation largely depends on the slope and current capabilities of the clamp valve. In practice it is better to use a tetrode or pentode with its screen grid fed from a fixed source rather than a triode.

The current which has to be passed by the clamp valve anode should be based on the total current which could flow through the screen resistor if it were shorted to earth. In the case of a QV06-20 or 6146 operated at 600 volts, the screen resistor is 50 K ohms, and therefore the maximum current is 12 mA. To allow a generous margin, the clamper valve should be capable of passing three to four times this current, and in this case a 6BW6 or EL84 valve would be suitable. Where a pentode or tetrode is used, it is essential to derive its screen grid voltage from a potential divider. Care should be taken to ensure that the screen dissipation of the valve is not exceeded when the anode voltage falls to a value which is well below that of the screen grid. If the supply to the screen grid of the clamp valve is from a potential divider connected

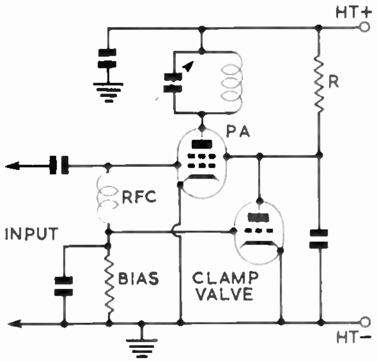


Fig. 6.61. Basic arrangement of a clamp valve to limit the anode current of the power amplifier to a safe level in the event of failure of the drive.

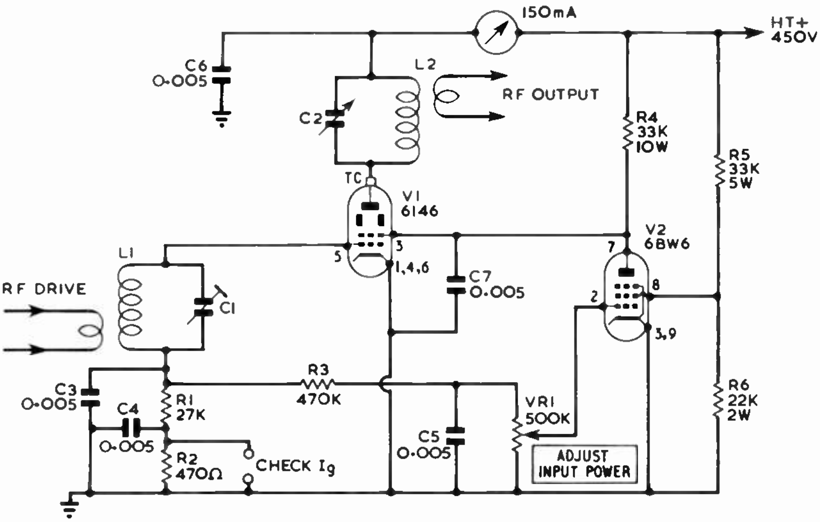


Fig. 6.62. Clamp circuit modified to provide control of the d.c. input to the power amplifier. Possible overdrive to the grid of the amplifier is discussed in the text.

across a high voltage source, the series resistor will usually ensure sufficient limiting action.

A clamp valve can be used as a convenient method of controlling the power developed in a power amplifier stage.

Since the anode current, and hence the output power, of a tetrode or pentode is directly related to the value of the screen grid voltage there will be a corresponding variation in the power output if it is varied. Where a clamper valve is employed to protect the power amplifier, such a facility can be arranged with very little change in the circuitry. A suitable circuit arrangement is shown in **Fig. 6.62.** Instead of being connected direct to the grid of the p.a. stage, the grid of the clamp valve V2 is connected to the slider of VR1 which functions as the power input control. To avoid affecting R1, the total value of R3 and VR1 is substantially higher. VR1 permits the bias on V2 to be varied from zero to about −35 volts so that, when the bias on V2 is maximum, the valve is cut off and the screen of V1 is at its normal operating voltage. As the value of the bias on V2 is reduced by means of VR1, the increased current flow through R4 lowers the voltage on V1 and hence the input.

With this arrangement it is possible for the grid of V1 to be driven too heavily under reduced anode input conditions. In such a case, the power reduction must be limited or the grid drive reduced.

Fixed Bias

The simplest method for the provision of fixed bias is a battery. For two reasons, however, its use may be found unsatisfactory:

(a) As the battery ages its resistance increases and an additional bias gradually develops owing to the grid current flowing through the battery.

(b) Grid current, which flows in a direction opposite to the normal discharge from a battery, tends to charge up the battery the voltage of which may be caused to rise well above the nominal value.

These drawbacks can be avoided by using a small mains-operated power unit incorporating the usual rectifier and ripple filter. The chief requirement in such a supply is that the output resistance must be low enough to prevent excessive auxiliary bias being developed across it by the flow of grid current. One method of ensuring this is to place a low-value bleeder resistor across the output terminals, in which case the power supply must be designed to supply a considerably larger current than is necessary for the actual biasing purposes. The difficulty is best met by the use of a voltage-stabilizing Zener diodes or valves which will maintain the output voltage constant irrespective of grid-current changes. The output from this voltage stabilizer must then be applied directly to the valve without the inclusion of an external resistor or a potential-divider network. In practice the minimum bias voltage obtainable by the use of voltage regulator valves will be 60–80 volts but Zener diodes are available for a wide range of voltages.

A most important advantage of fixed bias is that the valve will be fully protected in the event of a failure of the r.f. drive. This is quite likely to occur during tuning operations, and some form of fixed bias is therefore always advisable particularly in high-power stages.

Suitable fixed-bias circuits are described in Chapter 17 (*Power Supplies*).

Cathode Bias

Cathode bias is normally used in conjunction with fixed or drive (grid resistor) bias for a class C stage because of the high value of bias which such a stage requires. The voltage-drop across the cathode resistor must be subtracted from the h.t. supply in calculating the effective anode-cathode voltage, and it may be uneconomical to derive the whole of the required bias from the voltage-drop across the cathode resistor. The most valuable feature of cathode bias is that it automatically protects the valve by preventing the flow of excessive anode current if the auxiliary sources of bias fail. The cathode resistor, which must be bypassed for r.f., is generally chosen so that, without auxiliary grid bias, sufficient voltage is developed across it to limit the anode dissipation to within the maximum ratings. The method of calculation follows from Ohm's law ($R = E/I$):

$$\textit{Cathode resistor (ohms)} = \frac{\textit{Bias voltage (V)} \times 1000}{\textit{Cathode current (mA)}}$$

In the case of a tetrode or a pentode, the cathode current is equal to the sum of the screen and anode currents. Where it is not possible to determine the screen current at a given anode current, due to the wide difference in screen and anode currents in such valves, it is permissible to base the calculation on the required anode current limit alone and to dis-regard the screen current. This will result in a somewhat higher value resistor, providing a higher level of actual bias and an anode current below the maximum level fixed.

For example, consider a tetrode in which the anode current is to be limited to 70 mA. and which at this anode current has a screen current of 10 mA. Assume that the characteristic curves show that a bias of 10 volts is required at the applied h.t. voltage to produce these currents, then:

$$\textit{Cathode bias resistor (ohms)} = \frac{10 \times 1000}{70 + 10} = 125 \text{ ohms}$$

Disregarding the screen current:

$$\textit{Cathode bias resistor (ohms)} = \frac{10 \times 1000}{70}$$

$$= 140 \text{ ohms (approximately)}$$

The *actual* bias developed across the resistor calculated by disregarding the screen current will be:

$$E\ (\textit{volts}) = \frac{I\ (\textit{total current in mA}) \times R\ (\textit{bias resistor})}{1000}$$

$$= \frac{80 \times 140}{1000} = 11{\cdot}2 \text{ volts}$$

This will be seen to be within the 20 per cent tolerance likely to be encountered due to variations between the marked and actual value of the resistor fitted where the calculation included the screen current.

The power dissipated in the cathode resistor is given by I^2R, where I is the total current in amps; thus—

$$\textit{Power} = (0{\cdot}08)^2 \times 125 = 0{\cdot}8 \text{ watts}$$

Since this resistor is designed to protect the valve, its rated wattage should be appreciably higher than its working wattage: twice the working wattage is adequate and in the example given the resistor would be rated at 2 watts.

When a drive voltage is applied, the cathode current is augmented by the current flowing in the grid circuit but since this is relatively small it can safely be ignored in these calculations.

Drive or Grid Resistor Bias

In class C operation the r.f. grid drive is large enough to cause the grid potential to become positive at peak values, with the result that a rectified current flows in the grid circuit. If a resistor is inserted into the grid return lead the voltage drop across this resistor due to the grid current causes the grid to be biased more negatively. This provides an automatic biasing action, the amount of bias varying with the drive power and thus to some extent adjusting the actual peak drive voltage to the operating requirements of the valve.

Since failure of the drive would result in the loss of the whole of the bias if this were derived solely from a grid resistor, it is recommended that grid resistor bias should only be used in conjunction with sufficient fixed or cathode bias to limit the anode dissipation to a safe value should the drive fail. To illustrate the method of choosing the relative proportion of fixed, cathode and grid resistor bias, the following example is given.

EXAMPLE: *An 807 is to be operated under maximum power input conditions (ICAS rating) with a combination of grid-resistor bias and fixed bias. It is required to calculate the value of the grid and the fixed bias voltage.*

The relevant operating conditions obtained by reference to the published resistor data for the valve are:

Anode voltage	..	750	volts
Anode current	..	100	mA
Anode dissipation	..	30	watts
Grid bias ..	..	− 45	volts
Grid current	..	4	mA
Driving power	..	0·3	watts

First it is necessary to calculate the minimum bias required to limit the anode current to a safe value in the absence of r.f. drive. At 750 volts the maximum current permissible for a dissipation of 30 watts would be 40 mA. By referring to the characteristic curve, the bias needed is found to be approximately −20 volts. With the r.f. drive applied an additional bias of 25 volts is therefore required to raise it to the specified −45 volts. However, if the fixed bias is to be obtained from a mains-driven power unit having an internal resistance of, say, 1000 ohms, there will be an extra voltage developed due to the flow of grid current through this internal resistance: since the grid current is expected to be 4 mA, this extra voltage drop will be $1000 \times 0{\cdot}004 = 4$ volts. The fixed bias supply unit (nominally supplying 20 volts) will therefore develop an effective bias, under drive conditions, of 24 volts. This means that the series grid resistor which is to be relied upon to provide the additional bias to bring the total up to 45 volts need only contribute 21 volts. For a grid current of 4 mA, a resistor of $21/0{\cdot}004 = 5250$ ohms would thus be required.

If the fixed bias were to be obtained from a 20 volt battery (the internal resistance of which is negligibly small), the extra bias to be contributed by the series grid resistor would be 25 volts and its value would have to be increased to $25/0{\cdot}004 = 6250$ ohms.

In either case the total resistance in the grid circuit is 6250 ohms and the total resistive voltage drop amounts to 25 volts.

In practice the amplifier may be tested initially with battery bias in order to find the grid current. The value of the grid resistor can then be calculated from the following formula and the battery replaced by it:

$$\textit{Grid resistor in ohms} = \frac{\textit{Grid bias voltage} \times 1000}{\textit{Grid current (mA)}}$$

In the case of a 6146 valve with an anode voltage of 600 and a screen voltage of 150, the grid current and grid bias voltages are 2·8 mA and 58 volts negative respectively. Thus where the bias is derived solely from the grid resistor, this resistor will need to have a value of:

$$\frac{58 \times 1000}{2{\cdot}8} = 20 \text{ K ohms (approximately)}$$

Valve manufacturers specify a driving power for valves operated under class B or class C conditions but this power takes no account of circuit losses, and is simply that needed by the valve itself. Working on the assumption that the transfer efficiency between the driver and the power amplifier will be no worse than $33\frac{1}{3}$ per cent the driver should be capable of delivering three times the power needed by the driven valve. In the case of the 6146, the drive power required is 0·2 watt and the driver should be capable of delivering at least 0·6 watts.

Grid-drive Requirements

Grid Current. In normal class C operation the rectified grid current is sufficiently small to cause only a moderate rise in the temperature of the grid, but if the valve is over-driven the temperature rise may become great enough to produce grid emission and possibly cause permanent physical damage to the grid itself. Where the valve is a pentode or tetrode over-driving may also result in excessive screen dissipation. Another danger is the production of undesirable harmonics.

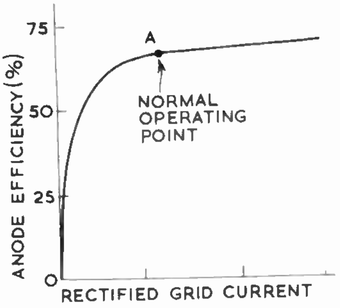


Fig. 6.63. The anode efficiency of a class C power amplifier initially rises rapidly with a small increase of grid current up to about 65 per cent, after which the increase of efficiency with drive becomes relatively slight.

A curve drawn to show the relation between grid current and the anode efficiency of a class C p.a. stage will have the form shown in **Fig. 6.63**. The efficiency initially rises quickly for a small increase of grid current from zero after which it remains substantially constant over quite a wide range. The grid current should therefore be kept as low as possible consistent with a reasonable efficiency being obtained (65–70 per cent); for example, by operating at point A.

Drive Power. The drive power required is made up of:

(a) the power required to drive the grid current against the bias voltage,
(b) the power dissipated in the grid of the valve and,
(c) the loss in the coupling arrangement.

The power in the bias source is the product of the grid current times the total bias voltage from all sources and is the same no matter how the bias voltage is obtained. The power dissipated in the grid of the valve is approximately equal to the product of grid current times the peak positive grid excursion which is often about the same as the power in the bias source. The loss in the coupling components will depend on the quality of the components used and on the layout and this is usually at least as great as the total drive power specified for the valve. The drive power available should be at least five times the power lost in the bias sources. Valves in parallel or push-pull require twice as much drive as a single valve of the same type.

In an amplitude modulated output stage the drive power and bias must be large enough to accommodate modulation peaks during which the anode voltage may rise to twice its normal maximum value in the unmodulated condition and the drive power requirement correspondingly increased.

Grid Tank Circuits

A grid tank circuit should be designed to have a Q value of about 12, for reasons which are outlined later in the discussion on *Anode Tank Circuits*. The required Q value is obtained by adjusting the LC ratio of the tank circuit, although the relative values of L and C are also dependent on the type of circuit in use. **Fig. 6.64** shows a selection of the more popular arrangements.

The optimum value of capacity for a grid tank circuit in any of these arrangements can be found directly from the abac in **Fig. 6.65**. To use this abac the data required are

(a) the operating frequency,
(b) the grid-drive power and
(c) the grid current.

In the case of push-pull circuits, the grid-drive power and the grid current will be double the value for a single-ended circuit.

Example: A single QV06-20/6146 valve is to be operated on the 3·5 Mc/s band under class C conditions. The driving power required is 0·2 watt, and grid current 2·5 mA. Find the grid circuit tank capacitance and the inductance of the associated coil.

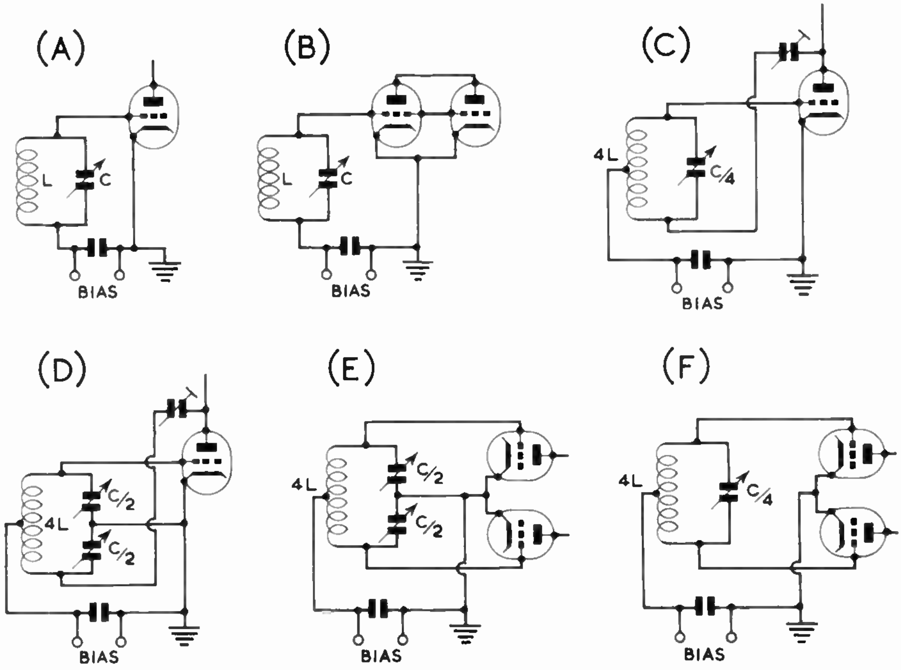


Fig. 6.64. Grid tank circuit arrangements. The values of C and L can be obtained from Fig. 6.76.

Using Fig. 6.65 join the values 2·5 mA on the grid current scale and 0·2 watt on the grid driving power scale by a straight-edge. This intersects the X_C/X_L scale at $1{\cdot}5 \times 10^3$ ohms. Join this value to the 3·5 Mc/s marker on the frequency scale. The tank circuit capacitance may now be read off the scale marked C—in this case 30 pF.

The capacitance found is that of the total across the coil, and therefore includes the stray capacitances of the circuit and the input capacitance of the valve. As these are likely to range between 10 pF and 15 pF, the net value to be provided is 30 pF − 15 pF = 15 pF. In practice a 25 pF variable capacitor would be fitted, to provide sufficient variation to compensate for all usual values of stray capacitance while still giving the total of 30 pF required.

Once the capacitance has been determined, the inductance may be calculated from:

$$L = \frac{X_L\ (ohms)}{2\pi \times 3{\cdot}5}\ \mu H$$

Which in the example corresponds to:

$$L = \frac{1{\cdot}5 \times 10^3}{2\pi \times 3{\cdot}5} = 71\ \mu H$$

When applying these calculations to push-pull circuits it should be borne in mind that the capacitance value arrived at will be the total capacitance across the complete coil, and therefore the capacity on each side of centre will be twice of the calculated value; the value of inductance calculated by use of the formula assumes that the coil is tuned by the total capacitance.

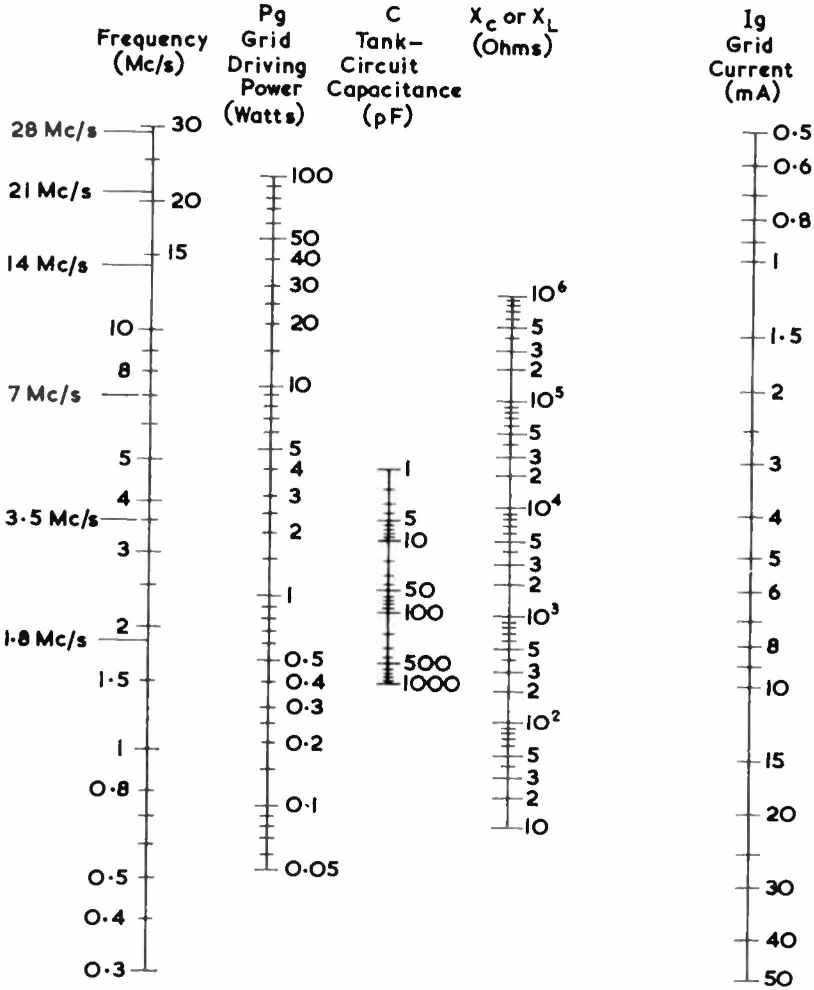
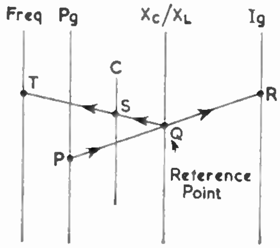


Fig. 6.65. Abac for determining grid tank circuit capacitance for a Q value of 12 in the circuit arrangements shown in Fig. 6.64. For push-pull and parallel connections, the appropriate values of grid current and power are those for the two valves taken together.

As the operating frequency is raised it will be found that the optimum value of C derived from the abac in Fig. 6.65 becomes smaller, and it may easily happen that it is less than the input capacitance of the valve. In such circumstances, the circuit will have to be allowed to operate with more than the optimum capacitance (i.e. a lower LC ratio): the only serious disadvantage will be a lowering of the efficiency of power transfer from the driver to the grid circuit of the p.a. stage. To offset this there will be the advantage of a reduction in the harmonic content of the grid-voltage waveform and a reduced likelihood of interference due to harmonic radiation.

Key to Fig. 6.65. Join the selected values of Pg and Ig by a line PQR. Note the point Q on the X_C/X_L scale. Join the point Q to the appropriate frequency T on the extreme lefthand scale. The required value of C is given at the point S. The corresponding value of L is given by the reactance value X_L at the point Q divided by 6·28 × frequency in Mc/s.



Anode Tank Circuits

The correct design of an anode tank circuit must fulfil the three following conditions:

(a) The anode circuit of the valve must be presented with the proper operating impedance in relation to the valve impedance to ensure efficient generation of power.

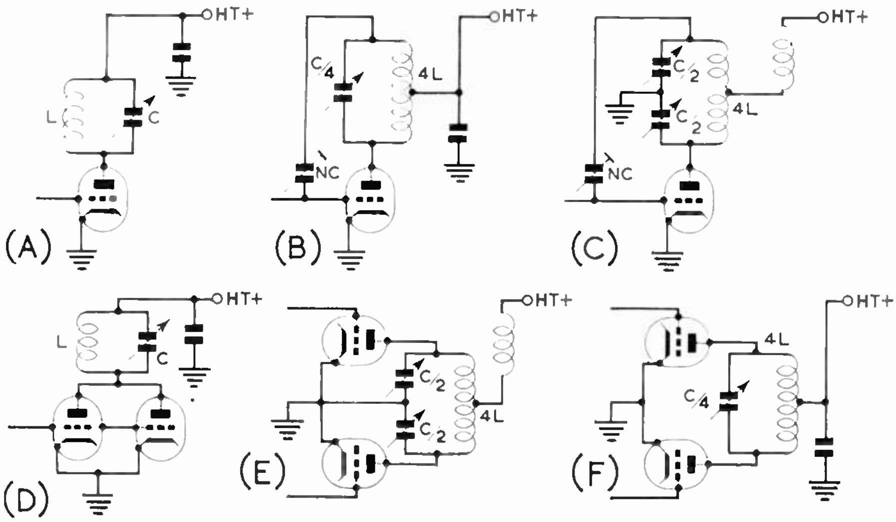


Fig. 6.66. Anode tank-circuit arrangements. The values of C and L corresponding to a loaded Q of 12 can be obtained from Fig. 6.67.

(b) The power delivered by the valve to the tuned circuit must be transferred to the load or aerial without appreciable loss.

(c) Owing to the nature of the anode-current pulses, the Q of the circuit when it is loaded by the aerial must be sufficiently high to maintain a good " flywheel " action and produce a close approximation to a sinusoidal r.f. voltage waveform.

A valve will deliver the maximum power when the output load is equal to the source impedance. For practical purposes a p.a. stage has an r.f. source impedance proportional to a figure which is obtained by dividing the anode voltage by the anode current, and the load is represented by the tank circuit shunted by the aerial load. The efficiency of the tank circuit depends on the ratio of its dynamic resistance when unloaded to its value when loaded. Because the dynamic resistance is proportional to Q, the efficiency depends on the ratio of the unloaded Q to the loaded Q.

For a given Q value the dynamic resistance can be raised or lowered by changing the reactance of the coil or capacitor, i.e. by altering the LC ratio. In a good tank circuit the unloaded Q is likely to be of the order of 100–300, while the loaded Q will be much lower. Reducing the loaded Q will reduce the circulating currents in the tank circuit, and because of the diminished heating effect it will be practicable to use smaller components. However, if the Q of the tank circuit under load is reduced too severely the " flywheel " effect becomes too small, and the harmonic content of the output increases as a consequence of the non-sinusoidal waveform. A compromise between the two requirements which has been generally accepted for use in amateur transmitters is represented by a loaded Q value of 12.

A slight increase of Q above 12 will in fact slightly lower the efficiency, but at higher Q values two other factors become significant. First, distortion can occur if the stage is modulated or if it is amplifying a modulated carrier (under class AB or B conditions) owing to the reduction in the impedance of the tank circuit at the sideband frequencies relative to the carrier frequency, but the Q would have to be greater than 200 to cause sideband clipping at 1·8 Mc/s. The effect is to give an apparent lack of modulating power at the higher audio frequencies. Secondly, the frequency band over which substantially constant power output can be obtained without re-tuning the tank circuit will become narrower, and the serviceable bandwidth then obtainable may be much less than is required for normal frequency changes within the permitted frequency band. The second effect will be much more significant than the first since the modulating frequencies are low in comparison with the carrier frequency.

The correct operating Q value can be obtained by adjusting the LC ratio or the loading but it must be remembered that the optimum proportions of L and C also depend on the particular form of anode tank circuit which is used. Several popular arrangements are shown in **Fig. 6.66.**

An abac for the direct estimation of capacitance for any given operating frequency will be found in **Fig. 6.67**: the only other data required are the operating anode voltage V_a and the anode current I_a.

EXAMPLE: *Find the anode tank capacitance and inductance required at 7 Mc/s for a pair of 5763 valves connected in parallel in a class C output stage: the anode current for a single valve is 50 mA and the h.t. supply is 300 volts.*

Using Fig. 6.67, join the values 100 mA and 300 volts on the anode-current and anode-voltage scales respectively by a straight-edge. This line intersects the X_L/X_C scale at the reference point of 150 ohms. Join this point to the 7 Mc/s-band marker on the frequency scale. The optimum tank-circuit capacitance is then read off as 160 pF. The anode-to-earth capacitance of the two valves (plus stray capacitances) is about 15 pF, leaving about 145 pF to be provided by the actual tank capacitor. In practice, a 200 pF unit would be used.

The inductance is given by

$$L = \frac{X_C}{2\pi f(Mc/s)} \text{ microhenrys}$$

$$= \frac{150}{6{\cdot}28 \times 7} = 3{\cdot}4\,\mu\text{H}$$

As in the grid circuit, the desired LC ratio cannot always be achieved in practice with some valve types at the higher frequencies and a lower LC ratio must then be accepted with some loss of efficiency.

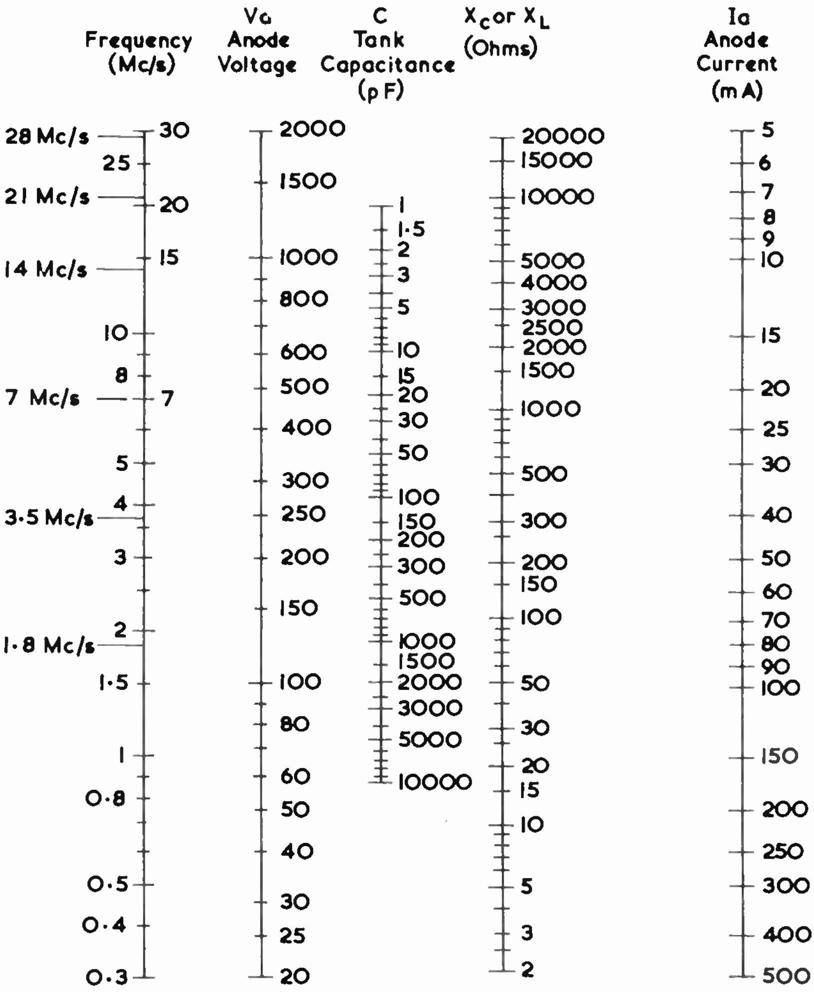
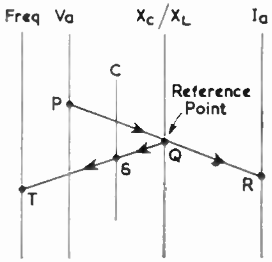


Fig. 6.67. Abac for determining anode tank-circuit capacitance for a loaded Q of 12 in the circuit arrangements shown in Fig. 6.66. For push-pull and parallel connections, the appropriate value of anode current is that for the two valves taken together.



Key to Fig. 6.67. Join the selected values of V_a and I_a by a line PQR. Note the point Q on the X_C/X_L scale. Join the point Q to the appropriate frequency and to the extreme left-hand. The required value of C is given at the point S. The corresponding value of L is given by the reactance value X at the point Q divided by 6·28 × frequency (in Mc/s).

PARASITIC OSCILLATION

Under certain conditions, the amplification of an r.f. drive voltage may be accompanied by the production of oscillations within the amplifying stage itself. The frequency of these *parasitic oscillations* is usually quite unrelated to the normal carrier frequency, and therefore interference may be caused to other services outside the amateur bands. They may be found to occur at a relatively low frequency or at a very high frequency.

Low-frequency Parasitics

Low-frequency parasitics are likely to be generated when the r.f. chokes in the grid and anode circuits happen to resonate at approximately the same frequency. The use of two similar chokes in these positions should be avoided. It is better to use a grid resistor in place of one of the chokes. The frequency of oscillation is normally below about 1–2 Mc/s and the reactance of the tank coils to such low frequencies is therefore very small except on the lower-frequency bands.

When drive (grid resistor) bias is employed, it is usually possible to dispense with the r.f. choke in the grid circuit and to connect the grid resistor directly to the tank circuit or grid, although this may mean increasing the loading on the driver to make up for the higher losses. When the grid circuit is tuned by a split-stator capacitor, direct connection of the grid resistor to the tap on the coil usually produces no measurable effect. However, if a clamp valve is used in conjunction with a push-pull input circuit, or if split-stator tuning of the grid circuit is used for neutralizing purposes, a grid choke must be employed. In such cases different types of choke may be used in the grid and anode circuits or the chokes may be damped by series or parallel resistors.

A general coverage receiver with a short aerial may be employed to detect low frequency parasitic oscillations. Alternatively, a sensitive wavemeter can be coupled to the suspected parts of the circuit. There will be no mistaking a parasitic when it is found: they are invariably highly unstable and any change in frequency is usually unrelated to v.f.o. tuning or crystal frequency.

The most difficult type of parasitic oscillation to trace is one that occurs only as the applied voltage reaches its peak during amplitude modulation. The effect on the signal is to make it sound rough. As this is similar to the effect produced by a slightly regenerative stage it is not possible to tell whether it is due to regeneration or to parasitic oscillation. The only approach is to neutralize the stage and then to look for parasitic oscillation if distortion persists.

V.h.f. Parasitics

The causes of v.h.f. parasitic oscillation are somewhat

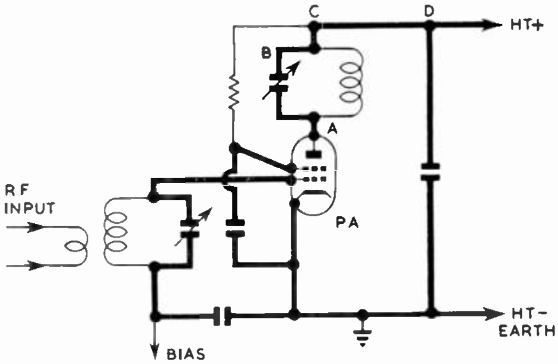


Fig. 6.68. Typical power-amplifier circuit showing how v.h.f. parasitic oscillations may be generated. The heavy lines indicate the respective v.h.f. grid and anode circuits, the inductance being constituted by the connecting leads. A neon lamp can be used to indicate the presence of such oscillation; if it is moved, for example, from A, round the path ABCD to earth, the intensity of the glow will be found to diminish progressively.

more subtle than those of the l.f. parasitics and it is not at first obvious how they can be produced. They are more common in tetrode and pentode amplifiers where the valve has a relatively high mutual conductance; often valves such as the 807, TT21 and 6146 are viewed with suspicion for this reason but the prevention of parasitic oscillation is quite simple provided that the cause is understood and the proper precautions are taken.

A tuned circuit is formed by both the grid and anode circuits as shown by the heavy lines in **Fig. 6.68** and is likely to have a resonant frequency in the range 100–200 Mc/s. At these frequencies the bypass and other capacitors and their connecting leads act as inductive reactances. The capacitance which resonates with them is that of the grid and anode circuits and other stray capacitances. If the resonant frequencies of the grid and anode circuits are not greatly different from each other, oscillation may occur. Similar conditions apply if the lead to the screen grid is resonant. By reducing the stage gain at the frequencies concerned, or alternatively by separating the v.h.f. resonance frequencies of the grid and anode circuits, oscillation can be prevented. By making connections to the tank circuits with short leads, preferably of dissimilar wire or copper strip, the inductance of the wiring may be reduced sufficiently to increase the resonant frequency to a value so high that the amplification afforded by the valve is insufficient to maintain oscillation.

A frequent cause of parasitic oscillation is the reactance at v.h.f. of bypass capacitors intended for use at h.f. For example, if the h.t. line in Fig. 6.68 were decoupled at v.h.f. at the point where the anode tank circuit joins it, in addition to being decoupled in the usual way for h.f., the v.h.f. resonant circuit would be limited to ABC instead of ABCDE.

The length of the cathode lead of a valve has a direct bearing on its ability to oscillate: the lead should therefore be made as short as possible. If the lead must be long, or has included in it a bypassed resistor for biasing, a v.h.f. bypass capacitor should be fitted with minimal leads from the cathode terminal on the valveholder to an earth point as near to it as possible. This capacitor should have a value of 470 pF and should be a type designed for bypassing purposes. Unfortunately, the addition of a larger capacitor in parallel in order to obtain sufficient decoupling at the working frequency defeats the object of the 470 pF capacitor unless the resulting parasitic circuit is damped by a low value (47 ohm) resistor close to the valveholder.

The screen grid requires special attention and it should be the rule to connect a 470 pF capacitor with minimal leads between the screen terminal of the valveholder and the cathode connection. This capacitor must be fitted between the screen grid and the cathode: to connect it from screen grid to earth would not prevent parasitic oscillation. A similar bypass capacitor should be fitted from the h.t. end of the anode tank circuit to earth with the shortest possible leads.

One of the most effective v.h.f. parasitic stoppers is a v.h.f. choke in series with the anode lead to the valve: a low value (5–10 ohm) wire wound resistor functions well as a v.h.f. choke for this purpose. This choke must be positioned directly on the anode terminal with no leads. If this is not effective, the screen grid lead should be removed and fitted with a ferrite bead, the lead reconnected and the bead positioned as close to the screen grid connection as possible. If oscillation persists, the anode choke should be removed, and a carbon resistor with a value of between 10–47 ohm fitted directly to the grid pin with minimal leads, the other end being connected to the grid circuit. The lowest value resistor needed to stop the oscillation should be used because it will introduce a power loss into the grid circuit. The positions for parasitic stoppers, the nature of the stopper, and the order in which they should be tried is shown in **Fig. 6.69**.

Small v.h.f. chokes can be constructed by taking approximately a quarter of a wavelength of wire at the parastic frequency, and winding it on a low loss former so that it has a length which is at least three times as great as its diameter. If it is possible to make the choke self-supporting it will be measurably more efficient. For parasitic oscillations in the range 100–200 Mc/s the wire length to be wound should be 18 in., and for the 200–300 Mc/s range 9 in.

Identification of Parasitics

The presence of either type of parasitic oscillation will render the tuning characteristics of the amplifier erratic and will reduce the efficiency at the operating frequency. In some

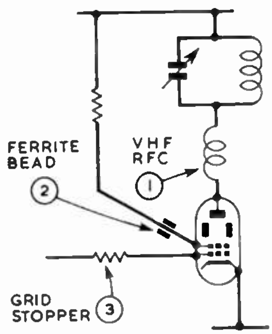


Fig. 6.69. Positions for v.h.f. parasitic stoppers, the nature of the stopper, and the order in which they should be tried.

cases, the valve is overloaded owing to the parasitic oscillation and may be damaged as a result. Apart from the possibility of interference by unwanted radiation, these parasitics should be eliminated therefore to allow the full efficiency to be obtained. Before checking for parastic oscillation it is desirable to carry out "cold" neutralization of the stage first by observing the grid current as the anode circuit is tuned through resonance with the h.t. switched off.

The tendency of the p.a. stage to oscillate can be investigated in the following way. The r.f. drive to the amplifier should be removed and the grid bias adjusted so that the anode current does not exceed the safe maximum value. If self-oscillation at a frequency corresponding to the actual tuning circuits is taking place, owing to lack of screening or neutralization, an adjustment of either the grid or anode tuning capacitors will alter the frequency (which will be heard on a receiver possibly as a rough unstable note): the anode current will also vary with the tuning adjustment. Such self-oscillation must first be eliminated by improving the screening arrangements and by ensuring that the amplifier is properly neutralized.

If the parasitic oscillation is of the low frequency type, the tuning capacitors will have very little effect on its frequency; a neon lamp held near the anode circuit will glow with a darkish red colour at any part of it, including the apparently "earthy" end. In the case of v.h.f. parasitics, the neon lamp will glow with a reddish-purple colour when held near the anode, but the intensity of the glow will diminish as the neon is moved round the "parasitic" anode circuit (ABCDE in Fig. 6.68). Sometimes parastics are more readily generated when the tank capacitors are at their maximum values, and the capacitors should therefore be varied throughout their full ranges to ensure all possible conditions are examined. A sharp upward "flick" of the anode current while this is being done indicates that the stage requires neutralizing.

In clearing parasitic troubles, it often helps to recognise that the spurious oscillation adopts a different mode of operation from that at the operating frequency. For example, in a push-pull stage a low frequency oscillation caused by the r.f. chokes could well occur due to the valves operating in parallel in the spurious mode. Similarly, when the valves are connected in parallel, a v.h.f. parasite could occur in a push-pull mode. Suppressing devices which affect the parasite but do not affect normal operation can sometimes be devised.

TRANSISTOR POWER AMPLIFIERS

The conditions which exist in a class C transistor power amplifier are similar to those of its valve counterpart but, unlike the valve, both the input and output impedances of a transistor are very low. The exception is the common base amplifier illustrated in Fig. 6.47(B) where the output impedance is high. The problems in a transistor power amplifier, and also in a transistor frequency multiplier, are mainly those of securing an accurate match between the output impedance and the tuned circuit. There is also the possibility of having to provide neutralization.

As the output impedance is so low, matching has to be to a fairly high degree of precision if maximum power output is to be secured. While it is possible to design a tank circuit with an impedance which matches that of the transistor directly and has a working Q of 12, the amount of capacity needed in such a circuit is so high that it is not usually a practical proposition. Since the impedance of the output link and its associated feed line is considerably higher than that of the tank circuit itself, there has to be a step-up ratio between the output circuit and the feed line rather than the step-down ratio usual in valve circuits.

Input Circuit

The input impedance of a transistor is relatively low whether the drive is fed to the base or to the emitter. As a guide, the input impedance of the base is usually in the region of the order of 600 ohms, while that of the emitter is normally of the order of 100 ohms. In view of these low impedances, matching to the input is essential if the best possible transfer of power is to be secured. Where drive bias is employed it becomes even more important. Tapped coil or link matching may be employed.

In transistor transmitters it is not common practice to provide a separate tuned circuit for the input as is the case with valve transmitters. Instead, the base or the emitter is driven by a link winding in association with the collector coil of the previous stage. The turns of the link are adjusted to give the necessary impedance transformation. This method of coupling has the advantage that if provision is made to vary the position of the link in relation to the coil, the drive can be adjusted easily.

Biasing

In class B operation no standing bias is provided. Unless a transistor has a forward bias current applied between its emitter and base, no current will flow through it until such a bias is provided. Thus a transistor can be used in class B simply by connecting it so that there is no forward bias applied.

Fig. 6.70 is the circuit of a typical transistor power amplifier in which the drive is applied to the base, and the base matched to the preceding stage by a link winding. While one end of the link winding is bypassed for r.f., the d.c. return is via a resistor R to earth; the emitter is also earthed. Under static conditions, i.e. without drive applied, the transistor rests in the class B condition, and the only collector current which flows is that due to leakage through the transistor. When drive is applied progressively, the transistor conducts on alternate half cycles. As the level of the drive is increased, diode rectifying action takes place between the base and emitter, and a biasing voltage is developed, the level of which will be in proportion to the drive power and

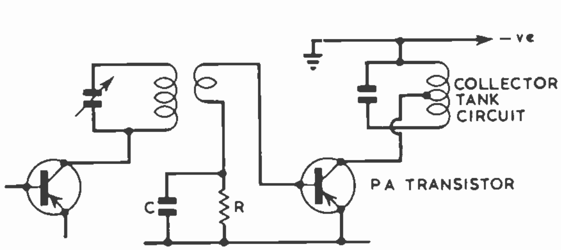


Fig. 6.70. Representative common emitter connected transistor p.a. stage.

the value of R. Thus by varying the drive power, the transistor can be moved progressively from class B to class C operation.

As it is not usual to find that manufacturers specify precise conditions for transmitter service, except for semiconductor devices specifically produced for this purpose, a certain amount of work may be needed in designing a transmitter employing general purpose transistors. So far as the bias level is concerned, it is important to ensure that the base emitter voltage specification is not exceeded, particularly the reverse voltage rating, and this should be continuously monitored as the drive is increased. Overdriving will quickly ruin a transistor.

Collector Tank Circuit

While an operating Q of 12 for a transistor tank circuit has been mentioned, this can only be achieved under special conditions. The more usual value is a Q of five, the higher harmonic content of which is usually kept under control by employing an aerial tuning unit.

The correct output load impedance for a transistor may be determined from:

$$R_L = \frac{V_C^2}{2P_O}$$

where V_C is the voltage at the collector of the transistor and P_O the power output.

For example, assume a transistor has a collector voltage of 12 volts and a power output of 5 watts (AUY10), then its optimum load impedance will be:

$$R_L = \frac{12^2}{2 \times 5} = \frac{144}{10} = 14{\cdot}4 \text{ ohms.}$$

For the full 5 watts output to be developed, the transistor must therefore be matched into a load of 14·4 ohms.

To design a tank circuit which would work with such an impedance directly with a Q of five, the reactance of the capacitor X_C is given by:

$$X_C = \frac{R_L}{Q}$$

from which

$$X_C = \frac{14{\cdot}4}{5} = 2{\cdot}8 \text{ ohms.}$$

At any frequency in the h.f. range this results in a capacitor of very large size, so large that practically no coil would exist in order to achieve resonance $X_L = X_C$. From a practical point of view such a procedure is clearly impossible.

The solution to this is to tap the collector on to a well-designed tank circuit. If the collector is, for example, connected to the centre tap, then the 14·4 ohm load resistance will be transformed by the turns ratio squared and so in this case will become 57·6 ohms.

Substituting this in

$$X_L = X_C = \frac{R_L}{Q}$$

produces

$$\frac{57{\cdot}6}{5} = 11{\cdot}5 \text{ ohms.}$$

Although this is better, it is not good and corresponds to about 3500 pF in the 3·5 Mc/s range. If the collector is tapped in at one fifth the total number of turns on the coil, then the 14·4 ohm load is transformed up to 360 ohms, and for a Q of five on 3·5 Mc/s, X_C becomes 75 ohms which gives a value of about 600 pF which is not unreasonable from a practical point of view.

Similar calculations may be made for other transistor powers and voltages and the tapping point on the tank circuit coil determined. From this the value of X_C/X_L in relation to the Q may be evaluated. The value of X_C in terms of capacity may be obtained from the abac of Fig. 6.67 by placing a straightedge from the correct position on the X_C/X_L scale across to the desired frequency on the frequency scale, and reading off the value of C where it intersects.

Collector Voltages

All the voltages applied to a transistor need the most careful attention for, unlike valves, transistors are not sympathetic to even momentary voltage overloads. In the tank circuit of a transistor class C amplifier, and to a lesser extent a class B amplifier, the peak voltages will far exceed the supply voltage, and this must be taken into account when transistors are selected for this class of service. Especially is this so when the transistor is amplitude modulated. In such cases, it is desirable that the rated working voltage between collector and emitter should be five times the applied d.c. voltage. When tuning up any transistor transmitter, the stages should be loaded to some degree to reduce the peak voltages between the collectors and emitters.

Transistor Neutralization

Like valves, transistors have internal capacities, and for the same reasons, these may have to be neutralized depending on their magnitude and the frequency of operation.

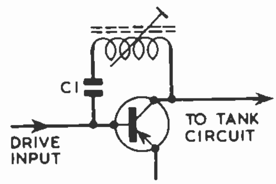


Fig. 6.71. Neutralizing circuit for a transistor p.a.

Transistors also have internal conductance which sometimes requires neutralization. If the neutralizing of a transistor is incorrect, there may be a loss of gain (degeneration) at one frequency setting of the output circuit while at another it may be highly regenerative.

The neutralizing system which appears to be the easiest to adjust is illustrated in **Fig. 6.71**. It was originally designed for use with triode valve amplifiers and was later used in cascode r.f. amplifiers.

In the circuit of Fig. 6·71, the inductance is arranged to resonate with the collector-to-base capacitance of the transistor at the operating frequency. At resonance, since $X_C = X_L$, the internal capacity of the transistor will be neutralized.

Output Matching

Output matching is arranged in the same manner as for valves, and is best accomplished by fitting a link to the collector tuned circuit, the turns of which are proportioned in relation to those on the tank circuit inductance.

TABLE 6.7

Self-resonant frequencies of capacitors commonly used in amateur transmitters.

Capacitor Type	Maker	Lead Length	Frequency
0·1 μF 350 volts paper foil (4702A)	Dubilier	½ in.	3·4 Mc/s
0·1 μF metalized paper (W99)	Hunts	¼ in.	5·95 Mc/s
0·01 μF ceramic disc (PZ)	Erie	¼ in.	17·8 Mc/s
2200 pF polystyrene	GEC	¼ in.	30 Mc/s
0·001 μF metalized paper (W99)	Hunts	¼ in.	43·6 Mc/s
0·001 μF ceramic disc (NY)	Erie	¼ in.	53 Mc/s

Bypass Capacitors

Due to the lower impedances encountered in transistor circuits compared with those in valve circuits, any residual impedance in bypass capacitors must be correspondingly lower. It will be found therefore that the bypass capacitors employed in transistor transmitter circuits are larger relative to frequency than those in similar valve circuits. It is essential that the type of capacitor is selected with care, since increasing the value of the capacitance may be accompanied *by an increase* in reactance in some cases. This is of particular importance so far as the capacitor across *R* in Fig. 6.70 is concerned, for if this shows reactance at the operating frequency it will either upset the matching of the link to the previous stage, or necessitate an increase in the drive to the p.a. transistor, or both.

OUTPUT COUPLING CIRCUITS

The output coupling circuit of the tank circuit of a transmitter is as important as the output circuit itself. There is no point in having potential energy available if it cannot be used.

Link Coupling

The most commonly used method of transferring the power available in the anode circuit is by means of a link winding. This consists of a few turns of wire, usually of the same gauge as the winding on the coil, placed adjacent to the earthy end of the coil and co-axial with it. In practice, and to allow the position of the link winding to be moved relative to the tank coil, it may be wound fractionally larger so allowing it to slide up and down the coil former.

Before winding the link, it is necessary to know the impedance into which the link will feed. The turns on the link and the turns on the tank coil constitute an impedance transformer. Just as it is necessary to match the anode tank circuit to the p.a. impedance for optimum power, so it is necessary to match the link to the impedance it has to feed if the most efficient transfer is to take place between the tank circuit and the link.

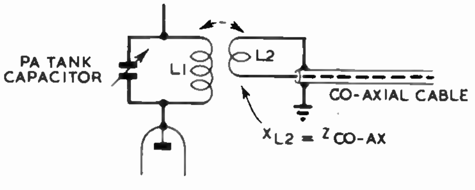


Fig. 6.72. Untuned link output circuit.

The load resistance which is placed across the link coil is transformed into the primary circuit to produce the load resistance for the valve or transistor.

The usual practice is to feed the power from the p.a. by means of a link connected to a co-axial cable which feeds an aerial array with an impedance which matches that of the co-axial line or an aerial matching device. The arrangement is illustrated in **Fig. 6.72.**

The ideal conditions are not always easy to meet in practice. If the link is accurately matched to the line, it is frequently found that as it is coupled to the p.a. tank coil, the leakage inductance of the link detunes the p.a. circuit. In order to restore the p.a. current to its loaded value, the tuning capacitor needs to be adjusted to compensate for this leakage inductance. As the coupling between the link and the p.a. coil is increased (as they are moved physically closer) the detuning becomes more pronounced. However, the effect is usually more irritating than serious.

Tuned Link Coupling

As an alternative to the simple link, a tuned link system can be employed, and in many instances this can show distinct advantages. The arrangement is shown in **Fig. 6.73.**

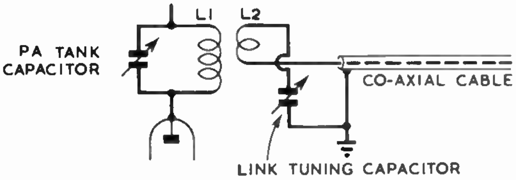


Fig. 6.73. Tuned link output circuit. The link L2 together with the capacitor forms a series tuned circuit resonant at the operating frequency. The link capacitor should tune through the value given in Table 6.8.

Provided that the far end of the coaxial cable is correctly terminated, the input impedance will be substantially resistive and equal in value to the impedance of the cable. By the use of a series tuned circuit this impedance can be matched. While the Q will be low, this is no particular disadvantage for, if the Q were high, the tuning of the link would require adjusting with excursions in frequency.

Table 6.8 gives the value of capacitance which, in association with a suitable coil, will produce a link circuit with a Q of two. The link should be wound so that when it is lightly coupled to the p.a. tank circuit, the anode current of the p.a. rises as the link tuning capacitor is tuned through the value

TABLE 6.8

Values of the link tuning capacitor in Fig. 6.73 with which the link L2 must be made to resonate at the operating frequency. The values given should not be exceeded at resonance, but a value of capacitor should be fitted which is somewhat higher than the maximum figure given.

Amateur Band	Impedance of Co-axial Cable 52 ohm	75 ohm
1·8 Mc/s	900 pF	600 pF
3·5 Mc/s	450 pF	300 pF
7 Mc/s	230 pF	150 pF
14 Mc/s	115 pF	75 pF
21 Mc/s	80 pF	50 pF
28 Mc/s	60 pF	40 pF

specified in the table. Once the coil has been adjusted to produce this condition, the link capacitor should be left at the value which produces the peak in the anode current of the p.a. The link coil should be moved towards the tank coil until the desired degree of loading is obtained. It is probable that the co-axial line, even if correctly terminated, will exhibit some reactive element. As the frequency is changed, the link tuning capacitor may therefore require slight adjustment. The degree of trimming that this capacitor will require will depend on the extent of the reactive component.

Due to the low Q of the tuned link, fairly tight coupling between the link and the tuned circuit will be required. However, as it is resonant, and any reactance can be tuned out, the tank capacitor will show less deviation from the unloaded dip position as the coupling is increased compared to the untuned link method. This is to be expected in a double tuned circuit using mutual inductance. Any change experienced is usually due to increases in stray capacitive coupling as they approach each other.

Pi-Network Coupler

The pi-network is a circuit which is capable of matching the p.a. impedance of a transmitter to a wide range of load impedances. It uses comparatively few components.

The commonly-used pi-network arrangement is shown in **Fig. 6.74.** R1 (shown dotted) represents the output resistance of the valve which is to be matched through the network to the aerial or load impedance (equal to R2 + X2). In order to achieve a correct match, the reactive part of the aerial impedance must be eliminated when presented through the network to the valve. At the same time, the loaded Q of the network when regarded as a tank circuit should normally be in the region of 12 in order to provide the necessary "flywheel" effect.

To meet these requirements, the two capacitors C1 and C2 and the inductance L should have certain values of reactance as given by the following formulae:

$$X_{C1} = \frac{R1}{Q}\left(1 + \sqrt{\frac{R2}{R1}}\right) \qquad \ldots(1)$$

$$X_{C2} = X_{C1}\sqrt{\frac{R2}{R1}} \qquad \ldots(2)$$

$$X_L = \frac{R1}{Q}\left(1 + \sqrt{\frac{R2}{R1}}\right)^2 \qquad \ldots(3)$$

where Q is the loaded Q of the circuit.

For any particular frequency, the corresponding values of C1, C2 and L can be calculated from the reactance charts shown on page 6.42 or from basic formulae.

The maximum practicable ratio between R1 and the resistive component of the load R2 is about 100 : 1. Correct matching will be more easily obtained if the ratio is appreciably smaller than this.

The value of R1 representing the effective p.a. anode impedance can be calculated by assuming that the peak r.f. voltage swing at the anode is about 80 per cent of the d.c. supply voltage, V_{ht}. It will then be given by—

$$R1 = \frac{(0{\cdot}57 \times V_{ht})^2}{P} \text{ ohms} \qquad \ldots(4)$$

where P is the power *output* of the valve in watts. It is much simpler, however, to measure the power *input*, and if an assumption is made with regard to the anode efficiency the expression can be converted into a more useful form. Thus, if the stage is assumed to be 70 per cent efficient, the expression becomes—

$$R1 = 460 \times \frac{V_{ht}}{I_a} \text{ ohms} \qquad \ldots(5)$$

where I_a is the anode current in milliamperes. Either of the expressions (4) or (5) may be calculated from the operating data given in the valve manufacturer's catalogue.

A chart derived from equations (1) and (3) is given in **Fig. 6.75** from which X_{C1} and X_L may be determined for a required ratio R2/R1, assuming that the Q of the loaded circuit is 12. The value of X_{C2} is then obtained from equation (2). The following example shows how this chart is used.

EXAMPLE: *A single ended QV06-20/6146 is to be operated on 3·5 Mc/s at an h.t. of 500 volts and an anode current of 100 mA. Find the constants of a pi-network with a Q of 12 to match a load impedance of 75 ohms.*

From equation (4)

$$R1 = 460 \times \frac{500}{100} = 2300 \text{ ohms}$$

The ratio of R2 : R1 is 75 : 2300 or approximately 0·03 which is well within the limit for securing a Q of 12 (i.e. R1/R2 $\ngtr$ 100).

Using Fig. 6.75 to determine X_{C1}, left-hand chart, proceed up the 0·03 vertical, and estimate the position of 2300 ohms between the 2000 and 3000 ohm curves. Read off X_{C1} on the lefthand scale: this is approximately 240 ohms.

Using the right-hand chart in the same manner, determine X_L. This is approximately 260 ohms.

Applying equation (2) calculate X_{C2}

$$X_{C2} = 240\sqrt{\frac{75}{2300}} = 45 \text{ ohms approximately}$$

The reactances of X_{C1}, 240 ohms X_{C2}, 45 ohms, and X_L, 260 ohms are now known.

From the reactance charts (Chapter 22) convert these into practical units: C1, 180 pF; C2, 1000 pF, and L1, 10 μH.

To allow for errors in reading the charts, it is wise to fit larger values than those calculated. In this example, the components fitted would be C1, 250 pF, and C2, 1000 pF variable with 500 pF fixed in parallel giving a total of

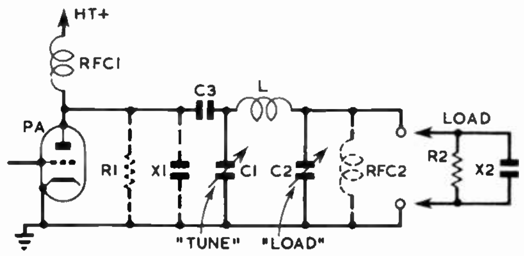


Fig. 6.74. Basic pi-network impedance matching circuit for feeding an unbalanced load. R1 and X1 represent the output resistance and reactance of the valve: R2 and X2 are the resistance and reactance of the load. The proper matching of R1 and R2 is effected by varying the relative values of C1 and C2; the reactances X1 and X2 can be cancelled by the positive or negative variation of these capacitors. In practice, a choke RFC2 is often added to prevent the appearance of the h.t. supply voltage across the load terminals in the event of an insulation-breakdown in C3.

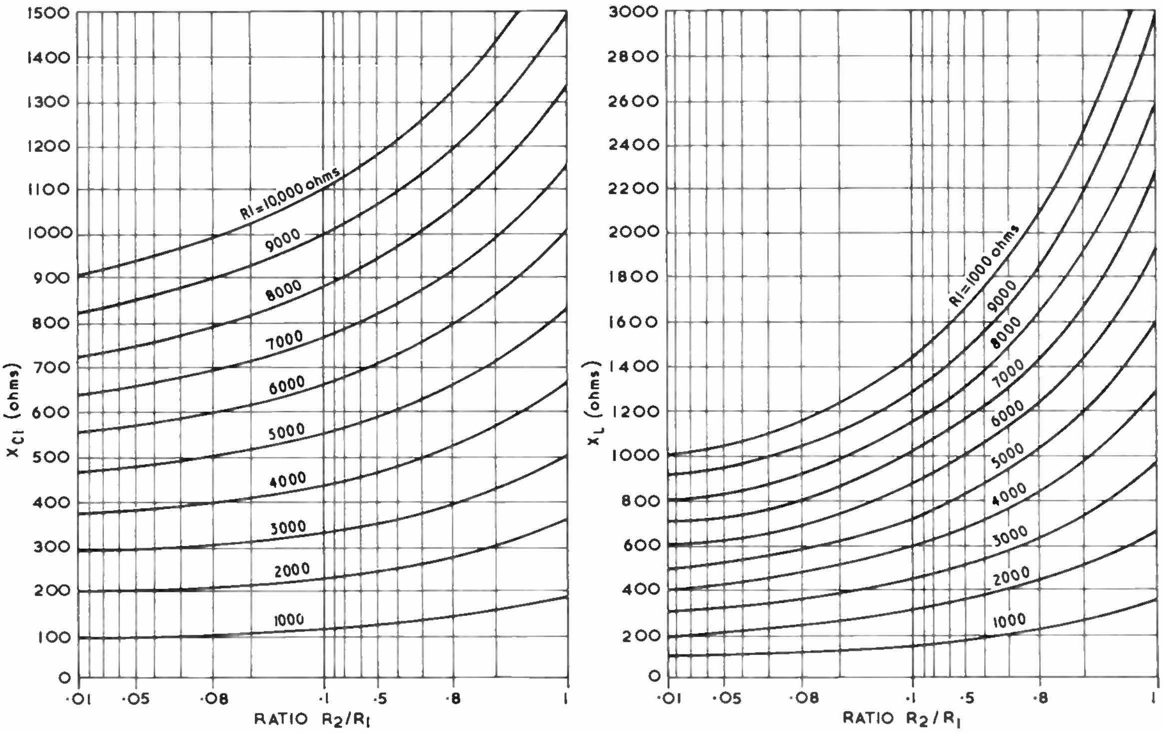


Fig. 6.75. Reactance charts for the design of pi-network couplers: see text for details.

1500 pF. Additional charts for the design of pi-network tank circuits are given in the *Radio Data Reference Book* (RSGB).

The pi-network coupler has one important advantage over the parallel tank circuit. As it functions as a low pass filter, it shows a high degree of attenuation to harmonics of the frequency to which it is tuned, but to secure the maximum degree of suppression, the components have to be carefully placed. The capacitors C1 and C2 should be mounted directly on the chassis with C1 as close to the p.a. valve as possible. The outer braiding of the co-axial cable should be connected to the same point as the earthing of the rotor of *C*2, and the inner conductor should be shielded right up to the point where it is connected to C2.

The earth connection through the chassis between *C*1, L and *C*2 should be of good conductivity. Since the chassis will carry the circulating current care must be taken that stray coupling to earlier stages does not exist. An excellent arrangement would be to connect the capacitors together with copper strip and connect them to earth at one point only.

In operation the pi-network offers the particular attraction that any reactance on the co-axial feed line can usually be tuned out.

The amount of capacitance used at positions C1 and C2 will be modified by the stray circuit capacitances, but because these are normally small when compared to the values of C1 and C2, and since it is usual to provide a somewhat higher variable capacitor than that of the calculated value, stray capacities can usually be ignored within the frequency range 1·8–30 Mc/s.

As the values of C1 and C2 decrease as the frequency increases, multiband operation can sometimes be arranged by designing the coupler for the lowest frequency band, and providing the coil with a number of suitable tapping points.

This will not give as high efficiency as that attainable with individual coils for each band, for as the tapping points are bridged the working part of the coil will " see " a number of shorted turns, and these will reduce the efficiency. Alternatively a variable inductance can be used and *C*1 switched for each band.

PA Choke Considerations

Owing to the tendency of an r.f. choke to resonate at various frequencies, depending on its design, some caution is necessary when using a parallel-fed tank circuit. A choke

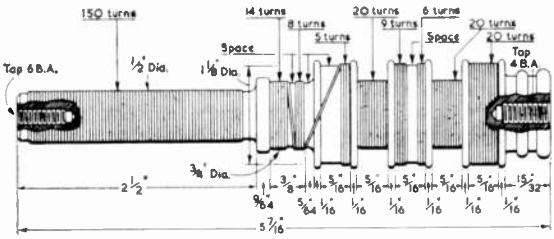


Fig. 6.76. Construction of an all-band r.f. choke suitable for a parallel-fed p.a. stage. The former may be made of Tufnol or p.t.f.e. The wire used is 30 s.w.g. d.s.c.

suitable for an all-band high-power parallel-fed p.a. stage is illustrated in **Fig. 6.76**. This has been designed so that no resonances are present within the amateur bands; nevertheless the possibility of an unexpected resonance should be borne in mind since a metal chassis or a screening partition in the vicinity of the windings can alter the characteristics appreciably. Resonance will be noted by a loss in efficiency and by overheating of the choke windings. The choke former should be made from a material with good electrical properties and capable of withstanding without softening the temperature which it may acquire by being mounted close to the p.a. valve. A suitable material is Tufnol, or p.t.f.e.

Pi-networks in Push-pull

It is practical to apply the advantages of a pi-network, to valves operating in push-pull and this is shown in **Fig. 6.77**.

In this the calculations are made for one half of the circuit according to the values given on the diagram, and similar components fitted to each half.

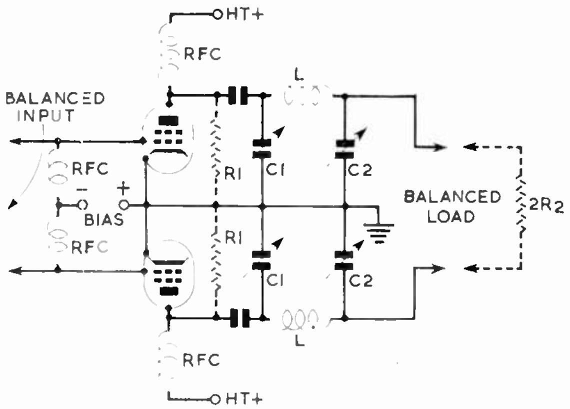


Fig. 6.77. Twin pi-network for a push-pull output stage. If preferred the pairs of capacitors C1, C1, and C2, C2 can be replaced by single capacitors, the earthed centre connections then being ignored.

Pi-output Safety Precautions

If the blocking capacitor shown as C3 in Fig. 6.74 develops a short circuit, there would be a possibility of the load assuming a dangerously high potential if it had no d.c. return. To prevent such a situation, the r.f. choke RFC2 should always be fitted, and the h.t. supply to the p.a. valve correctly fused. With certain aerial systems, this choke will provide a path for electrostatic voltages built up on the aerial. To avoid the transmitter enclosure assuming a dangerous potential under electrical storm conditions, it should always be connected to an efficient earth.

Pi-network Neutralization

P.a. stages employing pi-networks may be neutralized by means of the arrangements illustrated in Figs. 6.56 and 6.58. Cross-over neutralization may be used with push-pull pi-network stages. Contrary to popular opinion on the subject it is always best to neutralize p.a. stages using modern valves.

Adjusting the Pi-network Coupler

To tune a pi-network coupling the following procedure will be found satisfactory. First, with a dummy load connected, set the " loading " capacitor (C2 in Fig. 6.74) to maximum capacity and tune the anode capacitor (C1 in Fig. 6.74) for minimum p.a. anode current, with a low power input to avoid possible damage to the valve. Next connect the aerial load or its equivalent and decrease the capacitance of C2 slightly from its maximum value; this capacitor C2 can conveniently be regarded as a " loading " control. Then C1 should be readjusted for minimum anode current. The procedure is then repeated until a value of C2 is found at which the output into the load is a maximum and at the same time a dip in anode current of 10–15 per cent can still be obtained by varying C1, the dip being coincident with maximum output.

An alternative, and possibly more satisfactory, method is to monitor the output voltage or current and adjust the pi-network for the required output with normal input.

If the p.a. loading is found to increase as the value of C2 is increased, the ratio between the impedances R1 and R2 is too great to allow correct matching to take place. In this event, it is preferable to modify the load impedance R2, usually by connecting a capacitance in series; its optimum value will best be determined by experiment. This capacitor should be suitably designed to carry the relatively heavy output-load current and one having a good mica dielectric should prove satisfactory.

ADJUSTING POWER AMPLIFIERS

To avoid possible damage to the valve in a p.a. stage it is desirable to make the preliminary tuning adjustments with the anode and screen voltages removed. If a grid dip oscillator is available, it is a simple matter to tune the anode and grid circuits to the frequency of operation before any power is applied.

If a grid dip oscillator is not employed, normal bias should be applied to the valve. If link coupling is used, both links should be adjusted to give rather less than maximum coupling. A milliammeter in the grid circuit of the p.a. stage, **Fig. 6.78**, will indicate when optimum coupling has been obtained. The procedure is to tune the driver anode tank circuit to give minimum anode current in the driver, and, with the links set as above, the p.a. grid circuit is tuned to give maximum drive as indicated by the grid milliammeter. The link coils may then be readjusted to give tighter coupling, followed by further slight retuning of the driver anode and p.a. tank circuits, if necessary, to give maximum grid current. The latter should have a value 30-40 per cent greater than the specified value for the valve since it will fall by approximately this amount when the anode voltage is applied later. The frequency to which the grid circuit is tuned should be checked with an absorption wavemeter.

Neutralization

The stage must now be neutralized. Variation of the anode tuning capacitor through resonance should have no influence on the grid current, provided that the stage is correctly neutralized. Neutralization should be carried out at the highest frequency on which the amplifier is to operate.

There are two main methods of neutralizing an amplifier: with the h.t. removed from the anode and screen of the valve (" cold " neutralization) and with the h.t. applied (" hot " neutralization). Cold neutralization is to be preferred on the grounds of safety.

The procedure for neutralizing an amplifier is as follows:

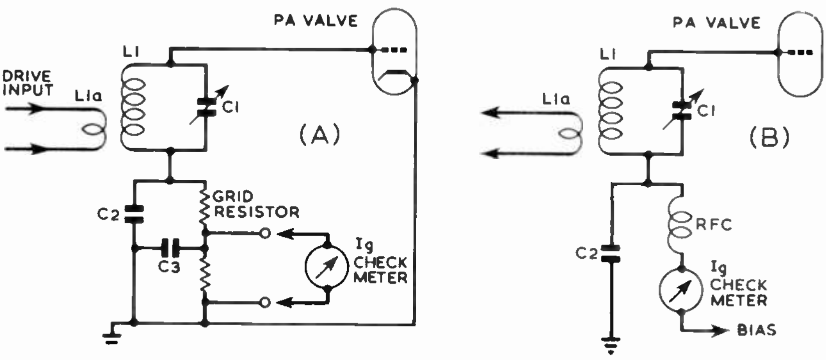


Fig. 6.78. Grid current meter positions. The use of a grid current meter is essential during the tuning-up of a p.a. and should preferably be a permanent feature in any transmitter.

Cold Method

(a) Disconnect the h.t. from the anode and screen of the valve.
(b) Disconnect any output load from the valve.
(c) Set the neutralizing capacitor to minimum capacity.
(d) Set the anode tuning capacitor to the point at which it influences the grid current.
(e) Increase the neutralizing capacitor slightly.
(f) Vary the anode tuning capacitor to determine whether neutralization has been achieved.
(g) If not, repeat steps (ii), (iii) and (iv) until the stage is correctly neutralized.

An alternative method of cold neutralization is to feed a signal to the anode of the valve and adjust the neutralizing capacitor for minimum indication on a valve voltmeter connected to the grid circuit.

Sometimes it is not possible to eliminate entirely interaction between the two circuits but this does not necessarily mean that the circuit will not function properly. However, the greatest care should always be taken to achieve the lowest possible degree of interaction.

In a push-pull amplifier the two neutralizing capacitors should be varied together (i.e. with similar values) so that an approximate balance is maintained during the process. The final correct balance may be found to occur with slightly different settings of the two capacitors but if the values are greatly different the balance may be upset when the operating frequency is changed to a different band. A well designed and constructed amplifier, when correctly neutralized, should remain so over the full range of frequencies 1·8–30 Mc/s.

Hot Method

This is carried out in a similar manner to that described for cold neutralization with the same object: to achieve the lowest possible interaction between the anode and grid circuits of the amplifier. As it is performed with anode and screen voltages applied to the valve, the greatest possible care must be taken to avoid coming into contact with the high voltage supplies. It is a good rule to keep one hand in your pocket while carrying out this type of adjustment. Before making any adjustment, the power should be switched off and the capacitors allowed to discharge; the h.t. line should then be shorted to earth. (See Chapter 20 for further advice on safety precautions).

A final check of the neutralization should be made by applying a low anode voltage, removing the grid drive, and adjusting the bias so that it produces a standing anode current corresponding to about half the maximum permissible anode dissipation. If the neutralizing setting is incorrect, self-oscillation is almost certain to occur. This can be detected in various ways, e.g. by the glow in a neon lamp held near the " hot " parts of the anode circuit, by the lighting of a flash-lamp bulb connected to a loop of wire when coupled to the tank coil, or by an absorption wavemeter. Anode-current variations will also be observed when the grid or anode tank capacitor is tuned through resonance.

If it appears impossible to achieve proper neutralization the cause of the difficulty may lie in the presence of parasitic oscillation. Methods for suppressing such oscillations are described on page 6.36.

When the amplifier has been neutralized, a suitable load should be coupled to the anode tank coil; the grid drive may then be applied though still with reduced anode and screen voltages. This load may consist of an ordinary domestic electric lamp of suitable power rating or preferably a carbon resistor of suitable rating to dissipate the power output (the Electrosil type H39, for example).

When h.t. is applied to the p.a. the anode tuning capacitor should be quickly tuned for maximum dip in the anode current to avoid damaging the valve. It can be assumed that the circuit is correctly loaded if the dip in anode current when the tank circuit is tuned through resonance is 10–15 per cent. If a lamp is used as the load it should be connected to the coil by the shortest practicable leads, the wires being soldered directly to the lamp-cap contacts.

After the amplifier has been tuned at reduced power, full power may be applied. A rough estimate of the power output may be made by observing the brilliance of the lamp. Finally the anode efficiency and dissipation should be calculated to ensure that the valve is being operated within its ratings.

If it is found that varying the anode tank capacitor about the resonance position causes the grid current to change sharply at any point, it is a sign that feedback is taking place and closer attention should be paid to the screening or to the neutralizing adjustments. Besides being troublesome in other ways such feedback can cause intermittent spurious emissions in phone operation and produce effects similar to those of over-modulation.

Sometimes the tuning adjustment that produces the maximum output from the power amplifier does not coincide with the tuning adjustment for minimum anode current. This can be due to faulty neutralization but it can sometimes be traced to a variation in the screen voltage as the circuit is tuned through resonance, and in this event the remedy is to stabilize the screen voltage.

Adjusting Transistor Power Amplifiers

It is essential to load the output circuit of a transistor power amplifier during the tuning procedure. Unless this is done, the peak collector voltage may exceed the collector-base rating and the transistor destroyed. A suitable low power dummy load is shown in **Fig. 6.79** although it is of course quite possible to use a high wattage carbon resistor of the type used as a dummy load for valve transmitters.

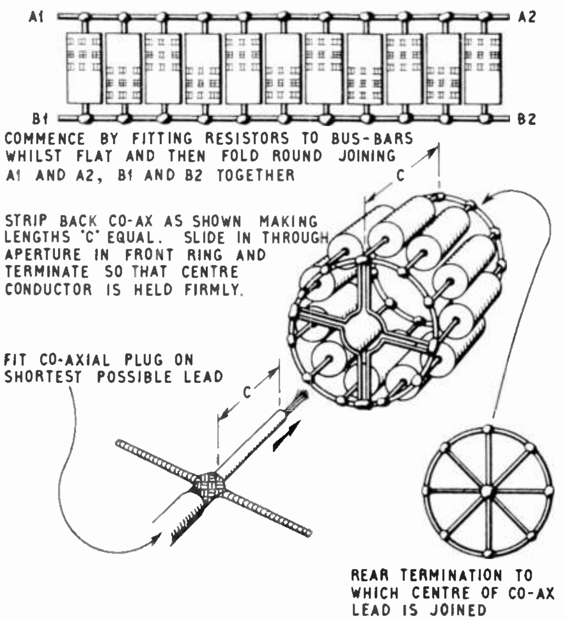


Fig. 6.79. A 75 ohm dummy load. For 10 watts dissipation, 11 820 ohm, 1 watt, 10 per cent carbon resistors should be fitted to form the " cage" ; for 50 watts, 25 1800 ohm, 2 watt, 10 per cent carbon resistors.

It is comparatively difficult to obtain the optimum output from a transistor p.a. with simple link coupling but a tuned link permits a reasonably high power output be to achieved. For 75 ohm feeder, the capacitances used to tune the link in a transistor p.a. should be three to four times those given in Table 6.8.

Apart from employing an oscilloscope to examine the waveform at the collector and a high resistance voltmeter across the biasing resistor there is no way of determining the circuit conditions in the case of an original design. When working to a published design it can be assumed that the necessary investigations have been undertaken: provided the right voltages are applied and the correct currents indicated, the transistor will be operating correctly.

In an original design, reduced drive should be applied initially. The drive is then adjusted by positioning the drive link until a voltage is indicated on the meter and the resulting r.f. waveform can be seen on the oscilloscope. The tank capacitor is then adjusted to peak the r.f. output voltage. The output link is adjusted to draw power from the tank circuit, and this will be reflected as a drop in the peak voltage indicated. The drive is then increased to a level which will indicate an angle of flow of about 150°, and the drive voltage checked to ensure that it is well below the base emitter rating of the transistor. The collector tank tuning and the output link tuning (or the position of the output link) are next checked for maximum output to the load by adjusting the tank for peak voltage and the link for a reduction in that voltage.

The drive is increased until the peak r.f. collector voltage rises to 75–80 per cent of the d.c. collector voltage at which point the loading is again checked for optimum power output. The drive is further increased until the peak r.f. collector voltage is $1{\cdot}25 \times V_c$, provided that the base-emitter voltage remains within its rating. The duty cycle of the p.a. transistor at this point will be short and it should operate for long periods without signs of overheating or a steadily rising collector current indicating thermal runaway. If either of these are experienced, the drive must be immediately reduced.

In view of the high harmonic content it is most desirable to follow a transistor with an aerial tuning unit to reduce the harmonics as far as possible.

POWER RATINGS OF R.F. AMPLIFIERS

The maximum permissible ratings at which a valve may be operated are specified in the valve manufacturers' published data. In the case of some US valves, the manufacturers specify two different ratings: (i) the CCS or *Continuous Commercial Service* rating, in which long life and consistent performance are the chief requirements, and (ii) the ICAS or *Intermittent Commercial and Amateur Service* rating in which high output from the valve is a more important consideration than long life. The term " intermittent " applies to ON periods (key down) not exceeding five minutes followed by OFF periods (key up) of the same or greater duration. This must be remembered if it becomes necessary, as for instance while testing, to keep the key closed so that the valve is then in *continuous* operation.

The following are the main features to be borne in mind when considering the power-handling capabilities of a valve.

Power Input

The power input to the valve is the d.c. power P_i fed to the anode circuit from the h.t. supply. It is given by—

$$Pi = \frac{I_a \times V_M}{1000} \text{ watts}$$

where I_i = anode current (mA)
V_M = anode voltage (V)

Power Output and Anode Efficiency

Not all of the h.t. power supplied to the valve is converted into r.f. energy; part is expended in heating the anode. The power-handling capacity of the valve is limited by the maximum permissible dissipation of heat by the anode through radiation and conduction. The anode efficiency is given by—

$$\text{Efficiency} = \frac{P_o}{P_i} \times 100 \text{ per cent}$$

where P_o = r.f. power output
P_i = d.c. power input to the anode circuit

The maximum r.f. power output obtainable from a valve may be predicted from a knowledge of the anode dissipation. For example, consider a valve with a maximum permissible anode dissipation of 12 watts, operating in class C, the efficiency being about 70 per cent. Of the d.c. power input, 70 per cent will appear as r.f. energy and 30 per cent as heat which must be dissipated by the valve anode. This amount of heat is limited by the valve rating to 12 watts. Thus—

$$\textit{Maximum dissipation} = 12 \text{ watts} = \frac{30}{100} \times \text{d.c. input power}$$

Therefore—

$$\textit{Maximum power input} = 12 \times \frac{100}{30} = 40 \text{ watts}$$

provided the maker's ratings are not exceeded.

For two such valves operating under the same conditions, either in parallel or in push-pull, the maximum power input would be 80 watts. If the anode efficiency is less than 70 per cent, the maximum power output would be correspondingly lower.

In a frequency multiplier, the anode efficiency will be substantially lower, and the maximum permitted d.c. input will be correspondingly reduced. In a frequency doubler, the efficiency is likely to be 50 per cent, leaving 50 per cent to be dissipated in the anode. In a tripler, as the output efficiency is likely to be 30 per cent, the anode power loss will be 70 per cent.

To arrive at the maximum permitted d.c. input for a frequency multiplier, the equation above should be modified so that the power loss figure is substitued as follows:

$$\text{Maximum d.c. input} = \text{Anode dissipation (Watts)} \times \frac{100}{A_{pl}}$$

where A_{pl} is the percentage anode dissipation, i.e. doubler 50 per cent, tripler 70 per cent.

Screen Dissipation

All the d.c. power supplied to the screen circuit is expended in heating the screen. Since the screen is constructed of relatively thin wire mesh, the valve may be permanently damaged if the permissible dissipation is exceeded even for a relatively short period.

When the screen grid is fed with h.t. from a separate supply provision must be made for it to be removed immediately if the anode supply fails. If screen grid voltage is applied in the absence of anode voltage, the screen will be burnt out.

Screen Voltage Supply

There are three common methods of supplying the screen grid voltage for a tetrode or pentode valve:

(i) by a voltage dropping resistor in series with the anode supply;

(ii) from a potential divider connected across the anode supply.

(iii) from a separate supply.

In the case of the series resistor, the voltage regulation of the screen grid is relatively poor due to variations in the anode supply voltage with variations in the valve loading, and consequential variations in the screen grid current. The advantage of the series resistor method is that since the value of the series resistor is normally high, this limits the maximum current of the screen grid, and so gives some degree of automatic protection. Some valves, however, will draw negative screen current; in such cases a series resistor alone is useless.

In the case of a supply derived from a potential divider across the h.t. supply, the current through the network is substantially higher than that taken by the screen grid, and this provides a measure of voltage regulation.

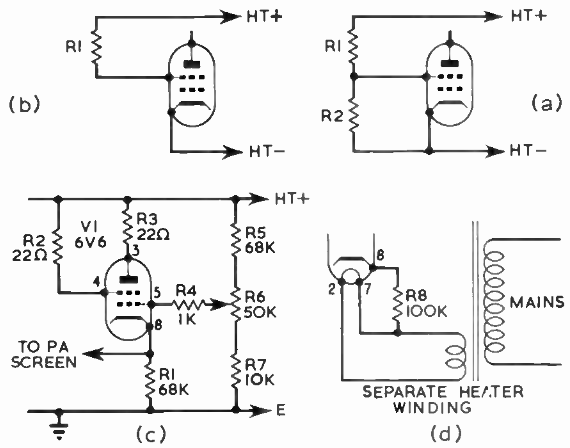


Fig. 6.80. (a) and (b). Calculation of resistances in screen voltage circuits. (c). Cathode follower p.a. screen supply circuit. (d) Heater supply for the cathode follower valve.

Methods (i) and (ii) are illustrated in **Fig. 6.80(a)** and **(b).** If the screen voltage is applied through a series dropping resistor, as shown at (A) in Fig. 6.80, the calculation of the correct resistor value is very simple. Consider, for example, the requirements for a typical p.a. valve:

Screen voltage	=	250 volts
Screen current	=	6 mA
Supply line voltage	=	750 volts

The series resistor must drop 750 — 250 volts (i.e. 500 volts) for a current of 6 mA. Therefore its resistance must be—

$$R1 = \frac{750 - 250}{0{\cdot}006} = 83{,}000 \text{ ohms}$$

Under these conditions the power dissipated in the resistor will be (750 — 250) V × 6 mA = 3 watts.

Fig. 6.80(b) shows the voltage supply to the screen grid derived from a potential divider. The design procedure is best explained by taking the same example. The resistor R1 in this case carries the screen grid current and the current through R2. Assume that the total current through R1 is three times the screen grid current (i.e. 18 mA). Then R1 can be calculated:

$$R1 = \frac{750 - 250}{18} = 27{\cdot}8 \text{ K ohms}$$

The resistor R2 carries the total current minus the screen grid current (i.e. 12 mA). So R2 can be calculated:

$$R2 = \frac{250}{12} = 20{\cdot}8 \text{ K ohms}$$

The power dissipated in each resistor is easily calculated using the formula $P = \frac{V^2}{R}$.

$$\text{Power dissipated in R1} = \frac{(500)^2}{27800} = 9 \text{ watts}$$

$$\text{Power dissipated in R2} = \frac{(250)^2}{20800} = 3 \text{ watts.}$$

Resistors of the nearest preferred value should be chosen with wattage ratings of double the above calculated power dissipations. The higher current consumption of the potential divider method is the price paid for improved regulation.

Fig. 6.80(c) shows a cathode follower circuit which provides a variable screen grid voltage with very good regulation. The output resistance of a cathode follower is low, being approximately $1/g_m$ so current variations affect the voltage output only slightly. The output voltage can be easily adjusted by varying the control grid voltage of the cathode follower. Most of the common triodes and pentodes may be used, though it should be remembered that the valve selected must be capable of carrying more than the p.a. screen current. A separate heater supply should be provided for the cathode follower with the cathode connected to one side of the heater through a 100 K ohm resistor. This eliminates the danger of breakdown of the heater to cathode insulation.

The cathode follower circuit may be modified to perform a number of other functions; see, for example the Series Gate Modulator in Chapter 9—*Modulation*.

Capacitor Ratings

A fixed capacitor which is used in r.f. circuits where any loss of power must be avoided should have a mica or ceramic dielectric. These types have the required low dielectric loss, low inductance values and high insulation resistance. Low inductance is a particularly important requirement of decoupling and other bypass capacitors since they should have negligible reactance over the entire working frequency range (and above). It is recommended that feed-through or other disc types be used for low-power circuits where the h.t. supply does not exceed about 500 volts. For normal h.f. transmitter use, the values of decoupling, d.c. blocking and r.f. coupling capacitors lie within the range 1000 pF to 0·01 μF.

The inductance of capacitors and hence their resonant frequency can be modified by mounting them so that their leads enclose as small an area as possible. With tubular types, mounting them in earthed tubes will reduce the inductance. If a capacitor is series resonant with its leads at or near to the operating frequency, it is ideal for bypassing the screen grid and cathode, particularly at v.h.f.

The self-resonant frequencies of a number of British fixed capacitors commonly used in amateur transmitters are given in Table 6.7.

To allow a margin of safety, the voltage rating of a capacitor should be at least 50 per cent greater than the peak voltage which is likely to occur across it, and its size should be adequate to carry the r.f. current without overheating.

Grid Coupling Capacitors. The peak applied voltage is approximately the sum of the grid bias voltage of the driven stage and the d.c. anode voltage of the driver stage; the voltage rating of a grid coupling capacitor should therefore be about 50 per cent higher than this peak value. Since some portion of the circulating current passes through the capacitor the current rating must be adequate.

Grid Tuning Capacitors. The peak voltage across the grid tank capacitor depends on the circuit arrangement and is indicated in **Fig. 6.81**. In some cases the grid bias voltage must be added to the drive voltage.

Screen Bypass Capacitors. The voltage rating should be 1·5 times the anode voltage, or 2·5–3 times this value when anode-and-screen modulation is used. The current rating must be also considered because some or all of the circuit current can flow through the bypass capacitor and $C_{a\text{-}sg}$ in series.

Neutralizing Capacitors. The neutralizing capacitor is subjected to peak voltages equal to the sum of the peak grid voltage and the peak anode voltage, amounting in the case of an unmodulated amplifier to about two to three times the d.c. anode voltage, and for a modulated amplifier about twice this value.

Anode Decoupling and Blocking Capacitors. In c.w. transmitters the peak voltage across the anode decoupling and blocking capacitors is equal to the d.c. anode voltage; in telephony operation it rises to at least twice this value on modulation.

Anode Tank Capacitors (including Pi-networks). The peak voltage across an anode tank capacitor depends on the arrangement of the circuit, e.g. single-ended, split-stator or push-pull. The values given in **Fig. 6.82** assume that the amplifier is correctly loaded. In circuits where the d.c. supply voltage also exists across the capacitor, as at (C) and (F), twice the rating is required, and it is therefore an advantage in high-power stages to remove this voltage from the capacitor by the use of parallel feed or other suitable arrangements. The application of 100 per cent modulation doubles the peak voltages occurring in any of the above circuits.

Although the necessary plate-spacing of an air-spaced capacitor for a given peak voltage rating depends to some extent upon constructional features such as the shape of the plates and the rounding of their edges, a useful guide

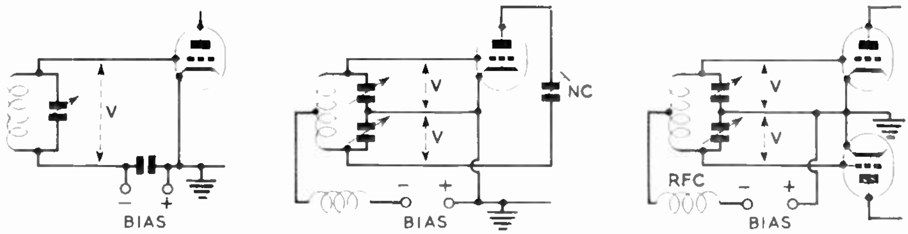


Fig. 6.81. Grid-tank capacitor voltage ratings. The peak voltage, V, across each section is equal to the bias voltage plus 25 per cent of the applied d.c. anode voltage.

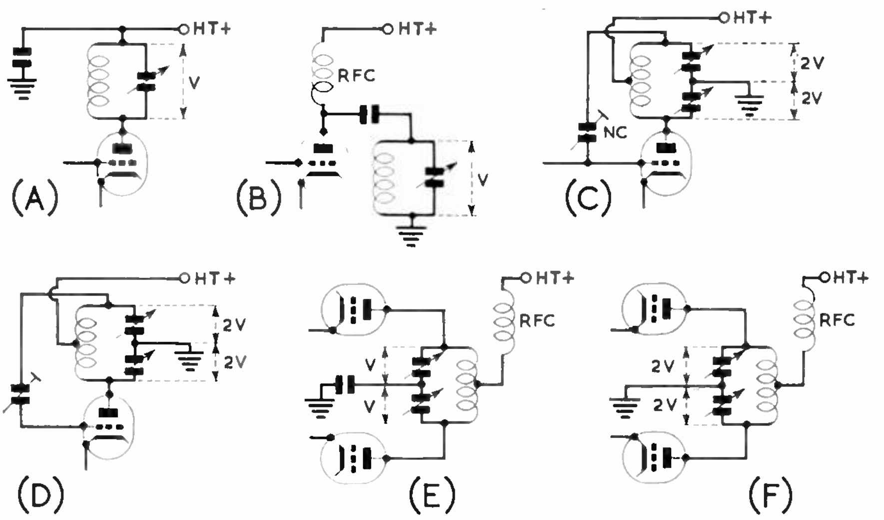


Fig. 6.82. Anode-tank capacitor voltage ratings. The peak voltage may be taken as 80-90 per cent of the h.t. supply voltage.

the p.a. stage is assumed to be 70 per cent, the peak power output will not exceed 420 watts. The peak voltage occuring across the pi-network output capacitor would therefore be—

$$E = \sqrt{2 \times 420 \times 75}$$
$$= 264 \text{ volts}$$

Pi-tank Tuning Capacitor

The tuning capacitor of a pi-tank circuit in a c.w. transmitter must be rated at not less than 1·25 times the d.c. voltage of the anode, and for a.m. and s.s.b., at not less than 2·5 times the d.c. anode voltage.

to the voltage at which break-down occurs for different values of spacing will be found in Fig. 6.83.

Pi-network Output (Loading) Capacitors. The peak voltage E occurring across a pi-network output capacitor is given by—

$$E = \sqrt{2P_oR_2} \text{ volts}$$

where P_0 = power output of the amplifier (watts)
R_2 = resistive component of the load or aerial impedance.

With a low-impedance load (e.g. a coaxial line not exceeding 100 ohms) the peak voltage is not likely to be higher

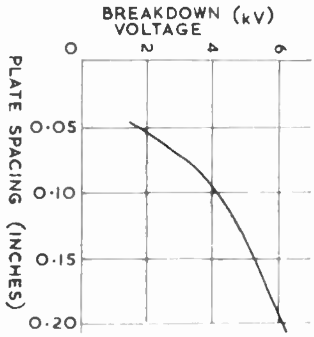


Fig. 6.83. Breakdown voltage of air-spaced capacitors.

than about 300 volts, even in a 150 watt modulated transmitter, and a receiving-type variable capacitor should be quite satisfactory. In many cases it may be convenient to use a good-quality three or four gang receiving-type capacitor to obtain a maximum capacitance of 1500–2000 pF, which may be necessary on 1·8 Mc/s and 3·5 Mc/s. The sections may be switched out progressively on changing to operation on the higher-frequency bands.

Example. *A 150 watt phone transmitter is operating into a 75 ohm line through a pi-network coupler. What is the peak voltage across the output capacitor?*

Since the peak power input occurring on modulation peaks is 4 × 150 = 600 watts, and if the anode efficiency of

METERING

Although it is necessary to meter the current and/or the voltage at several different parts of a transmitter to ensure that all the various stages are operating correctly, an economy of instruments is made possible by the fact that one single reading (of the anode current in the p.a. stage, for example) is usually all that is required for noting quickly whether the transmitter is operating normally. The simplest arrangement—to measure grid and anode current—is illustrated in Fig. 6.84.

A single instrument of about 1 mA f.s.d. in conjunction with a multi-way switch to allow connection to the appropriate points is therefore often found most suitable. Such an arrangement is shown in Fig. 6.85.

The resistors R1, R2 and R3 in Fig. 6.85 are the shunt resistors necessary to give a convenient range-multiplication factor for the current to be measured. They are left permanently in the circuit to be monitored but owing to their low resistance values, the operation of the circuit is not affected. On the voltage ranges, the series resistors R4, R5 and R6 will likewise be selected to give a convenient full-scale deflection for the meter. The actual values are calculated as suggested in Chapter 19 (*Measurements*).

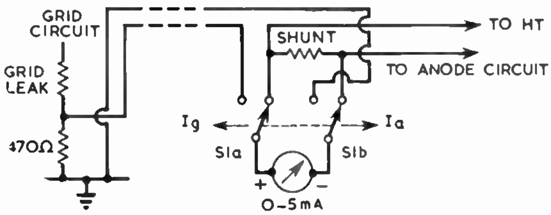


Fig. 6.84. Meter switching circuit measuring the anode and grid currents in a low power p.a. stage. The switch S1 should be arranged to have a spare earthed position between the two ranges to avoid arcing and damage to the meter. In medium and high power transmitters it is advisable to meter the cathode current of a p.a. rather than the anode current.

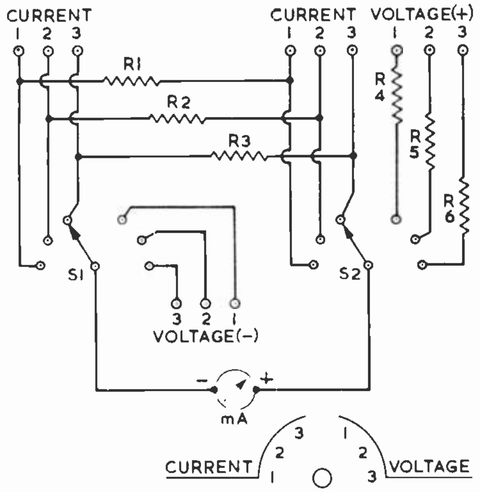


Fig. 6.85. A simple switching system enabling one meter to be used for measuring three different currents and three different voltages. The switches S1 and S2 can be of the ganged-wafer type. The resistors R1, R2, R3 (current range shunts) and R4, R5, R6 voltage-range series resistors) are selected according to the required ranges and the full-scale current of the meter.

The selector switch should be of the wafer type, with ceramic insulation for the higher voltage circuits; the contacts should be of the break-before-make type to avoid a short-circuit when changing from one circuit to the next. For voltages above about 300 volts it may be necessary to use double spacing between adjacent contacts to prevent flash-over, and the circuit should be arranged, where possible, so that adjacent contacts are at approximately similar potentials.

As an alternative to switching, closed-circuit jacks are sometimes used. These are jacks which keep the circuit closed except when the plug is inserted. The advantage of the method is that an independent external meter may be used but this advantage is outweighed by the simplicity and convenience of the switched meter arrangement. It must be remembered that spurious radiation is reduced when the meter and the associated wiring are completely contained within the transmitter enclosure.

TRANSMITTER DESIGN SUMMARY

The following summary is intended to help in the design and construction of amateur transmitters for use on the h.f. (1·8—28 Mc/s) bands:

(1) If the transmitter is to be v.f.o. controlled, the v.f.o. should be constructed as a separate unit and subsequently fitted to the main transmitter assembly.

(2) A v.f.o. must operate satisfactorily and be thoroughly checked before it is fitted to the transmitter. Final testing should be carried out in the complete transmitter.

(3) Only the highest quality components, rigidly mounted, and positioned for minimum temperature excursions should be used in the construction of a v.f.o.

(4) Where temperature compensation is needed, the adjustments should be made after the complete transmitter has reached its correct operating temperature.

(5) A v.f.o. must be followed by a suitable buffer amplifier.

(6) Frequency multiplier stages and driver amplifiers should be operated at a power level which is just sufficient to provide the required drive to the following stage.

(7) In the interests of reducing the radiation of unwanted harmonics of the v.f.o. or crystal oscillator, capacity coupling between the driver and the p.a. should be avoided.

(8) In high power designs, bandpass couplers between the stages significantly reduce the level of the harmonic output. While this is also true of low power transmitters, such couplers may not be needed to reduce harmonic radiation to an acceptable level in view of the lower power.

(9) A pi-network tank circuit tends to limit harmonic radiation.

(10) In certain areas where the television signals are weak an aerial coupler should be employed as an aid to harmonic reduction whatever type of p.a. tank circuit is employed.

(11) Provision for neutralizing should be included in all p.a. stages.

(12) Power amplifiers should be thoroughly checked for low frequency or high frequency parasitic oscillation.

(13) In power amplifiers employing grid-leak bias, some form of safety bias should always be included.

(14) It is preferable to construct the transmitter as a single unit in the interests of reducing stray radiation, but care must be taken to provide sufficient ventilation.

(15) If the transmitter is constructed as two units, one of which is the power supply, all leads leaving the transmitter enclosure must be fully decoupled at the point at which they leave the cabinet.

(16) The physical layout should be as near to a straight line as possible to ensure maximum separation between the master oscillator and the power amplifier.

(17) Both the v.f.o. and the p.a. should be thoroughly screened. All power supply leads, including heater circuits, to the v.f.o. should be decoupled at the point where they enter the v.f.o. box. This will help to prevent feedback from the power amplifier.

(18) Tuned circuits should be positioned so that they do not exhibit mutual coupling one to the other, unless this is intended.

(19) Always ensure that the h.t. supplies are properly fused.

(20) In the interests of personal safety, never make adjustments to the transmitter, and especially the power amplifier, while h.t. is applied.

TRANSMITTER CONSTRUCTION

The circuits given in this section are representative of current amateur constructional practice and provide a broad indication of present trends in the design of high frequency amateur transmitters. While layout plans are not given for all the projects described, no particular difficulties should be encountered provided the layout follows a logical sequence.

A 10 WATT TRANSISTOR TRANSMITTER FOR 1·8 MC/S

The transistor transmitter is v.f.o. controlled and has provision for both phone and c.w. operation. It is

Transistor transmitter for 1·8-2 Mc/s.

suitable for fixed or mobile use provided the d.c. supply line is suitably filtered.

Silicon *n-p-n* transistors are used in the r.f. stages and *p-n-p* types in the modulator and speech amplifier. The collectors of the r.f. transistors are at earth potential, i.e. connected to the earthed positive supply. The p.a. employs three transistors of the same type in the common emitter configuration: three RCA 2N3053 or three Mullard BFY51. The transmitter is constructed in an Eddystone diecast box which acts as a heat sink for the p.a. transistors, the collectors of which are connected internally to the cases. The arrangement can, therefore, only be used with a positive earthed supply.

The circuit is shown in **Fig. 6.86** from which it will be seen that a double wound transformer T4 is used to drive TR8, TR9 and TR10. The emitters of these transistors are connected to a tap on the tank coil T5.

A parallel tuned circuit is used in which the p.a. transistors are tapped down the tank coil so that the coil acts as an impedance transformer. The actual tapping point is on the middle turn of the coil to give the tightest possible coupling. T5 is tuned by a 900 pF variable capacitor C30 with a 1000 pF fixed capacitor C31 in parallel. This combination provides coverage of the 1·8–2 Mc/s band with some margin for tuning out reactive loads. The aerial is coupled to the tank coil by a four turn link.

Without bias, the p.a. transistors draw no current and the bias for class C operation is provided by resistors in the emitter connections. This also helps to equalize the currents through TR8, TR9 and TR10.

For 10 watts input at 11 volts (a volt or so is lost in the

TABLE 6.9

Components for the 10W Transistor Transmitter

C1, 11, 21, 31	1000 pF mica or polystyrene.
C2, 16	10,000 pF 30 volt metallized plastic.
C3	100 μF 12 volt electrolytic.
C4, 18	0·4 μF metallized plastic.
C5, 6, 36	100μF 3 volt electrolytic.
C7	22,000 pF 30 volt metallized plastic.
C8, 28	1 μF metallized plastic.
C9	220 pF ceramic, —750 p.p.m./°C.
C10, 17, 19, 22-27, 32, 34, 35	0·1 μF 30 volt metallized plastic.
C11, 15	680 pF polystyrene.
C12	176 pF + 208 pF in parallel, airspaced variable, with 6:1 epicyclic drive.
C13, 14, 20	6800 pF mica or polystyrene.
C29	0·1 μF plastic and foil.
C30	450 pF + 450 pF in parallel, airspaced variable.
C33	250 μF 12 volt electrolytic (in insulating sleeve).
CR1	OA5 or similar germanium diode.
CR2, 3	OA200 or similar silicon diode.
CR4	9·1 volt, 200 mW Zener diode.
F51	2 amp fuse.
J1	Igranic miniature jack socket with short-circuiting contact and additional normally open contact.
L1	Osmor QO4 with windings in series aiding, in 3/4 in. square i.f.t. can.
LP1	Miniature lamp, 12 volt, 60 mA.
M1	Moving coil meter, 0–1 amp d.c., 1 31/32 in. square.
R1	100 K ohms.
R2	68 K ohms.
R3	150 K ohms.
R4, 7	18 K ohms.
R5	3·9 K ohms.
R6, 18	1·8 K ohms.
R8, 15	4·7 K ohms.
R9, 11, 12	1 K ohms.
R10	470 ohms.
R13	0·5 ohm (see text).
R14	330 K ohms.
R16, 17	15 K ohms.
R19	330 ohms.
R20	150 ohms.
R21	820 ohms.
R22	27 ohms.
R23, 24, 25	3·9 ohms, 1 1/2 watt wirewound.
Rx, Ry	See text.
VR1	50 K ohms log. potentiometer.
All resistors 1/8 watt, 10 per cent tolerance, unless otherwise stated.	
RFC1, 2, 3	40 turns 28 s.w.g. on 1/4 in. o.d. ferrite tube, 1 1/4 in. long, or Henry's Radio type RFC8, 5 μH.
RFC4	1·5 mH, Henry's Radio type RFC.
Switch	Plessey push-button assembly, three interlocking buttons, each with three changeover contacts.
T1	1/4 in stack No. 450 0·015 in. Radiometal laminations butted (Coded grey, 1 in. × 3/4 in. overall, 1/4 in. centre limb). *Primary:* 1120 turns 43 s.w.g. *Secondary:* 200 + 200 turns 38 s.w.g., bifilar wound, or Ardente D3052.
T2	7/8 in. stack No. 187 0·018 in. silicon iron laminations butted (2 1/4 in. × 2 3/4 in. overall, 1 in. centre limb). 180 + 180 turns, 23 s.w.g., bifilar wound.
T3	11 mm. ferrite pot core, adjustable. *Primary:* 20 turns, 30 s.w.g. *Secondary:* 10 turns, 30 s.w.g.
T4	11 mm. ferrite pot core, adjustable. *Primary:* 6 turns, 30 s.w.g. *Secondary:* 3 turns, 24 s.w.g.
T5	Ferrite rod, 1/2 in. diameter, 1 1/2 in. long. *Primary:* 9 turns, 19 s.w.g., starting 3/16 in. from one end, tapped at 3 1/2 turns and 4 1/2 turns. *Secondary:* 4 turns, 22 s.w.g., on former sliding along finish end of core.
TR1	OC202 or 2S324.
TR2, 3, 11	OC82DM, GET104 or 2GT174.
TR4, 5	OC35, AD140 or 2PGT7.
TR6	2N706, BFY51 or 2N3053.
TR7–10	BFY51, 2N3053 (see text).

modulation transformer and r.f. chokes) the mean p.a. current is 0·9 amp divided between the three transistors. The instantaneous peak current when fully modulated is six times greater. The current gain of the transistors falls rapidly with increasing current in this region and to prevent flattening of the modulation peaks (" downward modulation ") the driver is also modulated.

Due to high currents and low voltages the decoupling capacitors must have very low impedances. The bias resistors are each decoupled by two parallel 0·1 μF metallized plastic capacitors. The p.a. supply decoupling, through which the whole p.a. r.f. current must flow, comprises a 1 μF metallized plastic capacitor C28 in parallel with a 0·1 μF foil and plastic unit, C29. Metallized capacitors are only just good enough for use at this frequency as the resistance of the metallizing film which appears in series with the capacitor is comparable with the reactance.

The p.a. tank circuit must match the output transistors to their load and have a Q of about 10 or 20 to attenuate harmonics without introducing excessive losses due to high circulating currents in the tank circuit. The output impedance of this p.a., calculated exactly as for a class C valve p.a., is about 6 ohms.

A Faraday screen between the tank coil and the aerial link helps to reduce the harmonic output.

The driver stage TR7 must supply about 100 mW to the p.a., the input impedance of which is 50 ohms. This is supplied via a tuned step-down transformer T4 wound on an adjustable ferrite pot core. Both this core and that on the driver input T3 are set to the middle of the band and do not require further adjustment. The driver transistor, a 2N3053, operates in class A at 40 mA collector current. At this lower current a small push-on heatsink is adequate, allowing the output circuit to be in its usual place in the collector lead.

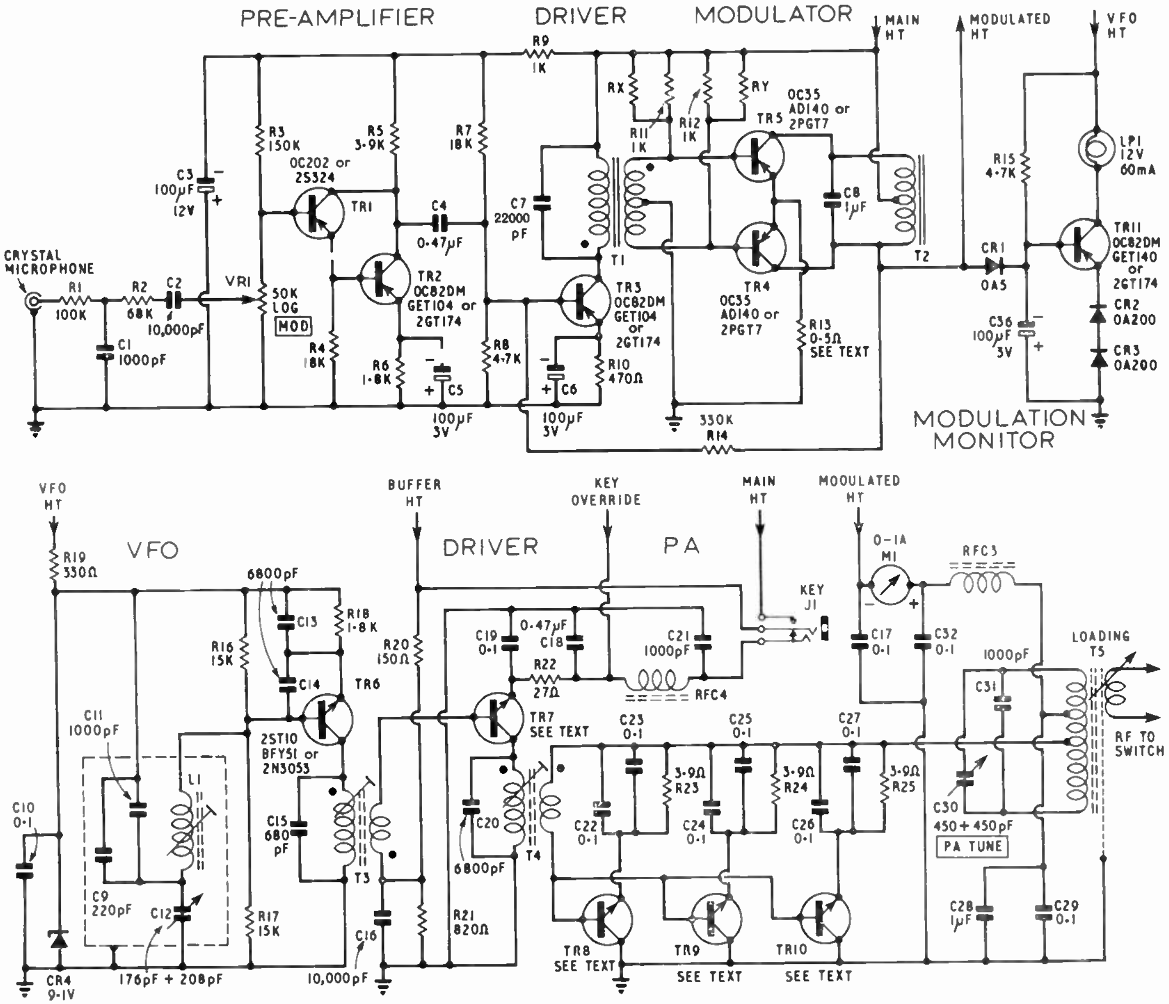


Fig. 6.86. The audio and r.f. sections of the 10 watt transistor transmitter for 1·8-2 Mc/s.

On c.w., the driver is keyed, and on NET it is switched on to make the v.f.o. audible.

The oscillator TR6 is designed so that either a series tuned circuit or a crystal may be used, although a crystal may oscillate only if C13 and C14 are reduced to about 1500 pF. The oscillator circuit is connected between base and emitter, and the second harmonic is extracted from a tuned transformer in the collector circuit, thus providing a considerable degree of isolation. If a crystal is used, it may be either fundamental or sub-harmonic. The supply voltage is stabilized by a Zener diode. At 1·9 Mc/s the change in frequency between NET and SEND is about 70 c/s and the frequency drift during the first 10 minutes after switching on from absolute " cold " is just under 200 c/s.

A g.d.o. is essential for checking the r.f. coils before assembly. If ferrite cores other than those specified are used, the number of turns may need to be altered to obtain the required resonant frequency, keeping the same turns ratios and the same tuning capacitors.

As the modulator and the p.a. share the same supply and have the same impedances, an autotransformer or tapped choke is used to couple the push-pull modulator output transistors TR4 and TR5 to the p.a. This minimizes the bulk of a transformer, although it is still the biggest component in the transmitter. A 15 watt mains transformer is suitable and is easily rewound by hand. The wire should be split on to two reels so that the two halves of the winding can be wound together. Wind 180 turns of the double wire and then connect the start of one wire to the finish of the other. There is considerable d.c. unbalance in the windings and to avoid saturation the laminations are gapped by simply not interleaving them. The same technique is used in the driver transformer T1.

The modulator operates in class AB1 and can deliver 6 watts output. The standing currents in TR4 and TR5 are adjusted by selecting RY and RX until 30 mA is flowing in each collector with no audio input. The two resistors must be adjusted repeatedly until both are correct, as they interact. On modulation peaks the total collector current will rise to about 900 mA.

Negative feedback is provided from the modulation transformer T2 to the base of TR3. If the sense of the connections to T2 is not as shown, the amplifier will oscillate. C7 and C8 ensure stability and restrict the high-frequency response to eliminate unnecessary sidebands.

The pre-amplifier is designed for use with a crystal microphone, and to obtain the required high current gain an emitter follower, TR1, precedes the amplifier, TR2. The unusual arrangement of the MODULATION control VR1 avoids disturbance of the d.c. voltage on TR1 base as the

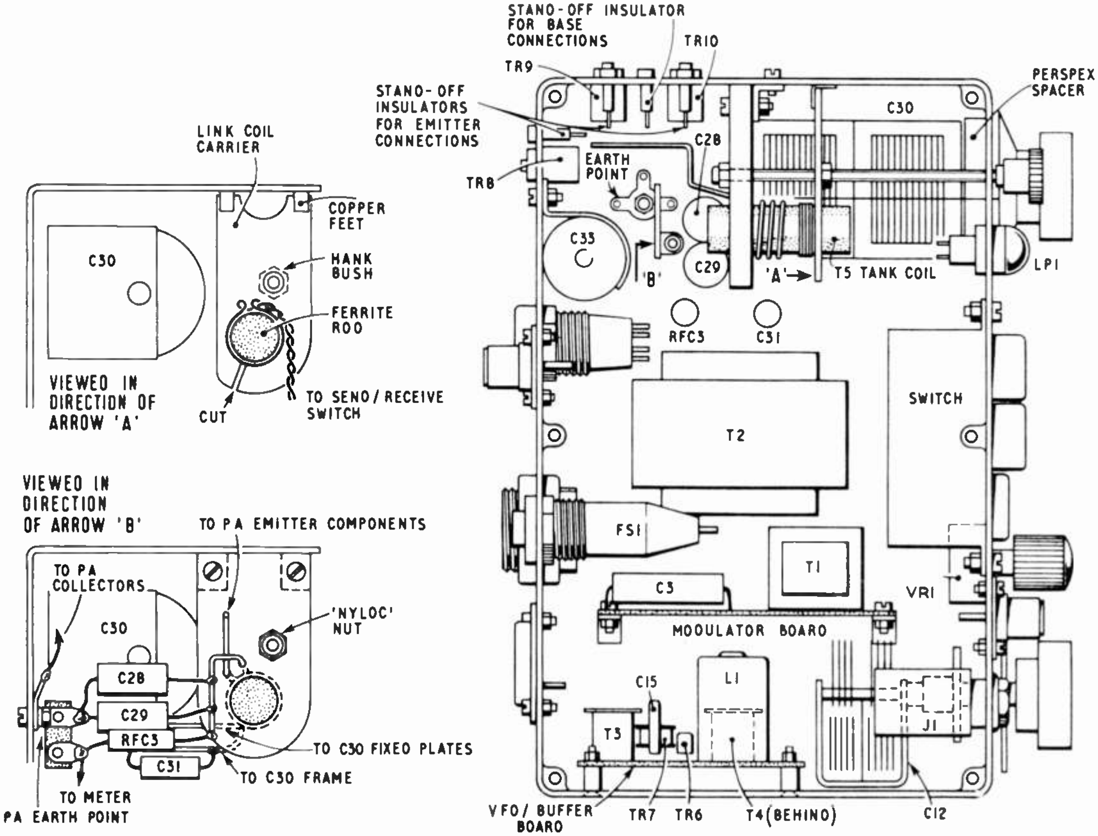


Fig. 6.87. Details of the p.a. assembly (left) and arrangement of the principal components inside the die-cast box used as a chassis.

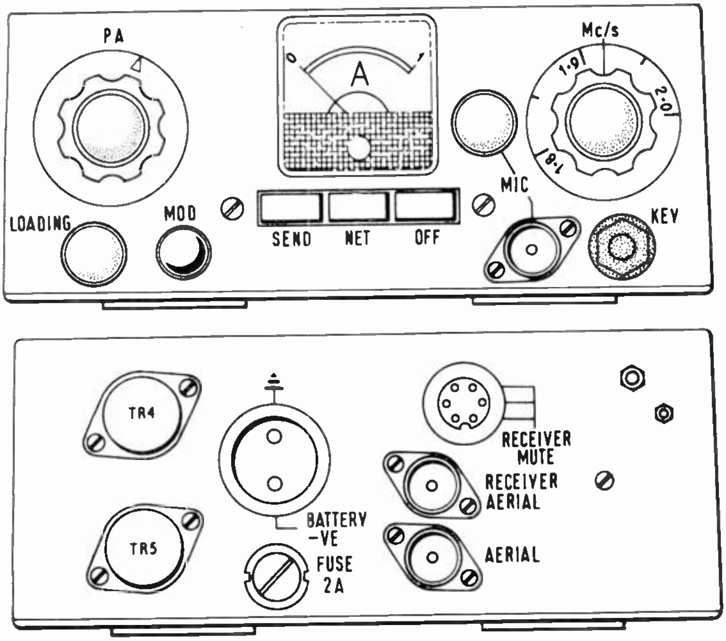


Fig. 6.88. Layout of the components on the front and rear of the die-cast box. The modulator transistors TR4 and TR5 must be insulated from the box but the p.a. transistors TR8, TR9 and TR10 may be mounted directly on the metal. The nut at the right-hand corner of the lower diagram holds the mounting bolt for TR8.

control is adjusted. As a crystal microphone behaves as a voltage in series with a 2000 pF capacitor, a degree of bass cut suitable for communication purposes is obtained with a load of 200 K ohms. This load consists of R1 and R2 in series with the r.f. bypass capacitor C1 connected at their junction. Oscillation of the modulator is possible if the earth connections are not grouped as shown in Fig. 6.87.

A simple modulation monitor is included, using a transistor, TR11, connected in series with LP1. TR11 is normally bottomed by current flowing through R15 from its base to h.t., but if a downward modulation peak is sufficient to make CR1 conduct, the base of TR11 is driven positive, turning off the lamp for a period dependent on C36. The diodes CR2 and CR3 are used in their forward direction as a low voltage Zener diode to bias the emitter of TR11 about 1·5 volts positive, so that the indicator lamp is extinguished if the supply to the p.a. falls below 1·5 volts on modulation peaks.

Construction

The transmitter is constructed inside an Eddystone 7½ in. × 4¾ in. × 3¼ in. diecast box which is used upside down so that the lid becomes the bottom cover. The positions of the principal components are shown in Figs. 6.87 and 6.88. The modulation transformer T2 is placed in the centre of the box so that it separates the p.a. from the low level r.f. circuits and the modulator.

The v.f.o., driver and modulation monitor are assembled on a piece of Lektrokit board bolted to the side of the box on spacers, so that the cores of L1, T3 and T4 may be adjusted through access holes in the side of the box. The v.f.o. is earthed directly to the tuning capacitor, the frame of which is bolted to the box. The r.f. driver stage is separately earthed to the box via one of the board fixing screws. An epicyclic drive is used for the v.f.o. tuning, with a round Perspex dial and an unskirted knob to match the skirted p.a. knob. A twisted pair of wires takes the r.f. output from T4 to the p.a. stage.

The modulator output transistors are bolted to the rear of the box, which acts as a heatsink. They must be insulated from the box with bushes and mica washers smeared with silicone grease. Their emitter resistor R13 is a length of resistance wire.

The r.f. filter on the microphone input and the gain control are behind the front panel, and all remaining modulator components, including the driver transformer, are on a second Lektrokit board between the v.f.o. and the modulation transformer.

As the p.a. transistors have their collectors earthed they may be fitted in the type of heatsink having a single bolt fixing, or merely pressed into holes drilled in the box (these holes must be exactly the right size). They are grouped in the rear corner of the p.a. compartment. The small components in the p.a. stage are fixed on miniature standoff insulators, with the earth connections taken to a star tag.

The p.a. tank coil is wound on a short piece of ½ in. ferrite rod. The ¼ in. thick support has a hole filed in it to fit over the winding, and the end of the coil and the core are fixed into the hole with Araldite. The rear connection goes to the frame of the tuning capacitor, and leads from the tapping points through the small holes in the paxolin support. The front end of the winding goes directly to the fixed vanes of the capacitor (it is particularly important to keep leads short in the p.a. stage). About ⅝ in. of the core is left exposed at the front for the aerial coupling link, which is mounted on a carrier of $\frac{1}{16}$ in. thick printed circuit board which can be moved along the axis of the coil by a 2 BA threaded rod rotated by the LOADING knob. A nut to accept the studding is soldered to the panel side of the circuit board. The carrier has copper foil feet which slide along the side of the box, earthing the copper surface and preventing it from rotating.

The aerial link is wound on a former of copper foil that is soldered to the carrier, acting as a Faraday screen. The foil must have a gap in it to avoid a short circuited turn, and at this point the carrier surface must also be slotted. The tank coil support and link carrier are both shaped to clear the tuning capacitor frame. The frame of C30 is insulated from the box by a Perspex spacer and nylon mounting screws; the voltage is low, therefore the spindle can pass through a ½ in. clearance hole and the knob be fitted in the usual way.

To avoid losses due to the proximity of the p.a. tank coil to the bottom plate, a 2 in. × 2½ in. hole is made and

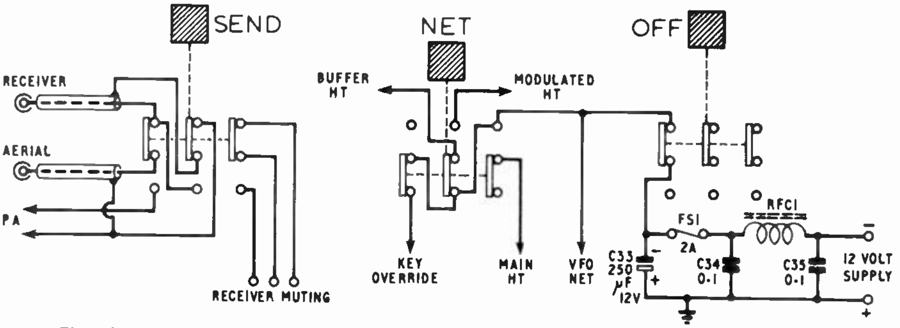


Fig. 6.89. Wiring of the control switch and filtering of the power input from the battery.

covered by a piece of Veroboard (perforated paxolin with copper strips on it). The strips of copper touch the lid on one edge only so that they act as a screen without allowing eddy currents to circulate. The Veroboard is covered by a piece of $\frac{1}{16}$ in. sheet rubber and a matching piece of the same material fitted at the other end of the lid to provide non-slip feet.

The components on the modulator and v.f.o. Lektrokit boards should be arranged as though on a printed board with the absolute minimum of crossed leads, and then wired up as far as possible with the component leads themselves or tinned copper wire. The essential crossing wires are then fitted with sleeving. Loops of tinned copper wire at the board edges may be used as terminals.

If there is any possibility of the 12 volt supply being plugged in the wrong way round, a 2 amp silicon diode should be wired in series with the fuse.

To fit the push-button switch in the centre of the panel it is necessary to remove part of the strengthening rib of the die-cast box.

Switching and Keying

Send-receive switching is provided by a triple push button unit which is wired as shown in **Fig. 6.89.**

In the OFF position the transmitter is turned off and the aerial is connected to the receiver. On NET the supply is connected to the oscillator and buffer and the key is shorted. When the SEND button is pressed the receiver is muted and its aerial terminals shorted out; the aerial is connected to the p.a. link coil. Slightly later, when the interlocking bar releases the NET button, the h.t. supply to the p.a. is connected and the transmitter is on the air. When either the NET or OFF buttons are pushed in, the action of the link bar again ensures that h.t. is removed just before the receiver is reconnected, avoiding any possibility of transmitter power blocking the receiver. If it is desired to tune up the transmitter off load, the SEND button is partly depressed to release the other button and then itself released, so that all three are out. This puts power on to the transmitter without changing over the aerial or muting the receiver.

The key is connected in the emitter of the buffer stage via a key click filter RFC4 C21 which gives rise and fall times of about a tenth of a millisecond. An additional contact on the key jack short-circuits the modulation transformer on c.w. to prevent the generation of voltage transients in the modulator and to avoid the voltage drop owing to the resistance of the transformer winding.

Alignment and Testing

When switched to NET the current consumption should be about 100 mA. Adjust the core of L1 so that the v.f.o. tunes from 0·9 to 1 Mc/s, and mark the scale 1·8 to 2 Mc/s. Press the SEND button, and check that the current rises to about 1·1 amp. The panel meter will indicate about 0·6 amp. Set the cores of T3 and T4 to give maximum meter current at 1·9 Mc/s, and tune the p.a. for minimum current. Connect a 12 volt, 12 watt lamp as a load and adjust the LOADING knob to give maximum brightness. The modulation may be checked through the output on a receiver or, if possible, on an oscilloscope. Check the operation of the modulation monitor.

V.F.O. CONTROLLED VALVE TRANSMITTER FOR 1·8-2 MC/S

This 10 watt transmitter has a built-in v.f.o. covering 1·8–2 Mc/s and provision for high level anode and screen modulation. C.w. operation is also possible.

Although an h.t. supply of 250 volts is suggested, the transmitter may be used on any voltage between 250 and 300 provided the screen dropping resistor is correctly chosen. The measured efficiency of the p.a. is 70 per cent.

The circuit of the transmitter is shown in **Fig. 6.90.** V1 (EF91/6AM6) functions as a Tesla-Vackar type v.f.o. (see Fig. 6.25) and is followed by a class A buffer stage V2. The grid circuit of this stage is purely resistive, R4 being provided to permit the connection of an external meter to check the grid current. The anode circuit of V2 is tuned to the operating frequency by L2 C13. If this circuit is resonated at 1·9 Mc/s it will be found that drive to the p.a. will remain substantially constant between 1·8 and 2 Mc/s. C13 may therefore be a pre-set trimmer. R7 provides a test point for checking the anode current during the alignment procedure.

The p.a. V3 (6BW6) is conventional. Drive to the grid is via C14 and to improve the efficiency a 2·5 mH r.f. choke is connected in series with the grid resistor R9. The anode

TABLE 6.10

Inductor details for Fig. 6.90

L1	70 turns 34 s.w.g. enam. close wound on ¼ in. former fitted with ferrite core.
L2	90 turns 34 s.w.g. enam. close wound on ½ in. former.
L3	26 s.w.g. enam. wire close wound for a distance of 1¼ in. on a 1½ in. diam. former. Over one end bind two layers of p.v.c. tape. This is the " cold " end of the p.a. coil, the anode being connected to the other end.
L3a	Starting at the end of L3, wind 8 turns 26 s.w.g. over the p.v.c. band on L3.
T1	Centre tapped output transformer. Primary 8 K ohms to 10 K ohms centre tapped, rated at 50 mA d.c. Secondary not used.

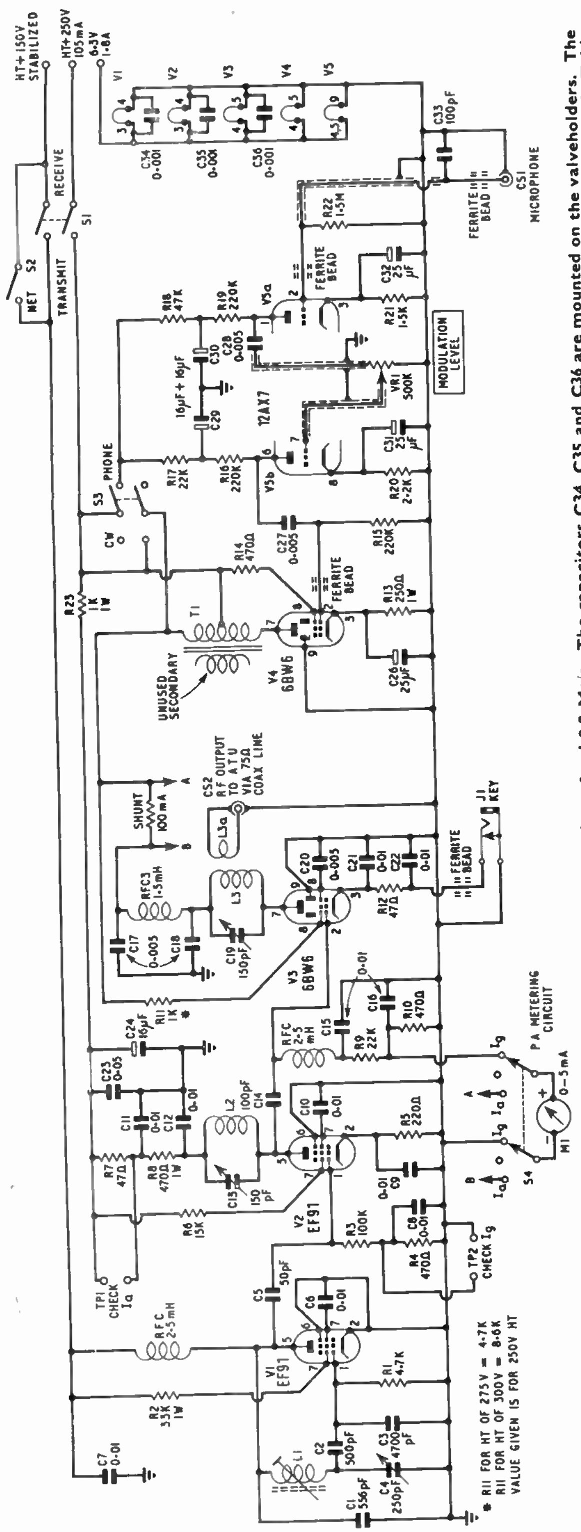


Fig. 6.90. Circuit diagram of the v.f.o. controlled 10 watt phone/c.w. transmitter for 1·8–2 Mc/s. The capacitors C34, C35 and C36 are mounted on the valveholders. The value of the screen resistor R11 for the p.a. valve depends on the h.t. supply. J1 must be a closed circuit jack. Details of the coils and the transformer T1 are given in Table 6.10.

tank is a simple parallel tuned circuit C19 L3, the output of which is taken by a link, L3a, proportioned to match 75 ohm co-axial cable. A 1·5 mH r.f. choke is used in the anode circuit to prevent r.f. feeding into the metering circuit or into the modulator.

M1, a 0–5 mA meter, switched by S4, is used to measure grid and anode currents. To avoid damage to the meter when switching, there is an unused position between the two current readings. When switched to read anode current, the meter is connected across a shunt permanently wired into the anode circuit which modifies the range to read 100 mA full scale. The leads to S4 should be screened to reduce the possibility of stray coupling.

The modulator valve V4 is also a 6BW6 and is coupled to the p.a. by T1, a standard centre tapped output transformer which is used as a 1 : 1 modulation transformer. As the impedance of the p.a. is almost equal to the recommended load for the modulator valve, a good match is obtained. T1 may be relatively small provided that the winding can carry the currents drawn by the p.a. and modulator valves.

V4 is driven by a two-stage speech amplifier, with the gain control between V5a and V5b. Ferrite beads are fitted to grids of V4 and V5 to prevent any stray r.f. picked up on the modulator leads being fed to these stages.

The transmitter is controlled by three switches: the main send/receive switch S1 controls power to the v.f.o., r.f. and modulator sections, S2 allows the v.f.o. to be switched on for netting and switching from a.m. to c.w. is provided by S3. In the c.w. position the modulator is disabled and a short circuit is placed across the modulation transformer to protect it from d.c. surges.

Many modern receivers incorporate a send-receive switch intended to control both receiver and transmitter. A suitable control circuit for use with such a switch is shown in **Fig. 6.91**.

Construction

The transmitter should be laid out in a straight line with vertical screens between each section and the coils in each section arranged so that they are at right angles to those in the adjacent sections. A screen is placed around the under-chassis r.f. circuits so that, with the bottom plate in place, they are effectively within a screened box. The layout is shown in **Fig. 6.92**.

The v.f.o. should be constructed as a separate unit. Most of the p.a. components are on top of the chassis, C19, L3 and RFC3 being contained in a screened compartment to the rear of which is fitted the co-axial output socket CS2.

To avoid earth loops which might give rise to hum or instability, the screened cable which runs from the microphone socket CS1 to the grid of V5a should have its outer screening earthed only at the socket, the screening at the grid end being left disconnected. The grid resistor R22 should be mounted directly on the grid pin of V5a and earthed by the shortest route.

Alignment

After the v.f.o. has been constructed, it should be tested as a separate unit. When it has been fitted to the chassis and the wiring checked, V2, V3 and V4 are fitted, S3 set to the

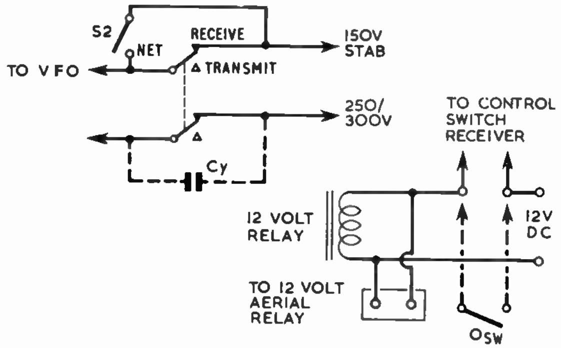


Fig. 6.91. Substituting a relay for S1 in Fig. 6.90. If there is arcing across the main contacts of the relay, a 0·001–0·01μF capacitor should be fitted at Cy. The smallest satisfactory value should be used. A manual override switch SW may be fitted to permit the transmitter to be switched on independently of the receiver.

c.w. position and S4 to read p.a. grid current. H.t. to the p.a. is disconnected by removing the connection from T1. Power is then applied by closing S1. C13 is tuned to produce maximum grid current (approximately 3 mA) and the frequency 1·8–2 Mc/s) checked with an absorption wavemeter.

With the drive to V4 still applied, C19 should be tuned through its range to see whether the p.a. grid current varies. If it does, the stage must be neutralized or the decoupling and screening improved. The power supply may now be switched off and the h.t. reconnected to the p.a. When this has been done, S4 should be set to read the anode current to V4. C19 should be at maximum capacity. The power supply may now be switched on again and C19 quickly tuned for maximum dip in anode current. Next open S1. Lightly couple an absorption wavemeter to L3, close S1 and confirm that the frequency is in the 1·8–2 Mc/s range. Open S1 again. A 75 ohm dummy load should now be connected to the r.f. output socket CS2, S1 closed and C19 tuned for maximum dip.

To avoid exceeding a d.c. input of 10 watts, the anode current must not exceed the following values under loaded conditions:

at 250 volts h.t. the current should be between 35–40 mA,
at 275 volts between 32–36 mA and
at 300 volts between 29–33 mA.

If the anode current exceeds these values, L3a should be moved away from L3 to reduce the coupling.

V5 may now be fitted and a high impedance microphone connected to CS1. VR1 should be at minimum gain. S1 may then be closed and the gain control advanced about half way. Speaking into the microphone should produce a marked increase in the brilliance of a flashlamp bulb loosely coupled to L3 by a three turn coil. The microphone should be kept away from the p.a. tank circuit to avoid r.f. pickup which might cause feedback.

The transmitter is now ready for connection to a suitable aerial through an aerial tuning unit.

A TT21 (7623) TRANSMITTER FOR 14, 21 AND 28 MC/S

The transmitter illustrated below employs a TT21 in the p.a. stage and is capable of 60 watts peak input on a.m. and 80 watts input when using narrow band frequency modulation or c.w. A.m. is provided by series gate screen modulation of the p.a.

In the interests of TVI prevention, the r.f. circuits of the transmitter are completely screened without reliance on a

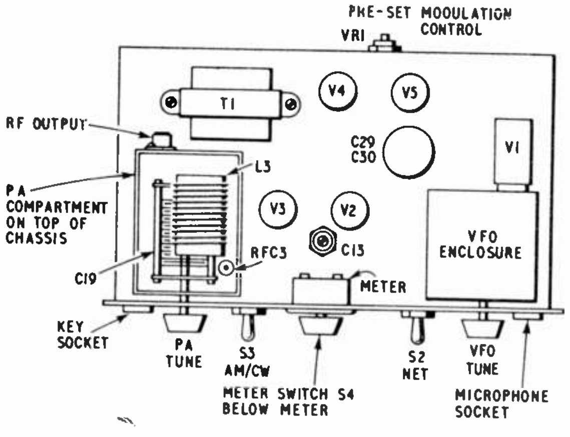


Fig. 6.92. Layout of the top of the chassis of the v.f.o. controlled 10 watt transmitter.

The TT21 transmitter for the 14, 21 and 28 Mc/s bands.

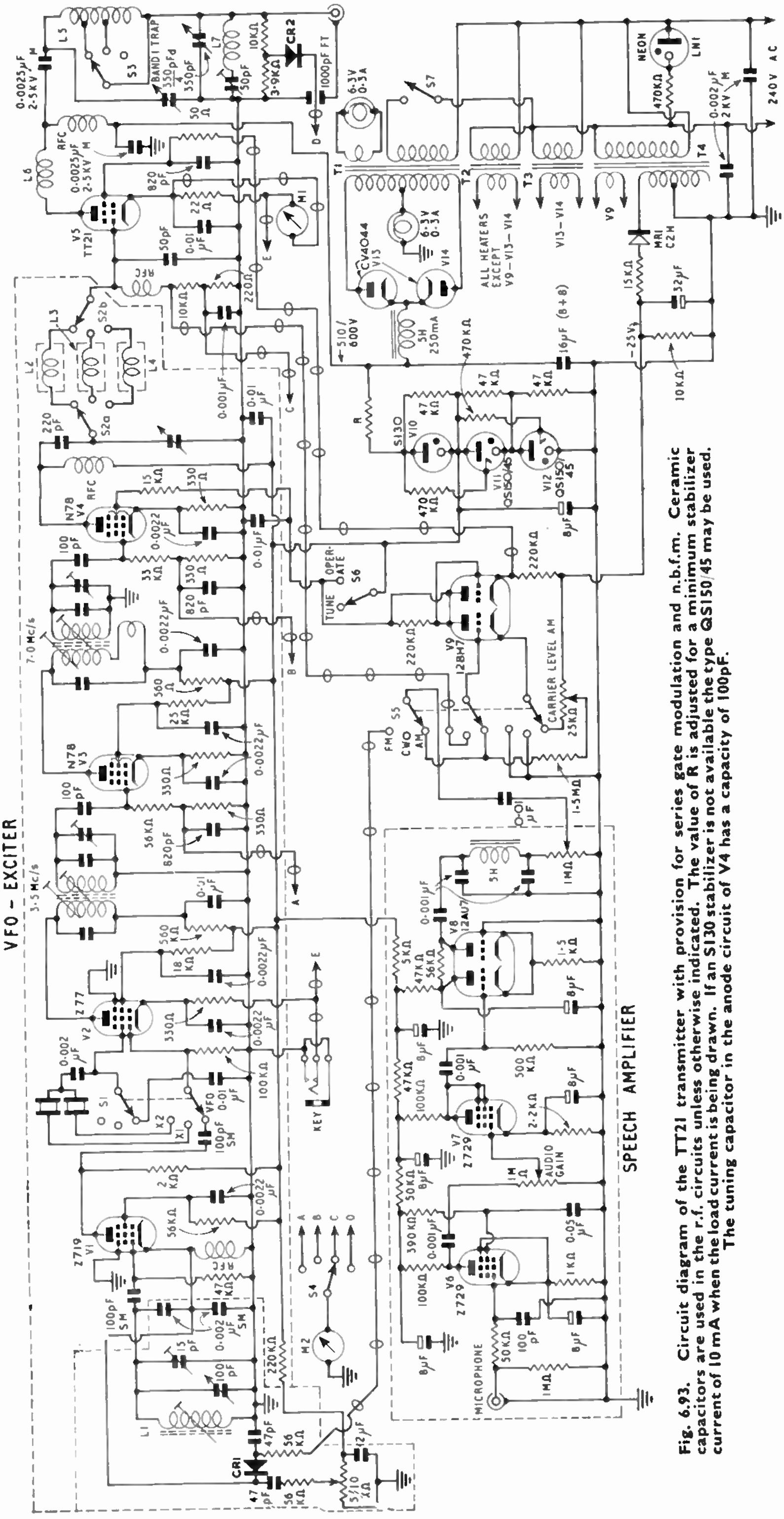


Fig. 6.93. Circuit diagram of the TT21 transmitter with provision for series gate modulation and n.b.f.m. Ceramic capacitors are used in the r.f. circuits unless otherwise indicated. The value of R is adjusted for a minimum stabilizer current of 10 mA when the load current is being drawn. If an S130 stabilizer is not available the type QS150/45 may be used. The tuning capacitor in the anode circuit of V4 has a capacity of 100pF.

cabinet so that the unit can be used either as a table top transmitter or installed in a rack with other equipment without any significant increase in unwanted radiation. The more complicated units are completely removable from the main chassis so that construction and servicing are simplified.

The circuit of the complete transmitter, including the built-in power supply, is shown in **Fig. 6.93**.

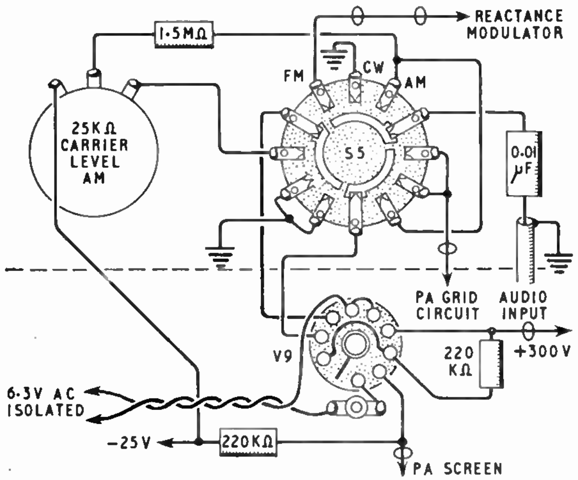
The v.f.o. employs a screened r.f. pentode V1 (EF80/Z719/6BX6) operating as a high *C* Colpitts electron coupled oscillator covering 3·5–3·65 Mc/s. Provision for frequency modulation of the oscillator is made by use of a silicon voltage variable diode CR1 across part of the tuned circuit, a potentiometer providing the required bias so that a suitably linear portion of the diode operating curve can be selected. In this case the bias is set to +1·0 to +1·25 volt which will give approximately equal capacity change for both sidebands.

The second stage, V2 (Z77/EF91/6AM6), acts either as a buffer stage or as a crystal oscillator, selected by S1. The anode circuit of this stage is tuned to 3·5 Mc/s and coupled to the next stage by a wideband coupler.

V3 (N78) is operated as a straightforward frequency doubler and feeds the next stage via a 7 Mc/s wideband coupler. The final stage of the exciter V4 is operated as a doubler from 7 to 14 Mc/s, as a trebler from 7 to 21 Mc/s and as a quadrupler from 7 to 28 Mc/s. The valve chosen (a type N78) gives somewhat better performance than the more usual 5763, due to its higher slope and lower drive requirement. Coupling between this stage and the p.a. is by means of a pi-network which provides harmonic attenuation while keeping the impedance of the grid circuit of the power amplifier relatively low, so helping to maintain the stability of the p.a. valve. The circuit is adjusted by tuning for maximum grid current to V5. When netting to a frequency, h.t. to the screens of this stage and of the power amplifier is switched off by S6. For c.w. operation, the cathodes of V2 and V5 are keyed.

The whole of the exciter is built as a unit so that it can be removed from the main chassis by removing four fixing screws and unsoldering the interconnections to the main chassis.

The final stage V5 uses a type TT21 (7623) tetrode, which is operated at 60 watts peak input on a.m. and at 80 watts input on c.w. and n.b.f.m. The anode circuit is a pi-network with the inductance tapped at suitable points for three bands by S3. Across the output are connected a TV



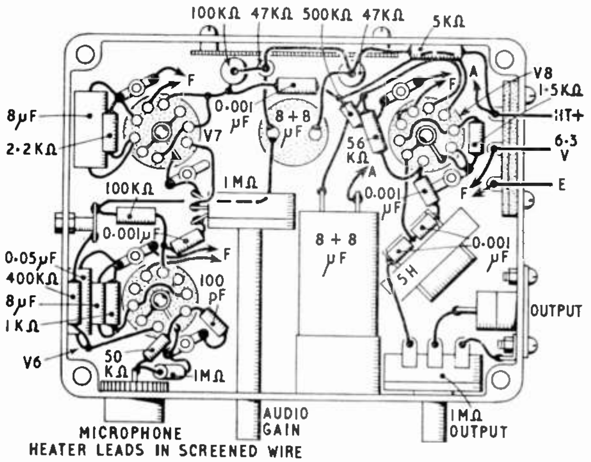


Fig. 6.94. Layout of the speech amplifier and clipper unit for the TT21 transmitter.

Band 1 series tuned trap circuit and an r.f. voltmeter which is a useful means of correctly tuning the output circuit, the power output being arbitrarily indicated on the meter on the righthand side of the front panel. By means of the switch S4, this meter also reads the grid current to V3, V4 and V5. The current input to the p.a. is indicated on M1 connected in the cathode circuit of the TT21. The valve is normally operated with 600 volts h.t. on its anode.

The whole of the power amplifier is built at the lefthand end of the main chassis and is completely screened. The grid circuit for the valve is below the chassis and also screened.

The speech amplifier comprises V6 (Z729/EF86/6267/6F22) as a microphone amplifier with a voltage gain of about 120 followed by a second stage V7 (also a EF86 but triode connected) having a voltage gain of about 25. V7 feeds the double triode clipper V8 (ECC82/12AU7) in the anode circuit of which a simple low pass a.f. filter is fitted. Suitable adjustments of the two gain controls enable a considerable amount of clipping to be applied which is desirable when using n.b.f.m.

The speech amplifier/clipper is built into an Eddystone 4½ in. × 3¾ in. × 2 in. die-cast box **(Fig. 6.94)** which is attached to the top of the main chassis at the front right hand corner. The output from this unit is switched by S5 either to the series gate screen modulator valve, V9 (12BH7), or to the reactance modulator CR1 (EW76 or SX761). When the switch is in either the c.w. or f.m. positions V9 operates as a clamp valve for the p.a. stage.

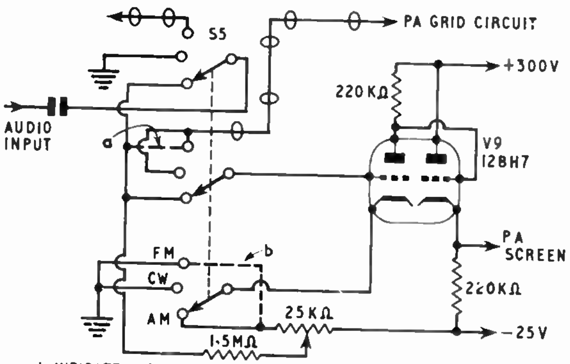
An alternative series gate clamp circuit to provide for adjustable carrier level when using n.b.f.m. is shown in **Fig. 6.95.**

The power supply is conventional and is provided by a 650 volt h.t. transformer and separate heater transformers. The main rectifiers, V13 and V14, are indirectly heated type CV4044/6443 half-wave valves, the pair providing the total current of about 225 mA. Smoothing of the supply is by a single 5 Henry choke and a 16 μF paper capacitor (two 8 μF units in parallel). The p.a. is fed directly but all other stages receive h.t. from the main potentiometer made up of a fixed resistance in series with three gas stabilizer tubes V10 (S130) and V11 and V12 (both type QS150/45). The striking electrodes are connected to suitable points to provide quick ignition.

To avoid breakdown between the heater and cathode of the screen modulator valve V9, a separate transformer is used. The heater must not be earthed. A small bias supply using a 0–125 volt transformer (T4) is provided for the series gate modulator.

Construction

The complete transmitter is built on a 17 in. × 13 in. × 3 in. chassis to which a standard 19 in. × 8¾ in. front panel



a – b INDICATE ALTERNATIVE CONNECTIONS FOR ADJUSTABLE CARRIER LEVEL WHEN USING FM

Fig. 6.95. Alternative series gate clamp circuit providing adjustable carrier level when using narrow band frequency modulation.

is rigidly fixed by panel brackets. The layout is shown in **Fig. 6.96.**

After the holes for fixing the chassis to the front panel have been drilled the hole for the v.f.o.-exciter unit should be made. This is probably most readily done with a tension file. The mounting of the p.a. screening box and the principal components on the main chassis is quite straightforward and should not present any difficulty. Some care is necessary in placing the fixing holes for the p.a. so that they are not fouled by operating shafts.

The v.f.o. and exciter unit is built on to a 10 in. × 5¼ in. plate of ⅛ in. thick aluminium. A screening box is fitted on the underside of this plate whilst the v.f.o. tuned circuit is built into an Eddystone die-cast box measuring 4½ in. ×

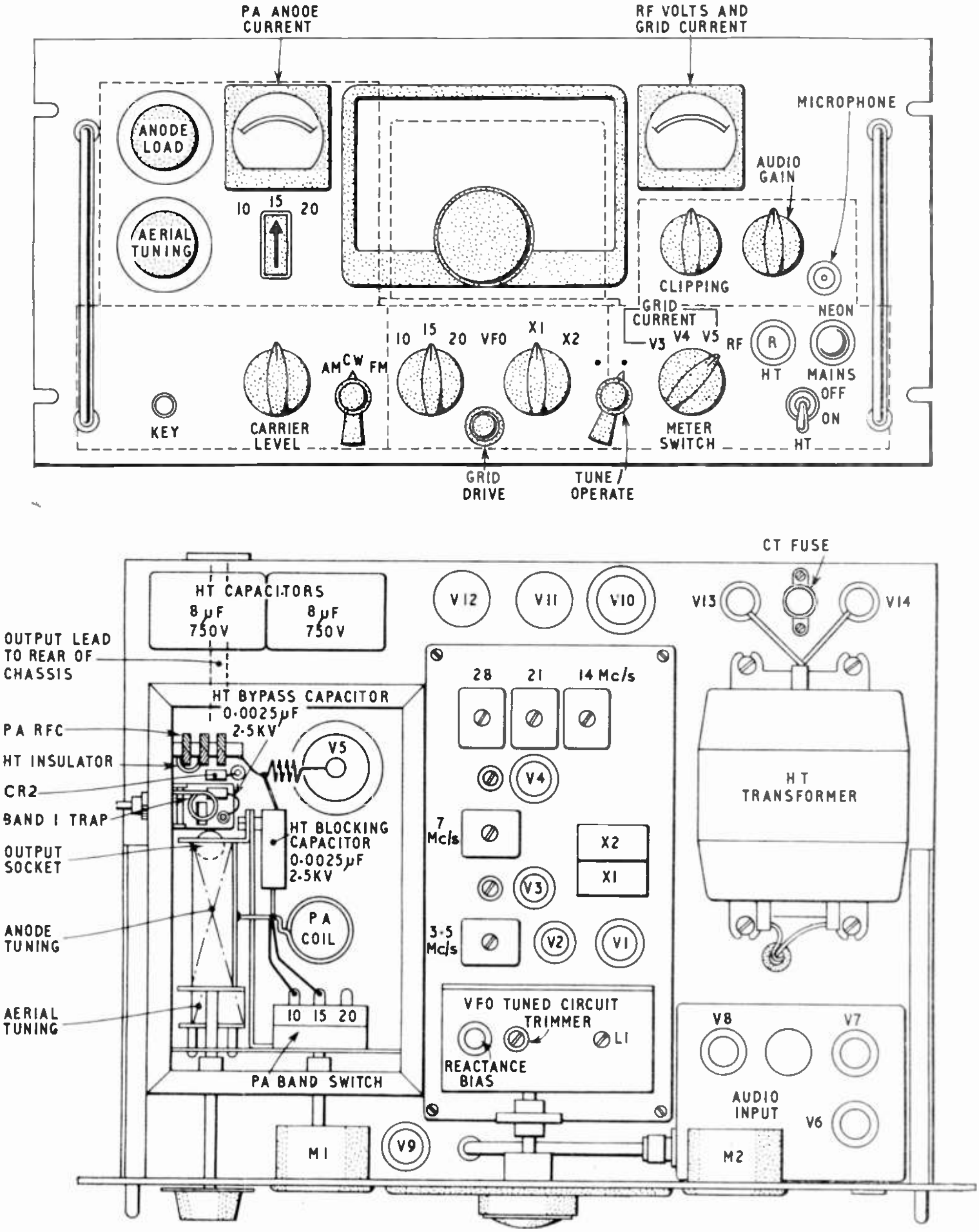


Fig. 6.96. Layout of the front panel and chassis.

3⅜ in. × 2 in. and attached to the top near the front end by its long side with the removable plate at the rear.

The speech amplifier/clipper is built into a similar die-cast box and mounted upside down so that the valves and other components are on the bottom of the box which is attached to the main chassis by its normal four fixing screws. The lid is not used. Considerable care is necessary to position the microphone socket (a Belling and Lee type L722/S) so that it is

(a) as close as possible to the corner of the box but with sufficient clearance for attachment of the back nut,
(b) does not foul the handle on the front panel.

Both of these units are electrically connected to the rest of the transmitter by soldered connections on the distribution terminal strip in the main chassis and are readily removable for servicing and testing.

Details of the coils are given in Table **6.11**.

The primary and secondary windings of the 3·5 and 7 Mc/s wideband couplers are wound in the same sense, the outer ends of the coils being connected to anode (top) and grid (bottom) with the earthy ends of the two coils at the centre. Coupling between the two windings is as follows:

3·5 Mc/s Top coupled through a 10 pF capacitor.

7 Mc/s Extra turns of the primary are wound over the earthy end of the secondary winding. This method was found to give more output than top capacitor coupling.

So that both the anode and grid circuits of these units can be adjusted from the top of the chassis a small trimmer is connected across each of the secondaries in addition to the core. The anode coil has a dust iron slug.

The driver anode coils are wound on the same type of former as those used for the transformer couplings. The 14 and 21 Mc/s coils are wound on the top half of the former. The inside diameter of the 28 Mc/s coil is about 7/16 in. All the coils and couplers are individually screened in cans.

TABLE 6.11

Inductor Details for the TT21 Transmitter

L1 (*v.f.o.*) 12½ turns 16 s.w.g. enam. close wound on ⅜ in. diam. ceramic former with dust iron core. Tuned by 100 pF and 15 pF trimmer.
L2 (*14 Mc/s*), 20 turns 20 s.w.g. enam. close wound on ⅜ in. diam. former with dust iron core (wound at top of i.f.t. former from type 373 unit).
L3 (*21 Mc/s*), 15 turns 18 s.w.g. enam. close wound on ⅜ in diam. former with dust iron core (as for L2).
L4 (*28 Mc/s*), 14 turns 16 s.w.g. close wound ⅜ in. diam. (self supporting inside i.f.t. can from type 373 unit).
L5 (*p.a.*), 11 turns 14 s.w.g. enam. spaced two wire diameters 1¼ in. diam. self supporting, tapped from anode end at 8 turns for 21 Mc/s and at 4 turns for 28 Mc/s.
L6 (*parasitic suppressor*), 4 turns 16 s.w.g. enam. spaced one wire diameter, 3/16 in. inside diam. self supporting.
L7 (*Band I trap*), 7 turns 18 s.w.g. spaced one wire diameter ½ in. inside diam. self supporting and tuned by 50 pF.

Wideband Couplers

3·5 Mc/s—primary and secondary both 45 turns 30 s.w.g. enam. close wound on ½ in. diam. former (i.f.t. former from type 373 unit), both tuned by 100 pF, dust iron core in primary and brass core in secondary, coupling 10 pF.
7 Mc/s—primary and secondary both 34 turns 26 s.w.g. enam. close wound on ½ in. diam. former (i.f.t. former from type 373 unit), primary tuned by 22 pF, secondary tuned by 30 pF, dust iron core in primary, brass core in secondary, coupling 3 turns turns extra on primary wound over earthy end of secondary (see text).

TABLE 6.12

Mode	Audio Gain Control	Clipper Control	Reactance Diode Bias	P.a. Anode Current
A.m. (Series Gate)	Half to maximum	Maximum	—	40–50 mA (no sig.) 110–120 mA (max. sig.)
N.b.f.m.*	Half to maximum	Quarter to half	+ 1 to + 1·25 volts	110–120 mA
C.w.	—	—	—	150–160 mA

* When the transmitter is wired so that S5 permits the carrier level to be adjusted on both series gate a.m. and on n.b.f.m.

Adjustment and Operation

After completion of the constructional work, the two smaller units should be adjusted before being fixed on to the main chassis. It is desirable that an alternative power supply giving about 300 volts at 60 mA and heater voltage should be employed while adjusting the v.f.o. exciter unit rather than the much more dangerous 650 volt h.t. source.

First, the v.f.o. should be set up and its tuning range adjusted, preferably with the help of a receiver switched to the 10m band. With the components specified the v.f.o. covers only 1·2 Mc/s (28–29·2 Mc/s) of this band as it was felt that the more open tuning for 15m and 20m was worth the sacrifice on the higher frequency band. If preferred a larger tuning capacitor could be used so that the whole band is covered but this might result in some loss of drive at the band edges.

The stability of the v.f.o. is reasonably good. Provided, however, that the main tuning capacitor and the fixed capacitors are of good quality it should be satisfactory. Having adjusted the v.f.o. range, S1 should be set to select one of the crystals and the operation of the crystal oscillator checked. Next, with V3 in position and using its grid current at point *A* on Fig. **6.93** as an indication on M2, adjust the tuning of the primary and secondary of the 3·5 Mc/s bandpass coupler until a satisfactory frequency response is obtained. The procedure should be repeated with V4 in position, using the grid current indicated at point *B*. Check that there is no instability in these stages before proceeding to adjust the anode circuits of V4. The total anode and screen input to the first three stages should be between 40 and 55 mA at 300 volts.

If, as recommended, this preliminary tuning is carried out with an external power supply, some form of temporary r.f. voltmeter will be required to check the output from V4.

When adjustment of the exciter stages is complete, the unit may be connected to the main power supply and the complete transmitter tested.

The p.a. anode circuit is adjusted in the usual manner for a pi-network: tune for maximum dip as indicated on M1 and then increase the loading until V5 is drawing 120 mA. Final adjustment should be made by adjusting the p.a. tuning controls for maximum output as indicated on M2 when switched to position *D*.

Satisfactory speech quality and modulation depth or deviation is obtained with the settings of the audio controls shown in Table **6.12**.

The Princess transmitter employing a mixer master oscillator and linear p.a. stage for c.w. and a.m. operation.

THE PRINCESS FIVE BAND TRANSMITTER

The Princess transmitter, is unusual in that the v.f.o. is a mixer master oscillator and the p.a. stage operates in class AB1 in contrast to the more usual class C. As a further aid to the reduction of harmonic output, a pi-network is used to couple the driver stage to the p.a.

The effective input to the linear amplifier on a.m. is 75 watts but as this represents a loss of only half an S point (3db) of signal strength compared with a 150 watt transmitter it is of little importance bearing in mind the economies it permits in the power supplies and modulator. In fact, a pair of small tetrode valves (type EL85) anode and screen modulate the driver valve, the exciter section thus comprising a 15 watt transmitter which feeds a low impedance load across which sufficient r.f. voltage is developed to drive the p.a. With such a low value of resistor in the p.a. grid circuit instability is most unlikely and neutralization of the TT21 (7623) valves is not required.

The circuit of the exciter stages is given in **Fig. 6.97**.

The master oscillator comprises V2, V3 and V4. V2a is a modified Colpitts type oscillator between grid and cathode the frequency coverage of which is from 1·5 Mc/s to 2·0 Mc/s, across about 150° of the traverse of C13, the v.f.o. tuning control.

Temperature compensation is effected by C14, which has a negative temperature coefficient of 750 parts per million. This capacitor, in series with the air-spaced trimmer, C15, gives close control of the overall thermal stability. Final adjustment of this trimmer should not be attempted during the first four weeks of use to allow component ageing. The transmitter, in its case, should be checked for drift against a high-stability receiver which has been switched on for several hours, or against a frequency meter or crystal oscillator. If the v.f.o. appears to drift, C15 should be altered slightly and the process repeated. Short-term drift is excellent even without careful setting of C15, but an improvement in the long-term stability results from patience used in its adjustment.

The series circuit L6, C24 is resonated at approximately 3·6 Mc/s. This is necessary because the oscillator second harmonic from 1800 kc/s is within the pass-band of the following stages and would cause some interference.

V2b functions as a fixed frequency crystal oscillator at 5.50 Mc/s. Output taken from the anode is fed to V3.

From the mixer valve, V3, the v.f.o. and crystal oscillator signals and their sum and difference products appear at S3a. For 80m operation, the difference product is selected by the broad-band transformer T4. The v.f.o., tuning from right to left of the dial, supplies 2·0 Mc/s to 1·5 Mc/s mixing with 5·5 Mc/s to give frequencies of 3·5 Mc/s to 4·0 Mc/s. The original v.f.o. and crystal oscillator products, as also the sum, are far enough from the passband of T4 and subsequent circuits to be rejected.

For 40m operation, the sum product is selected by T5. The v.f.o., tuning from left to right, gives 1·5 Mc/s to 2·0 Mc/s which when mixed with 5·5 Mc/s, provides output frequencies in the range 7·0 Mc/s to 7·25 Mc/s and beyond. For higher frequency bands this output is multiplied.

The design of the broadband transformers T4 to T10 is as simple as it is possible to make it. Two identical windings on the same former, with a given spacing between the end of one winding and the start of the second give a controlled degree of inductive coupling. Because the short 9mm iron dust cores give only a small variation in inductance, and as stray circuit capacitances will vary from model to model, it may be found necessary to alter slightly the values of parallel tuning capacitors; in certain cases these may be omitted altogether.

The output winding of either T4 or T5 feeds the amplifying valve V4 according to the band in use. An r.f. voltage test point (R29 and MR18) permits the alignment of the mixer anode circuit by connecting a 10,000 ohms per volt test-meter on a low voltage range to TP1. Similar broadband transformers for 80m and 40m are used in the anode circuit of V4, the output windings driving V7 directly on 80m or 40m. However, for operation on 20m, 15m or 10m, the 40m output at S3d is passed to the first doubler/tripler, V5.

Keying is by the grid-block method. Under key-up conditions, the bias cuts off V4. When the key is depressed, the grid circuit of V4 is earthed but due to the high value of R27 in the power supply, the bias to the driver is unaffected. R73 is necessary to make C39, the decoupling capacitor, effective. The DRIVE control, VR3, varies the screen potential on V4 up to a maximum value set by R32.

The value of R28 in the anode circuit of V3 is selected to provide the most efficient mixing action. At an anode potential of 100 volts, signal levels to the grids, quoted in r.m.s. values, will be of the order of 2 volts from the v.f.o.

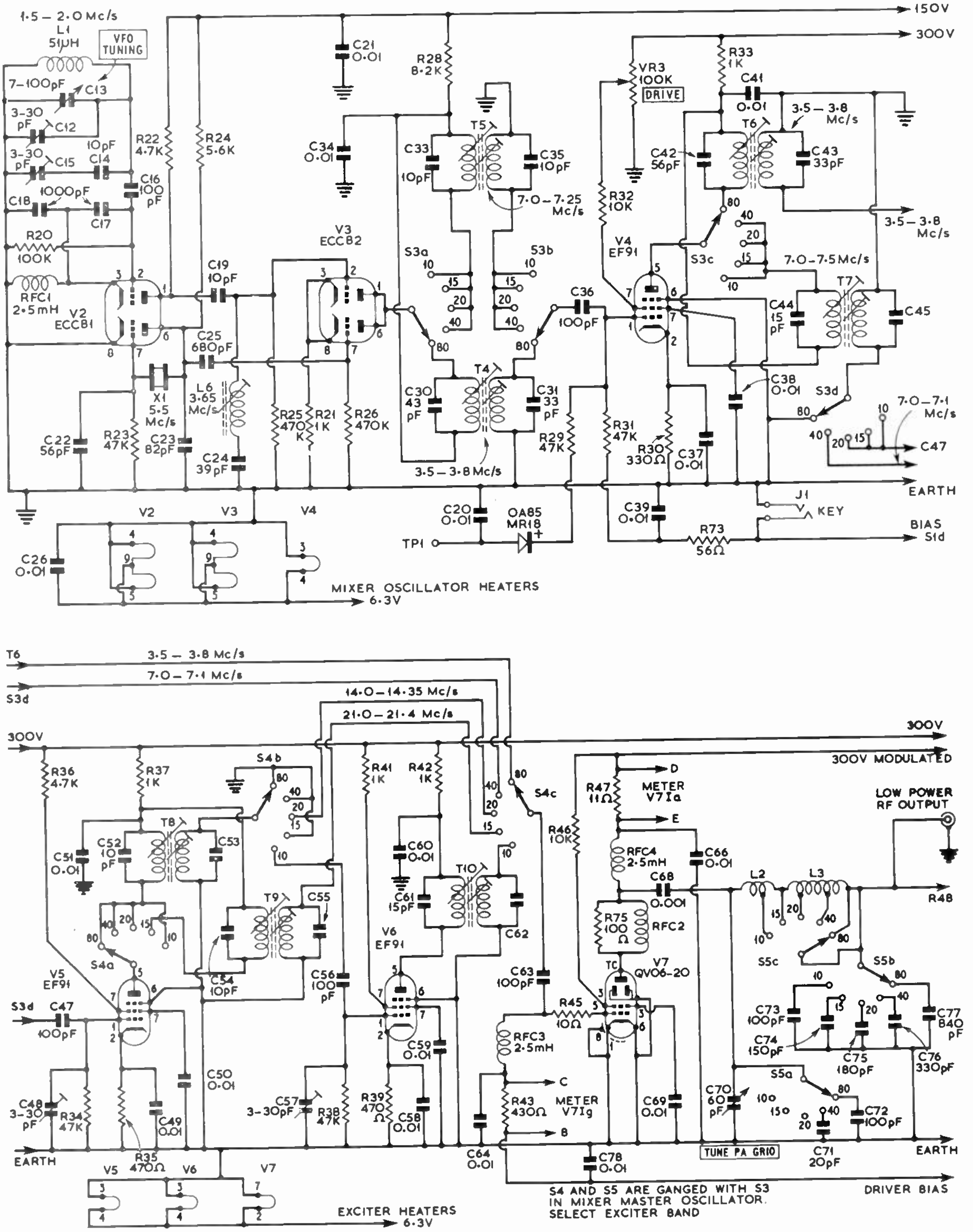
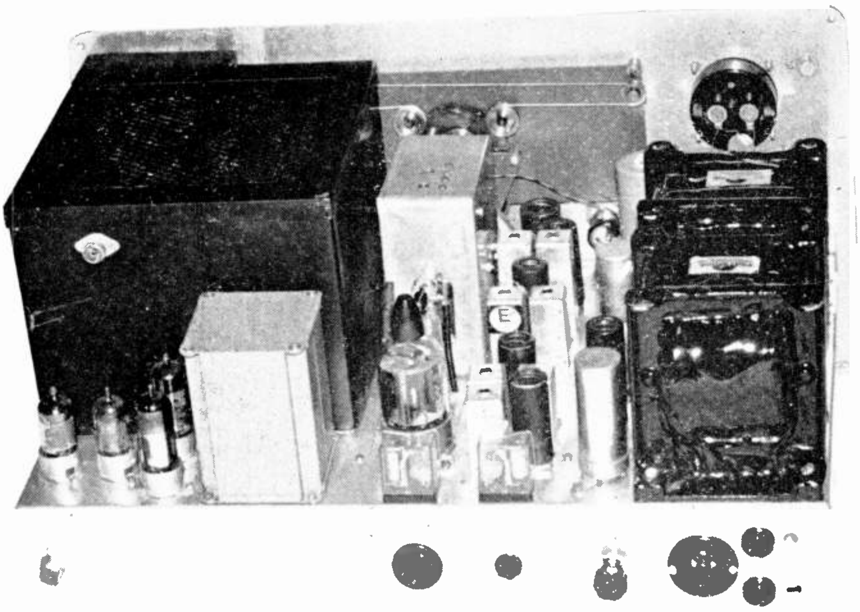


Fig. 6.97. Circuits of the mixer master oscillator, above, and the frequency multipliers and driver stage. The values of C55 and C62 are 10pF. The linear p.a. stage is shown in Fig. 6.98.

and up to 6 volts from the c.o. Wide variations may be due to the particular specimen of valve in use as V2.

Type EF91/6AM6 are specified for V5 and V6 positions because they are cheap and economical to run. It will probably be necessary to select valves in these positions to obtain the required drive.

V5 is used as a doubler to 20m or as a tripler to 15m. Broadband transformers T8 and T9 cover 14·0 Mc/s to 14·5 Mc/s and 21·0 Mc/s to 21·45 Mc/s respectively. On 20m and 15m, V7 is driven directly. For 10m operation, T8 is switched to V6, which doubles the 20m signal. C48 and C57 are tuned on a signal to equalize circuit capacitances when switching between a following doubler or the driver valve. Care must be exercised on 20, 15 and 10m that the valves are not overdriven, resulting in a reduction of drive at the wanted frequency. VR3 should be set to the minimum for the band in use consistent with the required drive level.



Rear view of the Princess transmitter. Note the shaping of the p.a. screen to clear the v.f.o. dial mechanism.

The driver, V7 (QV06/20 or 6146), has fixed grid bias and a parallel fed pi-network anode tank circuit. Grid current is monitored across R43 and anode current across R47. Anti-parasitic components R45 and R75 RFC2 are mounted as close to the valve connections as possible. The tank circuit is loaded by the passive grid resistor, R48, of the following linear amplifier (Fig. 6.98). A secondary r.f. output socket is provided, and with the feed to R48 disconnected, V7 becomes a stage of about 15 watts input, for low power applications.

With V7 as a driver, C70 is the p.a. grid tuning control and the loading is pre-set to give sufficient signal across R48 to drive the amplifier up to 150 watts d.c. input. Increasing the values quoted for C73 to C77 decreases the loading and so the drive to the final is reduced.

Modulation is applied to the anode and screen of the driver valve. For reasons explained later, it is necessary to reduce carrier power on A3 to 75 watts input, effected by switching R76 in series with the h.t. to V7. Full modulation capability of 100 per cent is available.

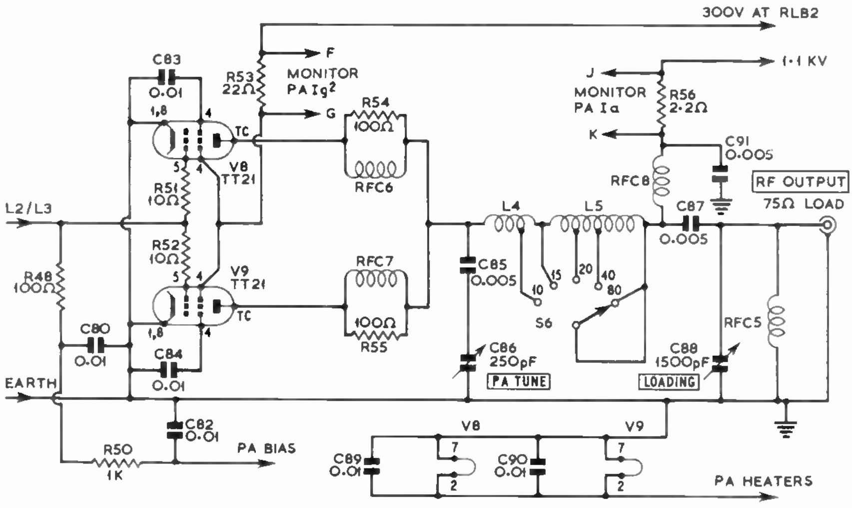


Fig. 6.98. The linear power amplifier stage for the Princess transmitter.

The linear amplifier Fig. 6.98 operates in class AB1 with passive grid input, as opposed to tuned grid. As grid current does not flow during any part of the input cycle, no driving power is required but only an r.f. voltage swing. The standing bias potential is a nominal −45 volts, which is adjusted to give a quiescent anode current of 70 mA.

So long as the peak value of the positive half-cycle of the driving waveform does not exceed this standing bias, the grids of V8 and V9 cannot go positive and draw grid current. The maximum permissible signal is therefore 32 volts r.m.s. This would correspond, with the present arrangement, to 300

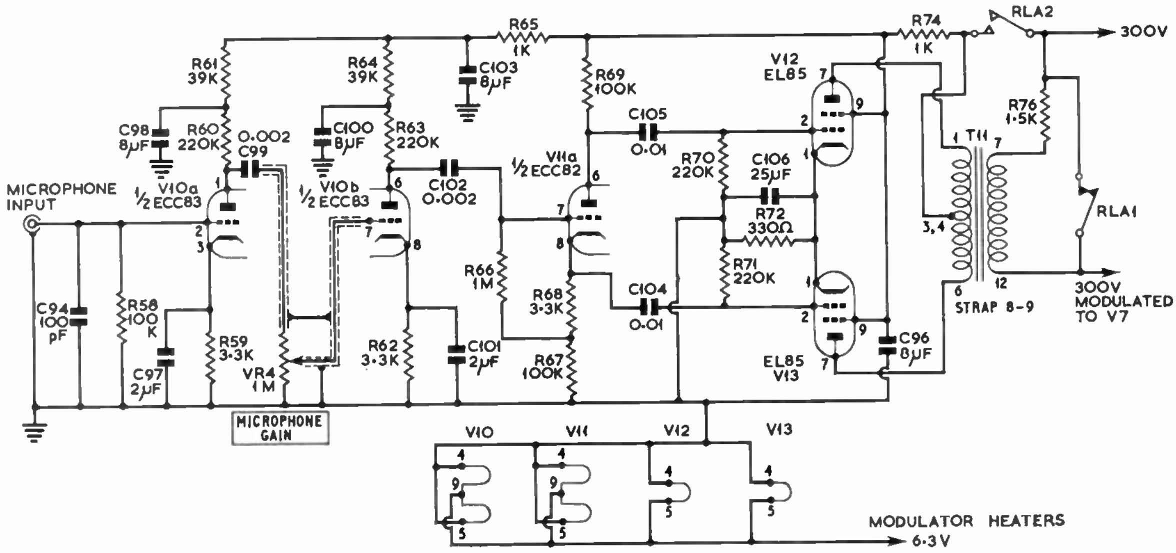


Fig. 6.99. The speech amplifier and modulator stages. V10, ECC83/12AX7, V11, ECC82/12AU7, V12, 13, EL85.

watts input. As only 150 watts is required, the drive will be 3db less, 22·5 volts r.m.s. Across 100 ohms, this corresponds to a c.w. drive power of about 5 watts. Allowing for circuit losses, at the lowest frequencies the fixed loading capacitors are selected for 10 watts input rising to 15 watts at 10m.

On phone use is made of the full capability of 300 watts peak envelope power. At 100 per cent modulation, this permits the use of a 75 watt carrier so R76 (**Fig. 6.99**) reduces the drive signal required. As a point of interest, a 150 watt transmitter at 100 per cent modulation runs up to 600 watts p.e.p. input.

A pi-network tank circuit is used, with the h.t. feed at the low impedance end of the coil. This avoids the necessity for a critical component at RFC8, as would be the case if d.c. was applied through the choke direct to the anode. Blocking capacitors C85 and C87 keep the d.c. from the tuning controls and r.f. output socket. With this system RFC5 is essential, because the feeder line would become over 1000 volts above earth if C87 failed.

Both anode and screen currents are metered and with an amplifier running in class AB1, screen indications are of great importance. The p.a. tuning control is adjusted for a peak in screen current and the loading control set so that this peak is about 25 mA on c.w. The valves are then running at the manufacturers' recommended figures for class AB1 operation. This method of tuning and loading is more precise than relying only on indications of anode current.

High stability carbon components are used as meter shunts, the values of which may be obtained, where necessary, by series/parallel arrangements. The resistor R57 is of such a value as to increase the internal resistance of the meter to 1100 ohms. By this means, as long as the internal value is known, most 1 mA meters will be suitable. (**Fig. 6.100**).

With a maximum input to the driver valve V7 (Fig. 6.97) of 8 watts on phone, it follows that 6 watts of audio will be sufficient. This can be obtained from a pair of EL84/6BQ5 valves, but the standing anode current for the two is about 80 mA and the full power capability is at no time required. A slightly smaller version, the EL85, is therefore used, with a quiescent current of 50 mA. The circuit of the modulator is given in Fig. 6.99. For differing types of valve the value of R72 must be varied. The Woden UM0 multi-ratio modulation transformer (T11) permits an accurate match in all cases.

V10 is a two-stage resistance-capacitance coupled speech amplifier with the values of C99 and C102 selected to attenuate the lower speech frequencies. Half of V11 is used as a phase splitter, the second half being unused. Relay contact RLA1 shorts out the secondary of the modulation transformer when not using phone. If this precaution is not taken, voltage peaks of sufficient amplitude to break down the transformer may result.

Control Sequences

The control and metering circuits are shown in Fig. 6.100. The system switch, S1, has five positions, the first of which is OFF. At EXCITER ON, all stages up to the linear amplifier are live. The transmitter cannot be switched to TRANSMIT at this stage, nor can the modulator be used. The NET position differs only that the key is short-circuited.

When the fourth position of S1, c.w. (A1), is selected, relay RLC in the power supply operates, putting high voltage on the linear V8-V9. No anode current flows as the screens are held at −12 volts. S2 may now be used to control the station. At TRANSMIT, relay RLB is energized, the p.a. screens are switched to 300 volts positive, −12 volts is available for the aerial change-over relay RLD, and the receiver is muted. Either short-circuit or open-circuit to earth conditions are provided for receiver control. If S1 is returned to NET, relay RLB opens and the station returns to RECEIVE automatically.

If phone (A3) is selected, relay RLA is energized, providing 300 volts for the modulator and removing the short-circuit across the secondary of the modulation transformer. The key is short circuited by S1d. Relay RLA can be energized only when S2 is at TRANSMIT.

Power Supplies

The power supplies Fig. 6.101 use two power transformers: T2 supplies all potentials with the exception of the high voltage to the linear amplifier. MR9 to MR12 form a bi-phase rectifier, which could be replaced by a GZ34 valve if a suitable heater winding were available. Choke input filtering is incorporated (CH1 and C6), the nominal 300 volt line rising to about 325 volts on c.w. operation. V1, OA2 (or VR150/30), in conjunction with R19, provides a 150 volt stabilized line for V2 and V3. An anti-surge 500 mA fuse (F4) is used to avoid unnecessary power failures.

All bias supplies are derived from the 150 volt secondary, rectified by MR13 and smoothed by C9, C10 and C11. Three separate lines, one variable, run to the p.a., driver and keyed stages. Note that excellent smoothing is necessary on these lines if ripple is to be avoided, hence the large values of the smoothing capacitors.

For the 12 volt relay supply, four diodes type SX631 are used in a full-wave bridge, partially smoothed by C8. Valve heaters are connected by four separate lines to the 7 amp secondary to avoid a heavy voltage drop. LP1 indicates that the transmitter is switched on.

When the system switch S1 (Fig. 6.100) is at A1 or A3, RLC/2 (Fig. 6.101) is energized and mains (line voltage) applied to T1, indicated by LP2. A standard 450-0-450 volt 250 mA secondary feeds a full-wave bridge of eight silicon rectifiers. During testing it is recommended that only half the secondary is used until everything appears to be satisfactory. This supply is fused at 1 amp, to avoid the fuse blowing at switch on, due to the very low impedance presented by the series capacitor chain C3, C4 and C5. The parallel resistors which equalize voltages across the electrolytics must on no account be omitted. ***Warning*:** the insulated cans of C3 and C4 will be 800 volts and 400 volts above chassis respectively.

Components

Components in the v.f.o. tuned circuit must be of the highest quality. The inductor, L1, is wound with the wire under tension and is then placed in a warm oven for an hour. After the coil has cooled, it should be coated with polystyrene cement. The capacitors must be either silver mica or air-spaced with ceramic insulation. Ideally the main tuning capacitor C13 should be of the double end-bearing type, but availability could be a problem and the item specified is a good compromise. In wiring both oscillator circuits 18 s.w.g. wire should be used in lazy loops; with the wire taut it may contract or expand suddenly accompanied by an instantaneous frequency change.

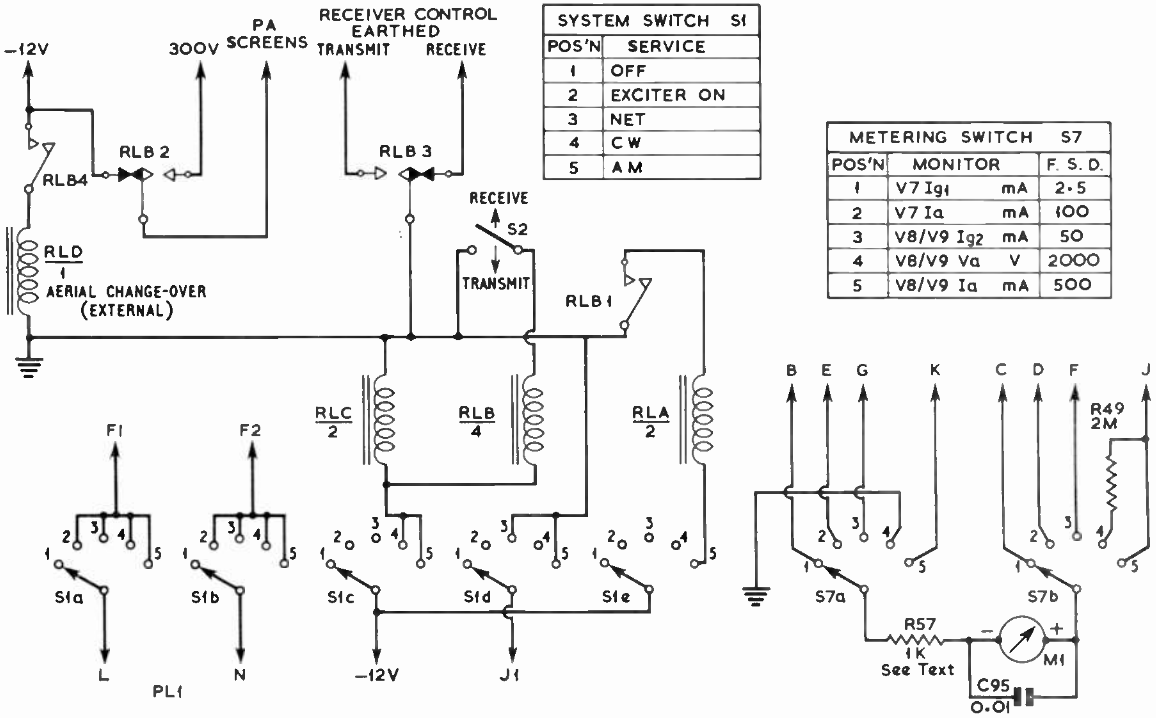


Fig. 6.100. Station control and transmitter metering circuits.

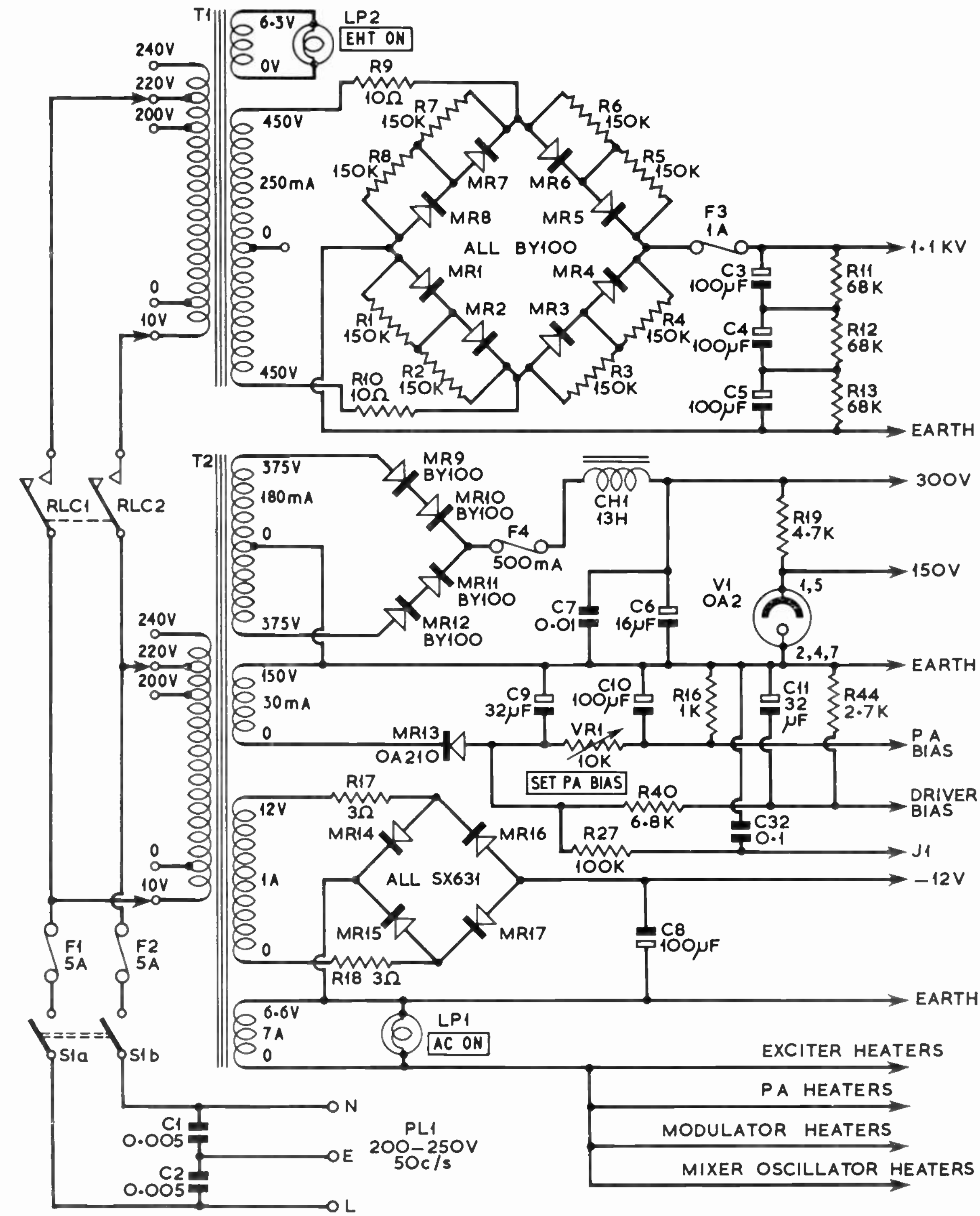


Fig. 6.101. Circuit diagram of the power supply for the Princess transmitter.

Wideband couplers T4 to T10 have been mentioned previously. Their construction is very simple: to obtain the correct spacing between the windings, cut a piece of Sellotape to the required dimension and stick it in the centre of the former. Do not forget to remove the Sellotape afterwards and also any pieces used to hold the wire terminations, otherwise the Q value of the coils will deteriorate with time. Use polystyrene cement to fix the turns. If the response of the 80m transformer is too narrow, about 1 pF top capacity coupling will widen the response sufficiently.

For S7, a ceramic Yaxley is essential with break-before-make contacts; this is most important with a potential difference of 1000 volts between adjacent positions. S6 has to carry 1000 volts d.c. as well as the full r.f. potential.

The bias and drive potentiometers must be wire-wound; the tracks of carbon controls in this application would soon fail. The linear amplifier passive grid resistor, R48, is made by forming two circles of 18 s.w.g. wire about 1 in. in diameter; with one wire above the other, each of the ten resistors is spaced around the circle. To decrease or increase r.f. drive, less or more resistors may be used. With only eight in circuit the transmitter will run up to 300 watts d.c. input on c.w. but will produce the wrong conditions for a.m.

If resistors with the value specified for R17, R18, and R56 are not available, they may be made with resistance wire. The valveholders for V5 and V6 should be p.t.f.e.; for V7, V8 and V9 ceramic is suitable.

No components in the speech amplifier and modulator are critical. The resistance of R69 should be higher than that of R67, although both are nominally 100 K ohms. All earth wiring is taken to a bus-bar earthed only at the microphone input socket.

TABLE 6.13

Inductor Details for the Princess Transmitter

L1	V.f.o. coil, 51 μH 89 turns 40 s.w.g. enam. close wound on Cambion 1533-0-2 former ⅜ in diam. No core. Tag ring at top and bottom.
L2, 3	Driver tank coil, 14 μH. Codar air spaced 35 turns 20 s.w.g. 1 in. diam. 3 in. winding length tapped at 23 turns (40m), 13½ turns (20m), 7½ turns (15 m), and 4½ turns (10m).
L4	Final tank coil, 1·6 μH. Codar air spaced 5 turns 14 s.w.g. 2 in. diam. 1 in. winding length tapped at 3¼ turns (15m), and 2 turns (10m).
L5	Final tank coil, 7·5 μH. Codar air spaced 17 turns 14 s.w.g. 2 in. diam. 3 in. winding length tapped at 8½ turns (40m), and 2 turns (20m).
L6	3·6 Mc/s trap, 25 μH. 70 turns 38 s.w.g. enamel close wound on Aladdin PP5938/S former with 9 mm. core.
T4, T6	Wideband coupler. 3·5–3·8 Mc/s. Primary and secondary 29·2 μH. Both windings 75 turns 40 s.w.g. enam., close wound. Spacing between windings 1/32 in. Aladdin PP5938/S former and can. 9 mm. cores.
T5, T7	Wideband coupler. 7·0–7·25 Mc/s. Primary and secondary 16·5 μH. Both windings 57 turns 38 s.w.g. enamel, close wound. Spacing between windings 7/32 in. Aladdin PP5938/S former and can. 9 mm. cores.
T8	Wideband coupler. 14·0–14·5 Mc/s. Primary and secondary 4·2 μH. Both windings 30 turns 32 s.w.g. enamel, close wound. Spacing between windings ¼ in. Aladdin PP5938/S former and can. 9 mm. cores.
T9	Wideband coupler. 21·0–21·45 Mc/s. Primary and secondary 2·0 μH. Both windings 20 turns 28 s.w.g. enamel, close wound. Spacing between windings 5/16 in. Aladdin PP5938/S former and can. 9 mm. cores
T10	Wideband coupler. 28·0–29·0 Mc/s. Primary and secondary 1·3 μH. Both windings 14 turns 20 s.w.g. enamel, close wound. Spacing between windings 3/16 in. Aladdin PP5938/S former and can. 9 mm. cores.

Note: The Aladdin PP5938/S former has a nominal diameter of [illegible] in. and the screening can is ¾ in. square. Design inductance values are quoted to assist if other formers are used, and are with cores fitted and the circuits resonated. All cores are h.f. types.

*Ready-wound coils are available from Electroniques (Prop: STD Ltd.) Edinburgh Way, Harlow, Essex.

Alignment Procedure

On completion of the transmitter, it is advisable to adopt a logical sequence of tests, thus ensuring that, for instance, h.t. supplies are not applied to a stage lacking bias due to a fault.

(1) Set S2 to RECEIVE, band switches to 80m, and key up.

(2) Before fitting the valves, fuses, relays or crystal, check the following lines to earth:

E.h.t. (C3 positive terminal) ..	200 K ohms.
H.t. (300 volts)	35 K ohms.
H.t. (150 volts)	40 K ohms approx.
Driver bias	2700 ohms.
P.a. bias	1000 ohms.
Relay supply (12 volts) ..	20 K ohms.
Mains at S1a/b	Infinity.

(3) Fit F1 and F2 only. Connect the mains (line voltage). Set VR1 to maximum resistance. Set SYSTEM switch (S1) to EXC. ON. LP1 should light. Check that the following voltages are present:

Junction MR10-MR11	350 volts positive to chassis.
Junction VR1-R16 ..	10 volts negative to chassis.
Junction R40-R44 ..	60 volts negative to chassis.
Across C8	16 volts negative to chassis.
Across LP1	6·6 volts a.c.

(4) Set the p.a. bias line to −45 volts by means of VR1.

(5) Fit the valves with the exception of V1. Check that the valve heaters light. Measure the heater voltage at pins of V7, V8 and V9: it should be between 6·0 volts and 6·7 volts a.c.

(6) Measure the negative voltage at pin 5 of V7: it should be approximately 60 volts.

(7) Fit V1 (OA2). Fit F4. The voltage across C6 should be 325 volts approximately. The voltage across V1 should be 150 volts.

(8) Check the anode current of V7 on meter position 2: it should be less than 30 mA.

(9) Switch to NET. Check that JK1 is shorted.

(10) Check the static voltages of V2 to V7 for any obvious irregularities.

(11) Disconnect the h.t. to the screens of V8 and V9 at R53.

(12) Fit relay RLB/4 and check that the disconnected screen supply (11) reads 15 volts negative.

(13) Set the SYSTEM switch (S1) to A1, and S2 to TRANSMIT. Relay RLB/4 should operate. Check that the disconnected screen supply (11) now reads 300 volts positive.

(14) Check that the receiver control (transmit) is earthed.

(15) Check that the feed to relay RLD/1 (aerial c/o) is 12 volts positive earthed.

(16) Set the SYSTEM switch (S1) to NET. Relay RLB/4 should open. Return S2 to RECEIVE. Check that the receiver

control (transmit) is open circuit and the receiver control (receive) is earthed. The supply to RLD/1 should be open circuit.

(17) Set C12 and C15 to approximately mid position, and C13 to 95 per cent of maximum capacity.

(18) Loosely couple a receiver covering 1·4 Mc/s to 2·1 Mc/s, to the anode circuit of V2. V2a should be oscillating in the region of 1·5 Mc/s. Adjust C12 so that 1·5 Mc/s corresponds to a dial setting of 20.

(19) The band 1·5 Mc/s to 2·0 Mc/s should be covered across the swing of C13 with an overlap at each end.

(20) Plug in the 5·5 Mc/s crystal. With a receiver, check that V2b is oscillating at this frequency. The note should be pure. Disconnect R46 at the screen grid of V7.

TABLE 6.14

Components List for the Princess Transmitter

Circuit ref.	Description	Quantity
	Capacitors (fixed)	
C1, 2	0·005 µF, 1500V wkg., 20%	2
C3, 4, 5	100 µF electrolytic, 450V wkg.	3
C6	16 µF electrolytic, 350V wkg.	1
C7, 20, 21, 26, 34, 37, 38, 39, 41, 49, 50, 51, 58, 59, 60, 64, 66, 69, 78, 80, 82, 83, 84, 89, 90, 95	0·01 µF disc ceramic, 350V wkg.	26
C32	0·1 µF paper, 350V wkg.	1
C8	100 µF electrolytic, 18V wkg.	1
C9, 11	32 µF electrolytic, 150V wkg.	2
C10	100 µF electrolytic, 100V wkg.	1
C14	10 pF ceramic, N750L, 20%	1
C16, 36, 47, 56, 63, 94	100 pF silver mica, 350V wkg., 10%	6
C17, 18	1000 pF silver mica, 350V wkg., 2%	2
C19, 33, 35, 52, 54	10 pF silver mica, 350V wkg., 10%	5
C22, 42	56 pF silver mica, 350V wkg., 10%	2
C23	82 pF silver mica, 350V wkg., 10%	1
C24	39 pF silver mica, 350V wkg., 10%	1
C25	680 pF silver mica, 350V wkg., 10%	1
C30	43 pF silver mica, 350V wkg., 2%	1
C31, 43	33 pF silver mica, 350V wkg., 2%	2
C44, 61	15 pF silver mica, 350V wkg., 2%	2
C68	0·001 µF mica, 750V wkg., 10%	1
C71	20 pF ceramic, Hunts CDG, 5%	1
C72, 73	100 pF ceramic, Hunts CDG, 5%	2
C74	150 pF ceramic, Hunts CDG, 5%	1
C75	180 pF ceramic, Hunts CDG, 5%	1
C76	330 pF ceramic, Hunts CDG, 5%	1
C77	840 pF (510 pF + 330 pF)	
C85, 87, 91	0·005 µF mica, 3 kV wkg., 20%	3
C96, 98, 100, 103	8 µF electrolytic, 350V wkg.	4
C97, 101	2 µF electrolytic, 12V wkg.	2
C99, 102	0·002 µF paper, 350V wkg., 20%	2
C104, 105	0·01 µF paper, 350V wkg., 20%	2
C106	25 µF electrolytic, 25V wkg.	1
	Capacitors (variable)	
C12, 15, 48, 57	3–30 pF air spaced trimmers	4
C13	100 pF Eddystone Type 585	1
C70	60 pF Eddystone Type 582	1
C86	250 pF Eddystone Type 817	1
C88	1500 pF (three-gang 500 pF)	1
	Resistors	
R1, 2, 3, 4, 5, 6, 7, 8	150 K ohms, ½ watt, 10%	8
R9, 10	10 ohms, 3 watts, wirewound, 10%	2
R11, 12, 13	68 K ohms, 6 watts, wirewound, 10%	3
R16, 65	1 K ohm, 1 watt, 10%	2
R17, 18	3 ohms, 1 watt, wirewound, 10%	2
R19	4·7 K ohms, 6 watts, wirewound, 10%	1
R20, 27, 58	100 K ohms, ¼ watt, 20%	3
R21, 33, 37, 41, 42, 50	1 K ohms, ¼ watt, 20%	6
R22, 36	4·7 ohms, ¼ watt, 20%	2
R23, 29, 31, 34, 38	47 K ohms, ¼ watt, 20%	5
R24	5·6 K ohms, ¼ watt, 20%	1
R25, 26	470 K ohms, ¼ watt, 20%	2
R28	8·2 K ohms, ¼ watt, 20%	1
R30	330 ohms, ¼ watt, 20%	1
R32, 46	10 K ohms, 1 watt, 10%	2
R35, 39	470 ohms, ¼ watt, 20%	2
R40	6·8 K ohms, 1 watt, 10%	1
R43	430 ohms, ¼ watt, 1% hi-stab.	1
R44	2·7 K ohms, 1 watt, 10%	1
R45	10 ohms, ¼ watt, 20%	1
R47	11 ohms, ¼ watt, 1% hi-stab.	1
R48	100 ohms. Consists of 10 1K ohm 1 watt 10% resistors in parallel. See text.	
R49	2 M ohms, ½ watt, 1% hi-stab.	1
R51, 52	10 ohms, ½ watt, 10%	2
R53	22 ohms, ¼ watt, 1% hi-stab.	1
R54, 55, 75	100 ohms, ½ watt, 20%	3
R56	2·2 ohms, ½ watt, 2%	1
R57	To increase total meter resistance to 1·1 K ohms ± 2%, ½ watt. See text	
R59, 62, 68	3·3 K ohms, ¼ watt, 10%	3
R60, 63	220 K ohms, ¼ watt, 5% hi-stab.	2
R61, 64	39 K ohms, ¼ watt, 20%	2
R66	1 M ohm, ¼ watt, 20%	1
R67, 69	100 K ohms, ¼ watt, 10%	2
R70, 71	220 K ohms, ¼ watt, 10%	2
R72	330 ohms, 3 watts, wirewound, 5%	1
R73	56 ohms, ¼ watt, 20%	1
R74	1 K ohms, 3 watts, wirewound, 10%	1
R76	1·5 K ohms, 5 watts, wirewound, 10%	1
VR1	10 K ohms, 3 watts, wirewound, 10%	1
VR3	100 K ohms, 3 watts, wirewound, 20%	1
VR4	1 M ohm, carbon log, 20%	1
	Miscellaneous	
CH1	12H 200 mA Partridge TF6811	1
F1, 2	Cartridge fuse, 5 amp	2
F3	Cartridge fuse, 1 amp	1
F4	Cartridge fuse, 500 mA, anti-surge	1
LP1, 2	Indicator bulb, 8V 0·15 amp	2
Meter	0–1 mA panel mounting. See R57 for value of internal meter resistance	1
MR1 to MR12	BY100, 800V p.i.v., 500 mA	12
MR14, 15, 16, 17	5X631, 100V p.i.v., 750 mA	4
MR18	OA85	1
MR13	OA210, 400V p.i.v.	1
RFC1, 3, 4	2·5 mH Electroniques	3
RFC5	470 µH Electroniques	1
RFC2, 6, 7	7 turns 22 s.w.g. enamel wound over R54, R55, R75	3
RFC8	100 turns 32 s.w.g. on ceramic tube 4 in. × 1 in., winding divided into 3 sections.	1
RLA/2, RLC/2	d.p.d.t. 12V operation	2
RLB/4	4 pole changeover, 12V operation	1
RLD/1	s.p.d.t., aerial changeover	1
S1	4 bank 2 pole 5 way 250V 5 amp make before break	1
S2	Toggle, s.p.s.t.	1
S3, 4, 5	2 bank 2 pole 5 way	3
S6	1 bank 1 pole 5 way, heavy duty	1
S7	1 bank 2 pole 5 way, ceramic	1
T1	Mains transformer. Primary 200–250V, 50 c/s. Secondaries 450–0–450V 250 mA, 6·3V 0·3 amp. Partridge TF6809	1
T2	Mains transformer. Primary 200–250V, 50 c/s. Secondaries 375–0–375V 180 mA, 12V 1 amp, 6·7V 7 amps, 150V 20 mA. Partridge TF6810.	1
T11	Modulation transformer. Woden UM0.	1
V1	OA2-150C2	1
V2	ECC81, 12AT7	1
V3, 11	ECC82, 12AU7	2
V4, 5, 6	EF91, 6AM6	3
V7	QV06/20, 6146	1
V8, 9	TT21, 7623	2
V10	ECC83, 12AX7	1
V12, 13	EL85	2
X1	Crystal, 5·5 Mc/s, 0·01%, 30 pF	1

(21) Set C13 to correspond to 1·970 Mc/s, giving a frequency of 3·530 Mc/s at the anode of V3.

(22) Connect a test meter on a low voltage range (2·5 volt f.s.d.) to TP1. Set VR3 to minimum.

(23) Peak the reading by adjustment of the cores in T4.

(24) Set C13 to 1·73 Mc/s to give a mixed frequency of 3·77 Mc/s. Readjust T4 to equalize the drive over the whole range 3·5 Mc/s to 3·8 Mc/s. Note that it may prove necessary to alter the values of C30 and C31 within small limits.

(25) Select 40m on the EXCITER switch. Tune T5 and select C33 and C35 if necessary. The tracking frequencies are 7·030 Mc/s and 7·220 Mc/s and the range 7·000 Mc/s to 7·250 Mc/s.

(26) Loosely couple the receiver to the grid circuit of V7. Set the band switch to 80m, frequency to 3·530 Mc/s, S7 to grid of V7 and VR3 to maximum.

(27) Tune T6. Note that the meter has a suppressed zero and may not read until resonance is closely approached.

(28) Compromise adjust T6 across the whole band, altering the tuning of T4 to improve the broadband response. Check that VR3 controls the drive level to V7.

(29) Repeat for 40m, tuning T7. C45 may not be needed due to the high stray capacities in circuit.

(30) Select 20m. Tune T8, then peak the reading with C48. Note that maximum drive to V7 on the three higher bands may not occur at the maximum setting of VR3 and this control should be set to the optimum position.

(31) Select 15m. Tune T9. Only a slight indication or even no reading may be noticed. Do not despair at this stage. Rely on the receiver for indications of resonance.

(32) Select 10m. Tune T10. Peak with C57. Similar remarks apply.

(33) Reconnect R46 to the screen grid of V7. Select 80m and set the v.f.o. to 3·530 Mc/s. Set S7 to read the anode current of V7. Tune P.A. GRID control. The anode current to V7 should be about 35 mA at resonance.

(34) Repeat for all bands. The figure for 10m operation at resonance should be in the region of 50 mA.

(35) It may be found necessary to alter slightly the values of C73-C77. An increase in capacity reduces the loading.

(36) Under (31) and (32) it was stated that little indication of drive would be obtained. Accurate tuning may be effected by setting C70 well off resonance and tuning T9 (for 15m) and T10 (for 10m) for maximum reading of anode current to V7. At all frequencies this should exceed 75 mA.

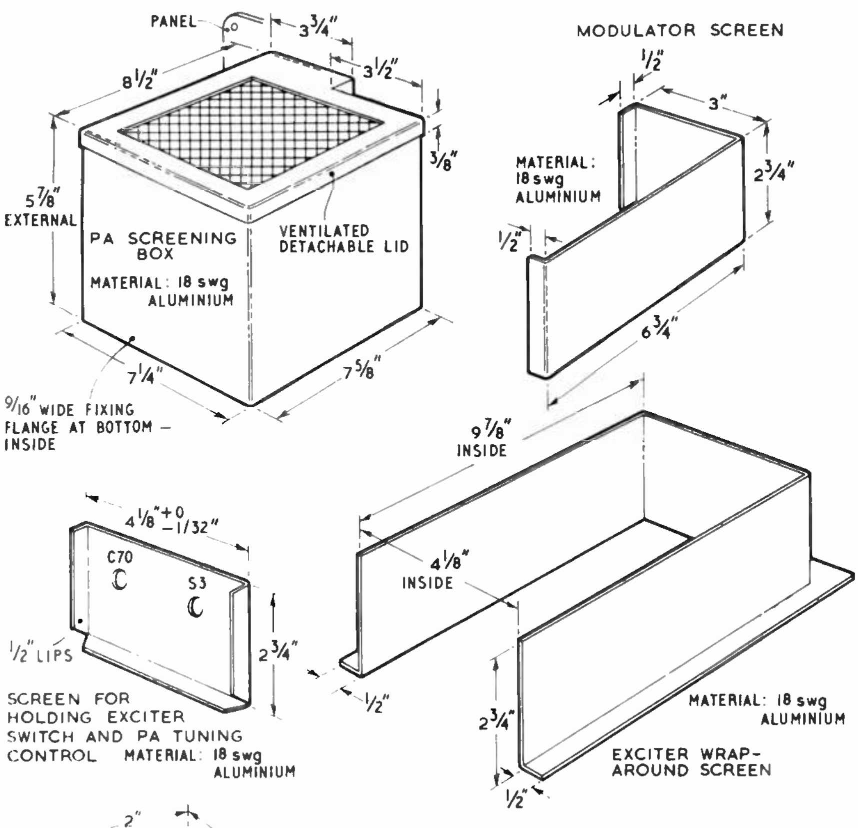
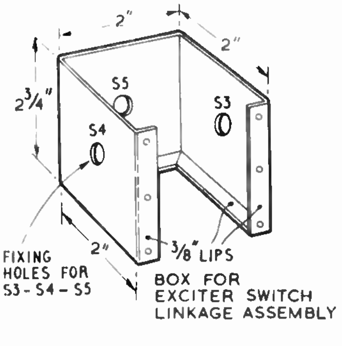


Fig. 6.102. Construction of the screens



(37) The exciter may be tried as a low power c.w. transmitter by disconnecting the lead to the p.a. grid resistor, R48, and feeding a low impedance aerial from the low power socket. After this test, reconnect R48.

(38) Fit relay RLC2. Check that F3 is not fitted. Connect either R9 or R10 to the centre tap of T1 as a precaution, Key up.

(39) Switch to A1. Check that the voltage at the bridge output is approximately 430 volts d.c.

(40) Fit F3. Check that h.t. supply is now approximately 650 volts d.c.

(41) Set the METER switch to p.a. screen current. Set S2 to TRANSMIT. A reading of less than 5 mA should be obtained.

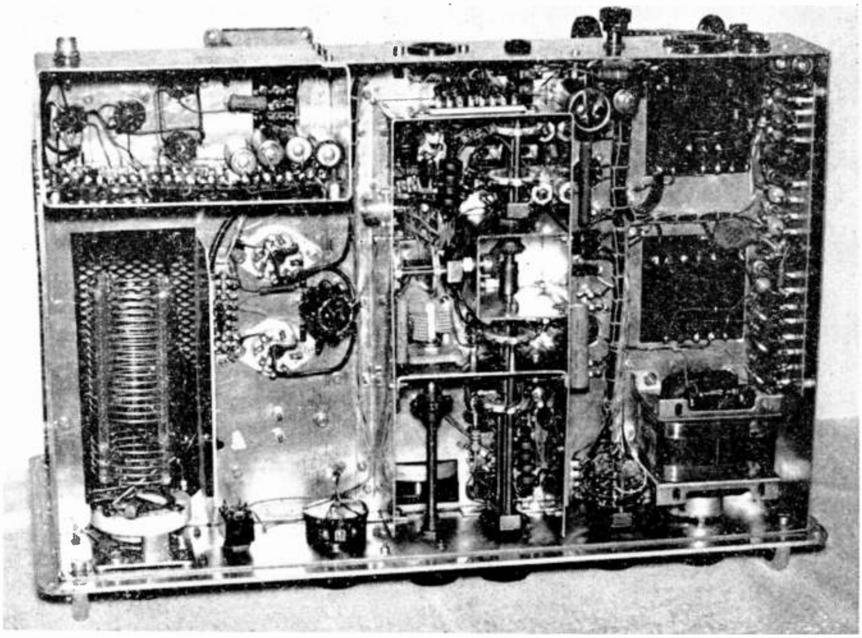
(42) Set the METER switch to p.a. anode current. A reading of approximately 20 mA should be obtained.

(43) Connect R9/R10 across the full h.t. output of T1. The d.c. output should be approximately double that obtained at step (40).

(44) Set VR1 to give a standing p.a. anode current of 70 mA.

(45) Connect a 75 ohm 150 watt load to the r.f. output socket.

(46) On 80m, with key down, drive the amplifier, loading to give a p.a. screen current at resonance of 25 mA. This corresponds to approximately 150 watts input.



Under chassis view of the Princess transmitter. The exciter occupies the central portion of the chassis and the modulator is at the top left corner.

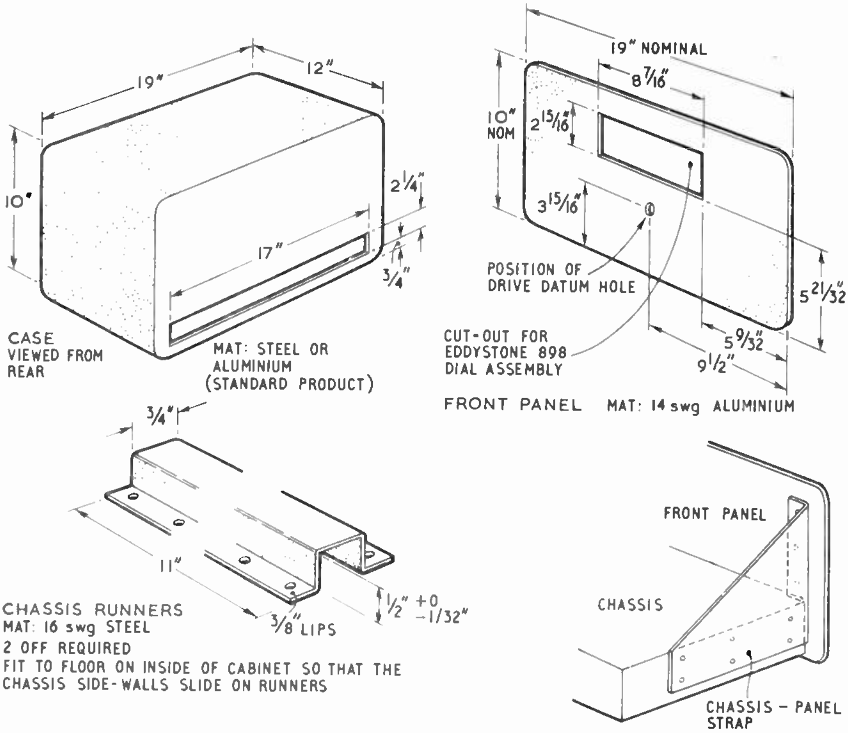


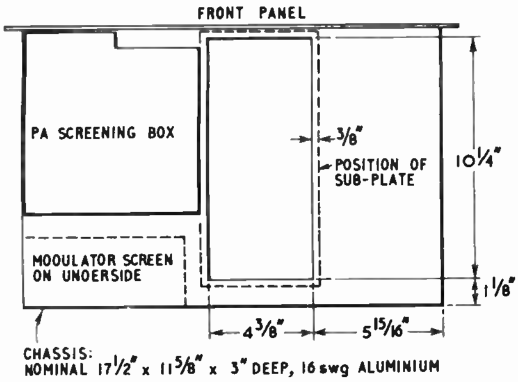
Fig. 6.103. Essential cabinet and panel dimensions for the Princess.

(47) Set the v.f.o. to an output frequency of 3700 kc/s. With the receiver tuned to 3600 kc/s a spurious output may be heard. Adjust the series resonant circuit L6/C24 to minimize this signal. It is due to the second harmonic of the v.f.o. on 1800 kc/s.

(48) Repeat step (46) on all bands. On the two higher frequency bands, the P.A. GRID tuning control must be adjusted for maximum drive.

(49) Final checking of the modulator may now be carried out. Fit relay RLA/2. Set VR4 to minimum. Turn SYSTEM switch to A3. P.a. anode current should drop to about 90 mA.

(50) Modulation can now be applied. The p.a. anode current should peak to only about 120 mA.



Layout of the main assemblies.

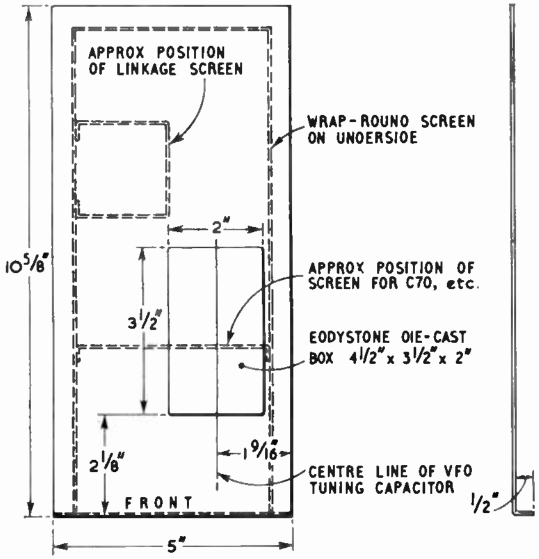


Fig. 6.104. Construction of the exciter sub-assembly.

Operation

Select the band required, v.f.o. on frequency, set METER switch to anode current of V7 and SYSTEM switch to NET. Set the DRIVE control at maximum, tune C70 for a dip in anode current; check that the grid current of V7 does not exceed 1 mA, adjusting VR3 if necessary.

Switch to p.a. screen current and to A1. With the key down, tune the p.a. controls C86 and C88 for a 25 mA peak in screen current. This should approximate to an input of 150 watts as indicated by the two remaining metering positions. With the key up, the anode current should be 70 mA, to which it had been set during alignment.

For phone operation, switch to A3; p.a. anode current should drop to around 90 mA. Adjust VR3 to obtain this reading (usually less drive is required). With VR4 set to suit the operator's voice level, modulation peaks should cause the anode current to rise to 120 mA. Care should be taken not to overdrive the amplifier, otherwise it will flat-top and splatter and TVI will result.

S.s.b. Operation

The construction of an s.s.b. exciter for the Princess transmitter was described in the July, August and September 1966 issues of the *RSGB Bulletin*. This exciter fits into the space occupied by the exciter described in this chapter.

A VALVE BUFFER FOR A TRANSISTOR V.F.O.

When a transistor v.f.o. is used with a valve transmitter, it is desirable to provide additional isolation by means of class

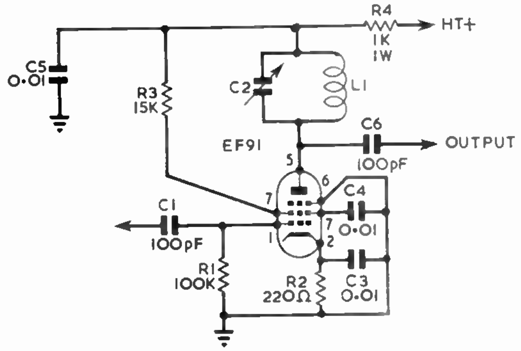


Fig. 6.105. Class A buffer amplifier suitable for use between the transistor v.f.o. of Fig. 6.36 and a valve transmitter. Sufficient output is given to drive either a 6BW6 or a 5763. Values of C2 and L1 are given in Table 6.15. The valve may be an EF91 or a 6AM6.

A valve buffer stage. The circuit of such a buffer, intended for use with the v.f.o. of Fig. 6.36 is given in Fig. 6.105. The input of the buffer amplifier is capacity coupled to the collector of the output transistor (Fig. 6.36).

The peak r.f. voltage available at the collector of TR2 is about 2·5 at 3·5 Mc/s and slightly less in the 11 Mc/s version, adequate in both cases to drive a valve buffer in class A. To ensure the valve operates in class A, a 500 μA meter should be connected in series with the earthy end of the 100K ohm grid resistor R1. If grid current is indicated, C1 should be reduced in value. The output of the stage is sufficient to provide 2·5–3mA grid current through 22K ohms to a QV03-12 (5763) or a 6BW6.

TABLE 6.15

Frequency	L1 Turns	Wire	C2
3·5 Mc/s	75	34 s.w.g.	100 pF
11 Mc/s	19	22 s.w.g.	25 pF

Coils are close wound on ½ in. formers. The wire is enamelled covered.

CHAPTER 7

V.H.F./U.H.F. TRANSMITTERS

TRANSMITTERS for the 4m, 2m, 70cm and 23cm (70, 144, 430 and 1296 Mc/s) bands employ the same general principles as those used on the h.f. bands: a stable oscillator followed by frequency-multiplying and power-amplifying stages.

A major difference between h.f. and v.h.f. transmitters is that the v.h.f. equipment is usually designed to operate on one band only. This is mainly due to the r.f. losses that would result from a band-switching device, and also to the difference in the form of tuned circuit, the transition from lumped circuits to parallel-line circuits and ultimately from parallel-line to coaxial-line circuits as the frequency is increased. Lumped circuits may be used at 70 and 144 Mc/s but line or cavity circuits are generally more satisfactory at 430 and 1300 Mc/s.

As the 2m band and part of the 70cm band are in harmonic relationship it is common practice to use an existing 2m transmitter as a driver, the output of the 2m transmitter being applied to a push-pull tripler stage which in turn is used to drive a power amplifier on 70cm. With this arrangement a transmitter operating between 144 and 146 Mc/s will produce an output having a frequency between 432 ahd 438 Mc/s. Similarly, a 70cm p.a. can be used to drive a tripler to 23cm, resulting in frequencies between 1296 and 1314 Mc/s. The usual operating range for frequency stabilized equipment on this band is 1296–1300 Mc/s.

Although rapid progress has been made in the production of semiconductors, they are still relatively expensive for operation on v.h.f./u.h.f. at comparable power levels to the valves available.

CRYSTAL CONTROLLED OSCILLATORS

The majority of v.h.f. transmitters are crystal controlled, although in some 2m transmitters the drive is derived from a stable variable frequency oscillator. To be able to tune the transmitter to the same frequency as that of the station being called, which is one of the customary requirements for operation on the h.f. bands, is not so necessary in v.h.f. practice.

The crystal-oscillator circuits used in v.h.f. transmitters are similar to those employed for the lower frequencies. Crystal frequencies are usually in the range 6–12 Mc/s, and owing to the high multiplication factor in the transmitter to reach the final radiated frequency, the oscillator arrangement adopted is generally one that produces a large harmonic output. A common example is the modified Pierce oscillator in which the anode circuit is tuned to the third or higher odd harmonic of the crystal frequency. Suitable valves in the miniature range to employ in this circuit are the EF91 (CV138, 6AM6), EL91 (CV136, 6AM5) and EF95 (6AK5).

Overtone Oscillators

In addition to operation on their fundamental frequency it is possible for crystals to oscillate on odd multiples, or overtones, of the fundamental. The term overtone is used in preference to harmonic operation, as the resultant frequency of oscillation is not an exact multiple of the crystal frequency.

The most suitable crystals for overtone operation are AT-cut or BT-cut, operating in the thickness shear mode, in which the frequency is determined by the thickness of the crystal. **Figs. 7.1 (a)** and **(b)** show the deformation in the

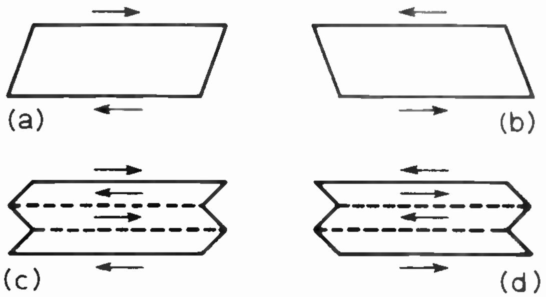


Fig. 7.1. Operation of a crystal on its fundamental and third overtone.

thickness of a crystal operating on its fundamental frequency at successive peaks of each half-cycle. **Figs. 7.1 (c)** and **(d)** show the deformation of a crystal operating on its third overtone; in each instance the arrows indicate the direction of mechanical movement. It will be seen that the crystal now operates as if it consists of three distinct layers; for fifth overtone operation it would appear to consist of five layers and so on. When the crystal is functioning as an overtone oscillator, there is no fundamental frequency component, the lowest frequency generated being that of the particular overtone.

For successful overtone operation crystals must be very accurately ground. At the frequencies used for oscillators in v.h.f. transmitters each layer of the crystal is only a few thousandths of an inch thick, and any irregularity in the grinding will result in failure to oscillate. Excessive pressure on the surfaces of the crystal by the mounting will also prevent oscillation.

Crystals specially processed for overtone are available giving output frequencies up to approximately 200 Mc/s. Certain crystals ground for fundamental frequency operation are suitable for use on the third and fifth overtones, notably the type FT243 in the frequency range 5·5 to 8·5 Mc/s. In transmitter circuits it is recommended that operation is restricted to the third or fifth overtone, due to the lower output voltage obtainable at the higher orders.

To cause a crystal to oscillate on its overtone additional feedback at the overtone frequency must be provided. Only sufficient feedback to sustain oscillation should be used: too much feedback will result in the crystal losing control and self-oscillation taking place. Adjustment of feedback is usually quite critical, fundamental frequency crystals requiring more feedback than those specially processed.

The evacuated B7G type of crystal with sputtered electrodes is very suitable for overtone operation, provided that a moderate h.t. voltage only is applied to the oscillator and feedback is not excessive. Too high an h.t. voltage will result in over-heating and possible fracture of the crystal.

The advantages of employing overtone oscillators in v.h.f. transmitters are that the lowest frequency generated is that of the overtone and there is less chance of causing spurious emissions than from a fundamental frequency oscillator. Disadvantages are that initial adjustment to obtain good frequency stability may be critical, and that the overtone frequency generated is not an exact multiple of the crystal frequency unless it is cut and calibrated on its overtone frequency; this must be borne in mind when working close to the edges of a frequency band.

Typical overtone oscillator circuits are shown in **Fig. 7.2.** The simple circuit of **Fig. 7.2 (a)** is suitable for use with crystals intended for overtone operation but cannot be employed with fundamental crystals as insufficient feedback is provided to cause oscillation. L1 is tuned to the overtone frequency.

The modified Pierce oscillator of **Fig. 7.2 (b)** uses a triode connected high slope pentode such as the 6AM6/EF91 and will give good harmonic output up to the ninth. The anode circuit LC is tuned to the required frequency.

Fig. 7.2 (c) shows the Squier circuit which is in common use with fundamental type crystals in transmitters and as the local oscillator in v.h.f. converters. Feedback between anode and grid circuits is by direct inductive coupling, the degree of feedback being determined by the position of the h.t. tap on the inductor L1; the tap is at earth potential to r.f. and feedback increases with the number of turns between the tap and the grid end of the inductor.

When overtone crystals are used in this circuit the amount of feedback required is small, but it must be increased with fundamental type crystals. L1 is slug tuned to the frequency of the particular overtone selected. The position of the tap should be adjusted to provide sufficient feedback to initiate and maintain oscillation only and should be approximately one-quarter of the total number of turns from the grid end of the winding. This may require adjustment when initially setting up the oscillator. As with all overtone circuits too much feedback will result in the crystal losing control and self-oscillation taking place.

The circuit shown in **Fig. 7.2 (d)** is a transistor overtone oscillator suitable for third or fifth harmonic use. L2 is tuned to the output frequency, the collector may be tapped down to about one-third up from the earthy end. An OC171 transistor or an equivalent is suitable for use in this circuit.

Fig. 7.2 (e) illustrates the Robert Dollar circuit. In this circuit feedback is controlled by the two capacitors C2 and C3. Decreasing the capacitance of C2 increases the feedback. Typical values for C2 and C3 are shown in the diagram, but if desired, C2 may be a pre-set capacitor to give some flexibility of adjustment. L1 is tuned to the overtone frequency.

A variant of the Butler circuit, due to Cathodeon Crystals Ltd., is shown in **Fig. 7.2 (f)**. This circuit, which is suitable for use with overtone crystals, consists of a grounded grid amplifier, (V1a) and a cathode follower (V1b). The anode circuit of the amplifier is tuned to the overtone frequency; the anode circuit of the cathode follower is tuned to a harmonic of this frequency, usually the second or third, as output at harmonics higher than these is generally too low to provide adequate drive in transmitter circuits. If r.f. output at the overtone frequency only is required this may be obtained from an inductive coupling to L1; in this instance

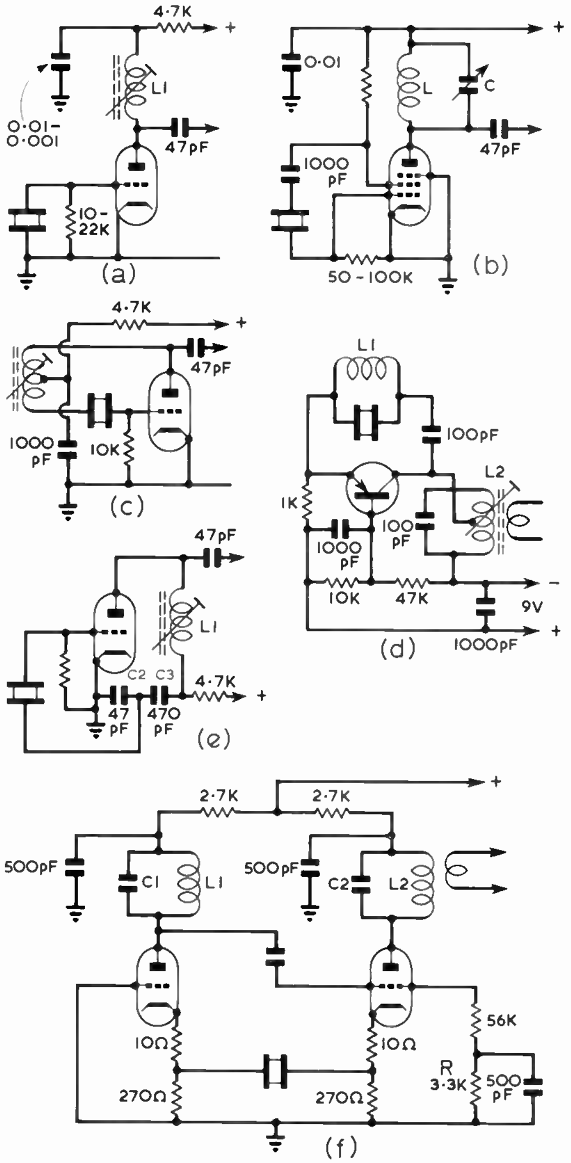


Fig. 7.2. Typical overtone crystal oscillators. (a) Simple type suitable for use crystal cut for overtone operation. (b) The modified Pierce circuit. (c) Squier type. (d) A typical transistor oscillator for third or fifth overtone operation. (e) The Robert Dollar circuit. (f) The modified Butler circuit.

the anode of the cathode follower should be earthed to r.f. and the tuned circuit L2, C2 omitted.

It is recommended that low Q inductors are used in this circuit. A coil wound on a polystyrene former and tuned by an adjustable dust-iron core is suitable. The resistors in the cathode circuits of the valves form a transmission line and should be matched in value to achieve lowest loss. It may be necessary to experiment to obtain the optimum value for these components because the r.f. output rises but the frequency stability of the circuit falls as the value of resistance is increased. Lowering the value of the resistors increases the stability but the output will be reduced to a point at which oscillation ceases. A compromise should be aimed at between good frequency stability consistent with r.f. output adequate to drive the succeeding stage of the transmitter.

Triode valves are shown in all the circuits previously described but tetrodes and pentodes may also be used satisfactorily. Valves in the miniature range are to be preferred; the normal practice is to use double triodes, one section functioning as the oscillator, the output of which is fed into the grid of the second section, which acts as a frequency multiplier. Suitable valves to employ in the circuits of Fig. 7.2 are the ECC81 (CV455, 12AT7), ECC91 (CV858, 6J6), ECC82 (CV491, 12AU7) and ECC85 (6AQ8). The triode pentode ECL80 (6AB8) is suitable for use as a triode overtone oscillator and pentode frequency multiplier. A recommended valve in the double triode range for the Butler circuit of Fig. 7.2 (f) is the ECC81 (12AT7); suitable pentodes for this circuit are two type EF95 (CV850, 6AK5).

As with the more normal type of crystal oscillator, resonance in an overtone circuit is indicated by minimum anode current. It will be found when tuning the circuit that the anode current will fall very sharply on the low frequency side and will rise rather more gradually on the high frequency side of resonance. The final adjustment should be such that the circuit is tuned slightly to the high frequency side of the resonant point, which will result in maximum stability being obtained.

The resonance condition of the oscillator circuits described in this Chapter may be conveniently determined by tuning for maximum grid current in the following stage. The overtone circuit L1, C1 of Fig. 7.2 (f) may be adjusted for maximum grid current in V1b, as indicated by a milliammeter connected across R.

FREQUENCY MULTIPLIERS

Frequency multiplying stages usually incorporate tetrode or pentode valves for greatest sensitivity and efficiency; high sensitivity is desirable because a large amount of grid drive may not be available from the preceding stage, particularly as the frequency of operation is increased. As the frequency increases the efficiency of multiplying stages falls off and for this reason it is advisable that any quadrupling or tripling is confined to the early stages of the transmitter and that doubling only takes place in the later stages where the efficiency is lowest.

To minimize the generation of unwanted frequencies, which would increase the possibility of interference to other services, multiplying stages should be designed to give maximum output at the desired harmonic and minimum output at all other frequencies. This is achieved by the use of high Q tuned circuits, inductive or link coupling between stages, and operation at the lowest power level consistent with obtaining adequate drive to the succeeding stage.

At frequencies up to 145 Mc/s single ended stages are usually employed. Most of the valves listed in **Table 7.2** are satisfactory for use as multipliers; suitable types for low power operation are the EF91 (CV138, 6AM6), QV03–12 (CV2129, 5763) and QV04–7 (CV309). R.F. double tetrode valves are commonly used as push-pull frequency triplers and can also be used as push-push doublers with the two anodes connected in parallel.

Triode frequency multipliers are sometimes employed in low power fixed and mobile transmitters, and their use in crystal-oscillator/multiplier circuits has already been mentioned in the section dealing with overtone oscillators. At higher frequencies, for example as triplers from 144 Mc/s to 430 Mc/s, triodes such as the A2521 (6CR4), 5842 (417A), DET22 (CV273, TD03–10), EC8010, DET23 (CV354, TD03–5), and DET24 (CV397, TD04–20) will be found suitable. Varactor diodes and some u.h.f. power transistors may also be used as frequency multipliers.

TABLE 7.1

For the early stages of much of the equipment used at v.h.f./u.h.f., small valves such as the ECC91/6J6 double triode, EF91/6AM6 and E180F(6688) pentodes are suitable and of high efficiency. The following data are a representative guide.

EF91/6AM6 with anode voltage of 300V, screen voltage of 250, and grid bias of −20V.

Class of operation	f_{input} (Mc/s)	f_{output} (Mc/s)	Anode current (mA)	Screen current (mA)	Grid current (mA)	Power ouput (watts)
Amplifier	60	60	11·5	3·2	1·5	1·9
	120	120	11·5	3·2	1·7	1·7
	200	200	9·7	3·2	1·9	0·4
Frequency Doubler	30	60	10·0	3·2	1·5	1·5
	65	130	10·0	3·2	1·6	1·1
	90	180	9·0	3·2	1·8	0·25
Frequency Tripler	20	60	10·5	3·2	1·6	1·1
	30	90	10·0	3·2	1·7	0·9
	50	150	9·0	3·2	1·8	0·2

ECC91/6J6 with anode voltage of 200.

Class of operation	Frequency (Mc/s)	Cathode Current (mA)	Anode Current ($I_{a(tot)}$) (mA)	Grid Current ($I_{g(tot)}$) (mA)	Grid Resistor (Common) (K ohms)	Power Output (Watts)
Amplifier (Grid bias −14V) Cathode resistor 330 ohms	50	22	17·2	4·8	1·5	2·1
	100	22	17·8	4·2	1·5	1·9
	150	22	18·2	3·8	1·5	1·7
	200	22	18·8	3·2	2·2	1·4
	250	22	19·3	2·7	2·7	1·1
Frequency Tripler (Grid bias −100V Cathode resistor 120 ohms)	50	22	16	6	15	0·95
	100	22	16·7	5·3	18	0·9
	150	22	17·2	4·8	22	0·8
	200	22	17·7	4·3	22	0·7
	250	22	18·2	3·8	27	0·6

POWER AMPLIFIERS

Power amplifying stages may employ triode or tetrode valves, used either singly or in push-pull. Tetrodes have the advantage that the power gain is greater and a lower grid drive is required, an important factor at the higher frequencies where in many instances only a low output is available from the preceding driver stage.

For low and medium power 2m and 70cm transmitters undoubtedly the most popular valve is the r.f. power double tetrode, some versions of which are suitable for operation up to 600 Mc/s. In the low power class with up to 10 watts anode dissipation, these valves are of miniature single-ended construction; examples are the Mullard QQV02–6 (6939) and QQV03–10 (6360). For higher power ratings up to a maximum of 40 watts anode dissipation a double-ended construction has been adopted, as exemplified by the Mullard QQV03–20A (6252/AX9910) and QQV06–40A. These valves incorporate internal neutralizing capacitances and no external neutralization is necessary provided that input and output circuits are shielded from each other.

Alternatively, grounded grid triodes for use as frequency multipliers, and amplifiers, offer somewhat simpler circuits. With such valves as the A2521 (6CR4), EC8010 (8556), DET22 (TD03–10) and DET24 (TD04–20) equally good power gain above 430 Mc/s may be attained.

Single tetrodes are commonly employed in high power 2m transmitters, or a pair may be used in push-pull; suitable valves are the Mullard QY3–65 and QY3–125. For 70cm a typical high power amplifier consists of a 4X150A forced air cooled tetrode in a coaxial circuit; such an amplifier can be operated at the maximum input of 150 watts at 435 Mc/s. Amplifiers for 23cm make use of planar electrode triodes in a coaxial circuit.

A list of valves suitable for operation as power amplifiers and in the later frequency multiplying stages of v.h.f. and u.h.f. transmitters appears in Table 7.2. Detailed information on these types and in many instances application notes may be obtained from the valve manufacturer.

TUNED CIRCUITS

At frequencies up to approximately 150 Mc/s lumped circuits are efficient and are generally used throughout the transmitter in both multiplying and amplifying stages. Above this frequency the efficiency of an inductor-capacitor tuned circuit falls off and use is made of parallel resonant lines up to frequencies of approximately 500 Mc/s At higher frequencies the spacing between parallel lines becomes appreciable compared to the wavelength and loss occurs due to radiation from the lines; for this reason coaxial resonant lines in which the outer conductor is at earth potential to r.f. are employed at ultra high frequencies. The transition from one type of tuned circuit to the other takes place gradually and some overlap exists in the choice of a particular tuned circuit for a given frequency; for instance although the majority of 144 Mc/s transmitters make use of lumped circuits throughout, a quarter wavelength line is sometimes employed as the tank circuit in a medium or high power amplifying stage. At 430 Mc/s the tendency is to use parallel-line circuits at low and medium powers and coaxial or cavity circuits in high power transmitters.

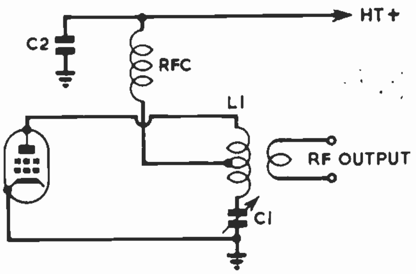


Fig. 7.3. Series tuned anode circuit.

Examples of each type of tuned circuit previously discussed appear in the transmitters described later in this chapter.

The inductors used in lumped circuits should be of self-supporting air spaced construction, and can conveniently be wound with soft drawn enamelled copper wire, having a gauge between 12 and 22 s.w.g., the wire gauge increasing as the power rating of the stage and the frequency of operation increases. For power ratings in excess of 100 watts, it is advisable to use $\frac{3}{16}$ or $\frac{1}{4}$ in. o.d. copper tube, preferably silver plated to minimize loss due to skin effect which can be appreciable at the frequencies under consideration. As a v.h.f. transmitter is essentially a one band device, an inductor should be soldered directly across the associated tuning capacitor; this eliminates any plug and socket connection which could cause losses if used in an r.f. circuit.

In tuned circuits associated with driver stages it is common practice to dispense with the tuning capacitor and to resonate the inductor with the stray and circuit capacitances. Tuning is effected by opening out or compressing the turns of the inductor.

Certain single-ended tetrodes have a high output capacitance which makes difficult their use at v.h.f. with a normal parallel-tuned circuit. This type of valve can be used, however, if a " series-tuned " anode circuit is adopted (Fig. 7.3). In this arrangement the tuning capacitor C1 and the output capacitance of the valve are in series across the inductor. Ideally the value of C1 should be equal to the valve output capacitance. The effective parallel capacitance is less than that due to the valve alone and the external inductance can be increased to a value that results in an efficient circuit.

Linear tuned circuits are most conveniently employed in push-pull stages and may be any multiple of a quarter wavelength ($\lambda/4$) long. This refers to the electrical length of the line, the actual length in practice being somewhat shorter due to the valve acting as an extension of the line, so that the added inductance of the leads and the inter-electrode capacitance result in a physical shortening of the line. A quarter wavelength linear anode circuit for a push-pull stage is shown in Fig. 7.4 (a). Tuning may be effected either by varying the physical length of the line by adjustment of a shorting bar or by a variable capacitor C1 connected across the open circuited end of the line adjacent to the valve. To maintain circuit balance a split-stator capacitor should be used for tuning with the rotor left floating and not connected to earth. For the same reason the h.t. feed connection at the voltage antinode of the line should be left unbypassed. An electrical centre-tap on the circuit is provided by the valve inter-electrode capacitances.

A push-pull anode circuit employing a half wavelength line is shown in Fig. 7.4 (b). The valve and tuning capacitance are connected at opposite ends and h.t. is applied at the electrical centre of the line, i.e. the voltage node. Half wavelength lines are sometimes employed in grid circuits at the higher frequencies where the valve leads and input capacitances are such that the natural frequency of the valve grid assembly is lower than the frequency of operation. All parallel-line circuits should be shielded to minimize loss due to r.f. radiation from the lines which becomes greater as the frequency of operation increases.

TABLE 7.2

Make	Type No.	Class	Base	Cathode			Limiting Values			Max. freq. full ratings (Mc/s)	Output at full ratings (W)	Max. freq. reduced ratings (Mc/s)	Output at reduced ratings (W)
				Type	V	A	P_A	V_A	V_{G2}				
Mullard U.S.A. CV	ECC91 6J6 858	Double-triode	B7G	IH	6·3	0·45	2 × 1·5	300	—	80	3·5	—	1·0 at 250 Mc/s
Mullard	EC56	Disc-seal triode	Octal	IH	6·3	0·65	10	300	—	—	—	4000	0·5
Mullard	EC57	Disc-seal triode	Octal	IH	6·3	0·65	10	300	—	—	—	4000	1·8
M-O Valve Mullard CV	DET22 TD03-10 273	Disc-seal triode	—	IH	6·3	0·4	10	350	—	1000	2·8	3000	0·5
S.T.C. CV	3B/152M 1540	Double-triode	B9G	IH	6·3	0·92	15	375	—	300	28	—	—
M-O Valve Mullard CV	DET24 TD04-20 397	Disc-seal triode	—	IH	6·3	1·0	20	400	—	600	23	2000	3·5
S.T.C. CV	3B/240M 2214	Triode	B8G	IH	6·3	1·1	24	375	—	200	24	—	—
M-O Valve CV	ACT22 257	Disc-seal triode	—	IH	6·3	4·0	75	600	—	1000	90	—	—
Mullard CV	QQV02-6 2466	Double-tetrode	B9A	IH	6·3 12·6	0·8 0·4	2 × 3	275	200	200	6	500	5
Mullard CV	QV04-7 309	Tetrode	B9G	IH	6·3	0·6	7·5	400	250	20	7·9	150	6·3
Mullard U.S.A. CV	QQV03-10 6360 2798	Double-tetrode	B9A	IH	6·3 12·6	0·8 0·4	2 × 5	300	200	100	16	225	12·5
M-O Valve CV	TT15 415, 4046	Double-tetrode	B9G	IH	6·3	1·6	2 × 7·5	400	—	200	12	250	—
Mullard U.S.A. CV	QQV04-15 832A 788	Double-tetrode	B7A-	IH	6·3 12·6	1·6 0·9	2 × 7·5	750	250	100	26	250	18
M-O Valve Mullard U.S.A. CV	TT20 QQV03-20A 6252 2799	Double-tetrode	B7A	IH	6·3 12·6	1·3 0·65	2 × 10	600	250	200	48	600	20
Mullard U.S.A. CV	QV06-20 6146 3523	Tetrode	Octal	IH	6·3	1·25	20	600	250	60	52	175	25
Mullard U.S.A. CV	QQV06-40A 5894 2797	Double-tetrode	B7A	IH	6·3 12·6	1·8 0·9	2 × 10	750	250	200	90	475	60
Mullard U.S.A. CV	QQV07-40 829B 2666	Double-tetrode	B7A	IH	6·3 12·6	2·5 1·25	2 × 20	750	225	100	87	250	60
Mullard U.S.A. CV	QY3-65 4-65A 1905	Tetrode	B7A	DH	6·0	3·5	65	3000	400	50	280	220	110
Mullard U.S.A. CV	QY3-125 4-125A 2130	Tetrode	B5F	DH	5·0	6·5	125	3000	400	120	375	200	225
Mullard S.T.C. U.S.A. CV	QV1-150A 4X150A 4X150A 2519	Tetrode— forced-air cooled	B8F special	IH	6·0	2·6	150	1250	300	165	195	500	140
M-O Valve ST&C U.S.A. CV	4CX250B 4CX250B 4CX250B 2487	Tetrode— forced-air cooled	B8F special	IH	6·0	2·6	250	1500	300	175	235	500	225 (at 2kV)
ST&C U.S.A. CV	2C39A 2C39A 2516	Triode Disc seal	— —	IH	6·3	1	100	1000	—	2400	20	—	—
M-O Valve CV	DET29 2397	Triode Disc seal	—	IH	6·3	0·5	10	450	—	3800	1·5	7000	—

Linear tuned anode circuits can be used in single ended stages as shown in Fig. 7.4 (c). The capacitor C2 is necessary to balance the output capacitance of the valve connected across the other half of the line.

Provision for adequate ventilation of the valve must be made to avoid excessive bulb temperatures being caused by enclosing this type of circuit.

Aerial Coupling to Parallel-line Tank Circuits

The usual method of coupling the output of the v.h.f. transmitter to a low impedance coaxial feeder is by a one or two turn coupling winding located adjacent to the earthy end of the p.a. tank inductor. One side of the coupling circuit is invariably earthed, either directly or through a small preset capacitor which is adjusted to cancel out the inductive reactance of the coupling coil. With a parallel-line tank circuit a hairpin loop is employed, as shown in Fig. 7.4 (b) and (c). This arrangement is satisfactory with the unbalanced circuit of Fig. 7.4 (c) but when used with the push-pull anode circuit of Fig. 7.4 (b) it results in some unbalance being reflected back into the tank circuit.

One method of overcoming this is to use a linear balance-to-unbalance and impedance transformer—balun (Fig. 7.5). The amount of coupling is determined by the distance between the balun and the anode lines, and also by the length of the balun lines which is dependent on the position of the shorting bar.

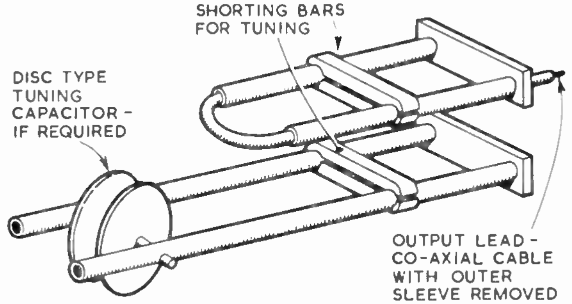


Fig. 7.5. Aerial coupling to a linear (parallel line) tank circuit.

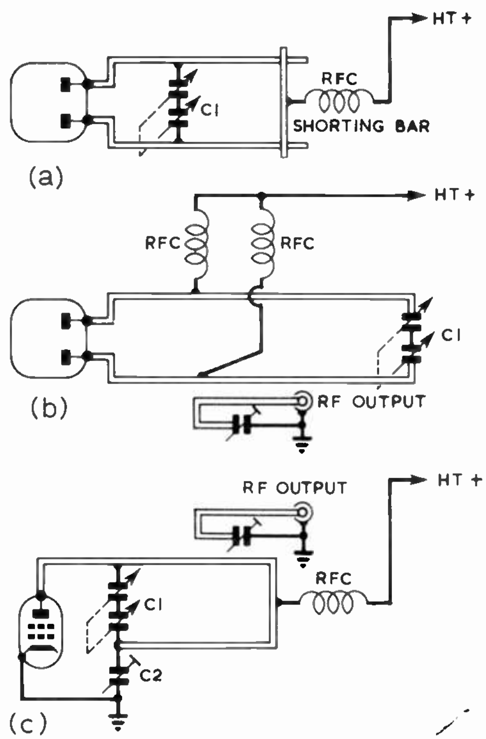


Fig. 7.4. Linear tuned circuits. (a) Quarter wavelength tuned circuit. (b) Half wavelength line for push-pull use. (c) Linear tuned circuit for use with a single ended stage.

Coaxial or Cavity Circuits

Coaxial or cavity circuits are usually employed at 435 Mc/s and above. At 1296 Mc/s planar-electrode valves are incorporated in grounded grid circuits with quarter wavelength cathode and anode cavities, though three-quarter wavelength cavities are sometimes found. Use of a grounded grid amplifier obviates the necessity for neutralization but as the amplifier and its driver stage are effectively connected in series, both stages must be modulated for telephony transmission. The audio power is not wasted, however, as the modulated r.f. output of the driver stages is added to the output of the amplifier.

Design of Transmission Line Resonators

When designing a resonator to be used as a tank circuit it is necessary to know first how long to make the lines. The resonant frequency of a capacitatively loaded shorted line, open-wire or coaxial, is given by the following well-known expression:

$$\frac{1}{2\pi fC} = Z_0 \tan \frac{2\pi l}{\lambda}$$

where F *is the frequency*
C *is the loading capacity*
λ *is the wavelength*
l *is the line length*
Z_0 *is the characteristic impedance of the line.*

The characteristic impedance is given by:

$$Z_0 = 138 \log_{10} \frac{R_1}{R_2}$$

for a coaxial line with inside radius of the outer R_1 and outside radius of the inner conductor R_2.

or $$Z_0 = 276 \log_{10} \frac{2D}{d}$$

for an open-wire line with conductor diameter d and centre-to-centre spacing D.

The results obtained from these expressions have been put into the form of the simple set of curves shown in Fig. 7.6.

In the graphs, Fl has been plotted against FC for different values of Z_0, with F in Mc/s, C in pF and l in centimetres.

In the case of coaxial lines (the righthand set of curves) r is the ratio of conductor diameters or radii and for open-wire lines (the lefthand set of curves) r is the ratio of centre-to-centre spacing to conductor diameter.

The following examples should make the use of the graphs quite clear:

Example 1

How long must a shorted parallel-wire line of conductor diameter 0·3 in. and centre-to-centre spacing 1·5 in. be made to resonate at 435 Mc/s, with an end-loading capacitance of 2 pF (the approximate output capacitance, in practice, of a QQV03-20 (6252) push-pull arrangement)?

First, work out $F \times C$, in Mc/s and pF.

$$FC = 435 \times 2 = 870 = 8{\cdot}7 \times 10^2.$$

The ratio, r, of line spacing to diameter is:

$$r = \frac{1{\cdot}5}{0{\cdot}3} = 5{\cdot}0$$

Then, using the curves marked "parallel-wire lines," $r = 5{\cdot}0$ in. project upwards from $8{\cdot}7 \times 10^2$ on the horizontal "$F \times C$" scale to the graph and project across from the point on the graph so found to the vertical "$F \times l$" scale, obtaining:

$$F\,l = 2800$$

$$\textit{therefore, } l = \frac{2800}{435} = 6{\cdot}45 \text{ cm approximately.}$$

The anode pins would obviously absorb quite a good deal of this line length but, if the lines were made 6 cm long, with an adjustable shorting-bar they would be certain to be long enough.

Example 2

A transmission line consisting of a pair of 10 s.w.g. copper wires spaced 1 in. apart and 10cm long is to be used as part of the anode tank circuit of a QQV06-40 (5894) p.a. at 145 Mc/s. How much extra capacitance must be added at the valve end of the line to accomplish this?

For a pair of wires approximately $\frac{1}{8}$ in. in diameter spaced 1 in. r is about 8. Also $F \times l$ is equal to 145×10, i.e. 1450. Estimating the position of the "$r = 8$" curve for a parallel-wire line between "$r = 10$" and "$r = 7$," $F \times C$ is found to be about $1{\cdot}55 \times 10^3$, i.e. 1550. Hence the total capacity C required is given by:

$$145 \times C = 1550$$
$$C = 1550 \div 145 = 10{\cdot}7 \text{ pF.}$$

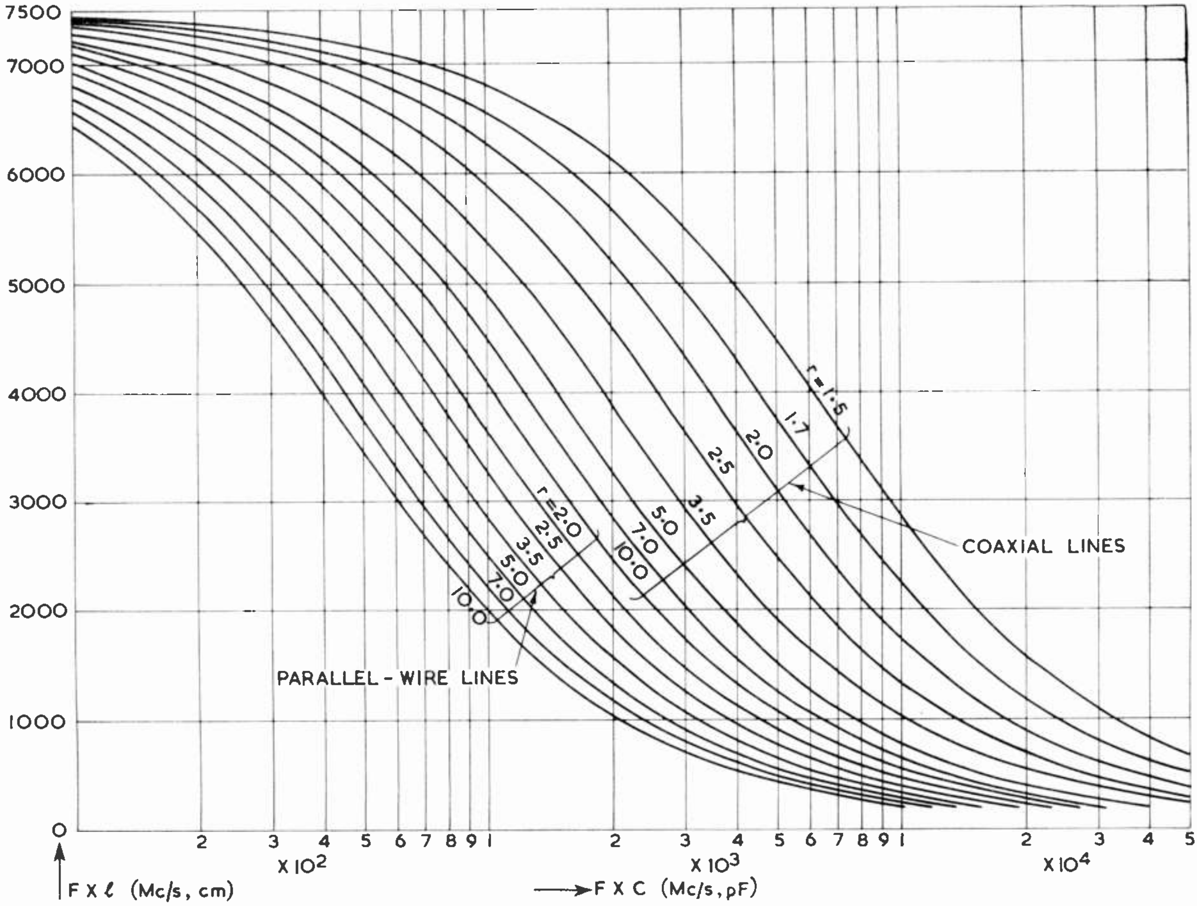


Fig. 7.6. Resonance curve for capacitively loaded transmission line resonators.

Now the output capacitance of a QQV06-40 (5894) push-pull stage is around 4 pF in practice, so about 7 pF is required in addition. A 25 pF + 25 pF split stator capacitor should therefore be quite satisfactory giving 12 to 15 pF extra at maximum capacity.

Example 3

A coaxial line with outer and inner radii of 5·0 and 2·0cm, respectively, is to be used as the resonant tank circuit (short-circuited at one end of course) for a 4X150A power amplifier on the 70cm amateur band. What length of line is required?

In this case:

$$F \times C = 435 \times 4{\cdot}6 = 2001.$$

Using the "$r = 2{\cdot}5$" curve for coaxial lines,

$$F \times l = 4620$$

Hence

$$l = 4620 \div 435 = 10{\cdot}6\text{cm approximately.}$$

This length would include the length of the anode and cooler of the 4X150A of course but, as in Example 1, a line 10cm long would be certain to be long enough, especially as the output capacity used in the calculations is that quoted by the manufacturers for the valve, the effective capacity being somewhat greater in practical circuits. A shorting bridge would be the best method of tuning the line to resonance.

Designing for Maximum Unloaded Q

The tank circuit efficiency is given by:

$$\textit{Efficiency}\text{ (per cent)} = \frac{\textit{unloaded } Q - \textit{loaded } Q}{\textit{unloaded } Q} \times 100$$

It is obvious that the highest possible unloaded Q is needed to get the greatest tank circuit efficiency. The Q is greater for radial and coaxial resonators than for comparable parallel-wire circuits and the former types should always be used where possible. It should perhaps be explained that unloaded Q is the Q of the tank circuit with the valve in position and all voltages and drive power applied, but with no load coupled up to it. The loaded Q is, of course, that measured when the load is correctly coupled to the tank circuit.

For unshielded parallel-wire lines, the unloaded Q is usually quite low because of power loss by radiation from the line and the best value is obtained by using a small conductor spacing (low Zo).

To obtain the best Q, the material of the line should be copper or brass fairly smoothly finished, although a highly polished surface is not necessary. To improve its conductivity, the surface of a coaxial or radial line can be silver plated. In an industrial or city atmosphere, however, the silver plating is rapidly attacked by atmospheric gases and the surface conductivity suffers far more than does that of a copper or brass line. The best solution is to apply a "flash" of rhodium to the silver plating but this is rarely possible for the amateur. The Q of a coaxial line depends also upon the ratio of conductor diameters and, for fairly heavily capacitatively loaded resonators, which is true in most practical cases, this ratio should be between about 3 and 4·5 to 1.

Care should be taken that the moving contacts on the bridge (if one is used) are irreproachable: they should preferably make contact with the line a little way from the shorting disc, where the line impedance is somewhat higher than at the current antinode. It is better to use a large number of springy contact fingers rather than to use a few relatively rigid ones.

Attention to these points will often make all the difference between a reliable, satisfactory resonant circuit of high Q and one which possesses none of these qualities.

Disc-Type Capacities

Parallel lines or concentric (coaxial) tuned circuits are conveniently tuned by means of a variable air capacitor comprising two parallel discs, but the calculation of the capacitance range of different size discs with differing spacing can be tiresome.

The chart in Fig. 7.7 gives the capacitance between two parallel discs of various convenient diameters with spacings

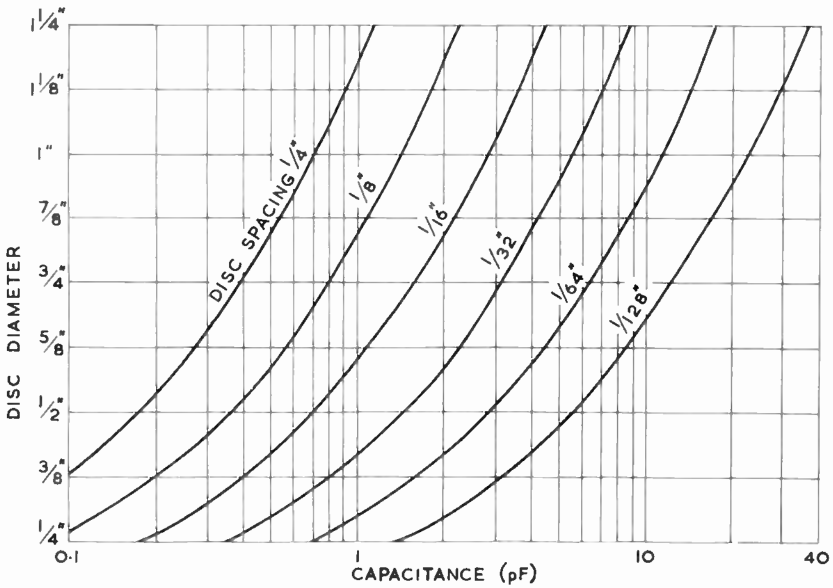


Fig. 7.7. Capacitance between two parallel discs of various diameters.

between ¼ in. and 1/128 in. calculated according to the formula

$$CpF = \frac{0{\cdot}244 \times \textit{area (inches)}}{\textit{spacing (inches)}}$$

The diameter of the disc employed may be fixed by space considerations but it determines the minimum and maximum capacitance and the range available. Very close spacings should be avoided unless extremely accurate parallelism can be maintained, and must be avoided where high voltages exist, as in transmitters.

For example, a cathode concentric tuned circuit tuned by a disc capacitor is required for a 2C39A valve to be used as a tripler from 432 to 1296 Mc/s. This valve has an input capacitance of 6·5 ± 1 pF. If the cavity has an outer diameter of 2⅜ in. and an inner conductor of ⅜ in. the ratio is 7 : 1 and at 430 Mc/s the curves in **Fig. 7.6** show that if C is taken as 7·5 pF plus a minimum disc capacitance of, say 0·5 pF, i.e. 8 pF, the inner conductor length is 4·2cm or 1·65 in. If a ½ in. disc capacitor is chosen then the range of tuning is from 0·2 pF at ¼ in. spacing to 1·5 pF at 1/32 in. spacing. As the drive voltage is of the order of 100 volts closer spacing must not be used. It is clear that a ½ in. disc will not do as the spread between top and bottom limit values is 2 pF so that ¾ in. discs will have to be used giving a range of 0·4 to 3·0 pF which is sufficient for fixed frequency working; if it is necessary to cover a frequency band then a greater range is required. If at times the drive frequency were 420 Mc/s then this would require a maximum capacitance across the tuned circuit of the original 8 pF multiplied by the square of the frequency ratio, i.e. $\frac{(430)^2}{(420)^2}$ or 1·05. This means a maximum disc capacitance of 2·4 pF and the ¾ in. diameter will still just do but ⅝ in. would not.

If it is found that the range of the capacitance required is larger than that given by the maximum size disc that can be accommodated, then the solution is to use two sets of discs located, for example, each side of a cavity and use one as tuning and one as a kind of band-set. Valve capacitance limits must be taken into account otherwise the replacement of a valve can result in the circuit no longer tuning. Where the tolerances in input and output capacitance are not known they can be taken as ± 30 per cent which would adequately cover most cases.

Neutralization

With triode valves and tetrodes operating below their self-resonant frequency the normal methods of neutralization apply i.e. the application of an antiphase feedback voltage from anode to grid of the valve. The residual anode-to-grid capacitance of modern v.h.f. tetrodes is quite low, in some instances being less than 0·1 pF. Provided that adequate screening exists between grid and anode circuits, operation of certain types of valve in the lower part of the v.h.f. spectrum (up to 70 Mc/s) is permissible without neutralization being required. At higher frequencies although the reactance of this capacitance is still not sufficiently low to cause oscillation, it will result in the stage being regenerative and neutralization will be necessary.

A convenient method of providing the small capacitance required for neutralizing push-pull double tetrode stages

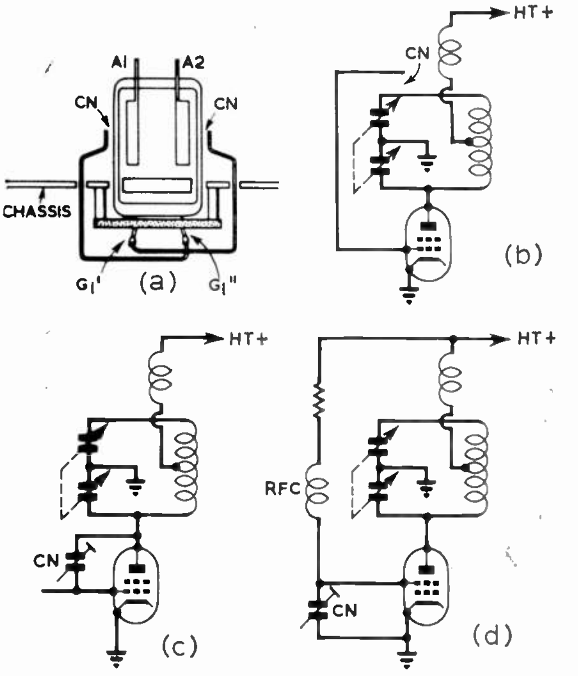


Fig. 7.8. Neutralizing circuits. (a) Neutralization of push-pull double tetrode stage. (b) Similar method for single ended stage. (c) Neutralization of a valve operating above the self neutralizing frequency. (d) Screen neutralization of single ended stage.

consists in using each valve anode as one plate of the capacitor, and a wire connected to the grid of the opposite valve as the other plate (**Fig. 7.8 (a)**). The capacitance (C_n) is adjusted by bending each wire nearer or further away from the respective anode. In single ended circuits the capacitance is formed by a wire connected at one end to the grid and positioned so that the other end is adjacent to the appropriate side of the anode tank circuit (**Fig. 7.8 (b)**). Heavy gauge enamelled copper wire (12 to 16 s.w.g.) should be used, and a metal tab may be attached at the end of the wire to increase the capacitance should that provided by the wire alone be insufficient.

When the valve is operated above the self-neutralizing frequency the neutralizing capacitor must be connected directly between anode and grid of the valve (**Fig. 7.8 (c)**). Alternatively a variable capacitor may be connected between screen grid and earth and tuned with the screen lead inductance to form a series resonant circuit at the frequency of operation, thus providing a low screen-to-earth impedance and effectively placing the screen grid at earth potential with respect to r.f. (**Fig. 7.8 (d)**). It must be remembered that this method of neutralization is frequency-sensitive, and any considerable change in the operating frequency will necessitate re-adjustment of the variable capacitor.

In some double tetrode valves such as the QQV03-20A (6252) and QQV06-40 (5894) internal neutralizing is incorporated in the valve structure so that external neutralizing is unnecessary provided the circuit design and layout is good.

Further information on neutralizing procedures is given in Chapter 6.—*H.F. Transmitters*.

MODULATION

For telephony service a v.h.f. transmitter may be modulated by any one of the normal methods, though high level modulation of the final amplifier is the system usually adopted. For satisfactory modulation it is essential that there is adequate grid drive to the modulated amplifier and that the stage is correctly neutralized.

Experience has shown that it is desirable to maintain a high average level of modulation in a v.h.f. transmitter and the use of speech clipping or compression is therefore recommended (see Chapter 9—*Modulation*).

Instability in the modulator may be experienced due to r.f. feeding back into early stages of the speech amplifier. To eliminate this the microphone input circuit should be completely shielded and the effect of inserting grid stopper resistors, or ferrite beads threaded on to the grid lead close to the valve socket connection, particularly in low level speech amplifying stages, and the inclusion of an r.f. filter in the grid circuit of the first valve should be investigated.

SEMICONDUCTORS IN V.H.F. AND U.H.F. TRANSMITTERS

Transistors and Varactor diodes capable of handling reasonably high power—a few tens of watts—are available and make attractive designs possible. They compete strongly with some of the special valves designed for the v.h.f. and u.h.f. range, especially in portable and mobile equipment.

Different problems, however, arise, such as adequately stabilized low voltage, high current power supplies, transient voltage and current overload protection, provision of suitable heat sinks for cooling, and the use of much lower impedance circuits. None of these problems is particularly difficult to resolve but it is important to recognize that the methods differ considerably from those used with valves.

Valves are relatively high impedance devices and capable of withstanding considerable—often quite severe—overloads for short periods without suffering permanent damage. Semiconductors can be damaged much more easily and wherever possible fast acting protective arrangements should be provided.

Another problem is that there is much less direct interchangeability between different types of semiconductors. This is also true of nominally the same types made by different manufacturers.

It is desirable to start with the highest possible oscillator frequency in order to minimize the number of multiplier stages required to reach the final frequency. It is good practice to use low power stages in the frequency multiplier chain and to use amplifiers to raise the power to the required output level. The gain per stage in semiconductor circuits is generally lower than in comparable valve arrangements and more stages are usually required for a given power level.

Modulation of transistors is normally quite straightforward but it should be remembered that under a.m. conditions the device must be drastically de-rated to avoid damage. For this reason, it is advantageous to use frequency or phase modulation whenever possible, a method which also reduces the problem of the dynamic performance of the power supply.

Amplitude modulation should be applied to the driver stage as well as the final power amplifier in transistor transmitters. In some cases, in which three or four amplifiers are used, it may be beneficial to apply modulation to the last three stages.

Typical transistor amplifier circuits are shown in **Fig. 7.9.**

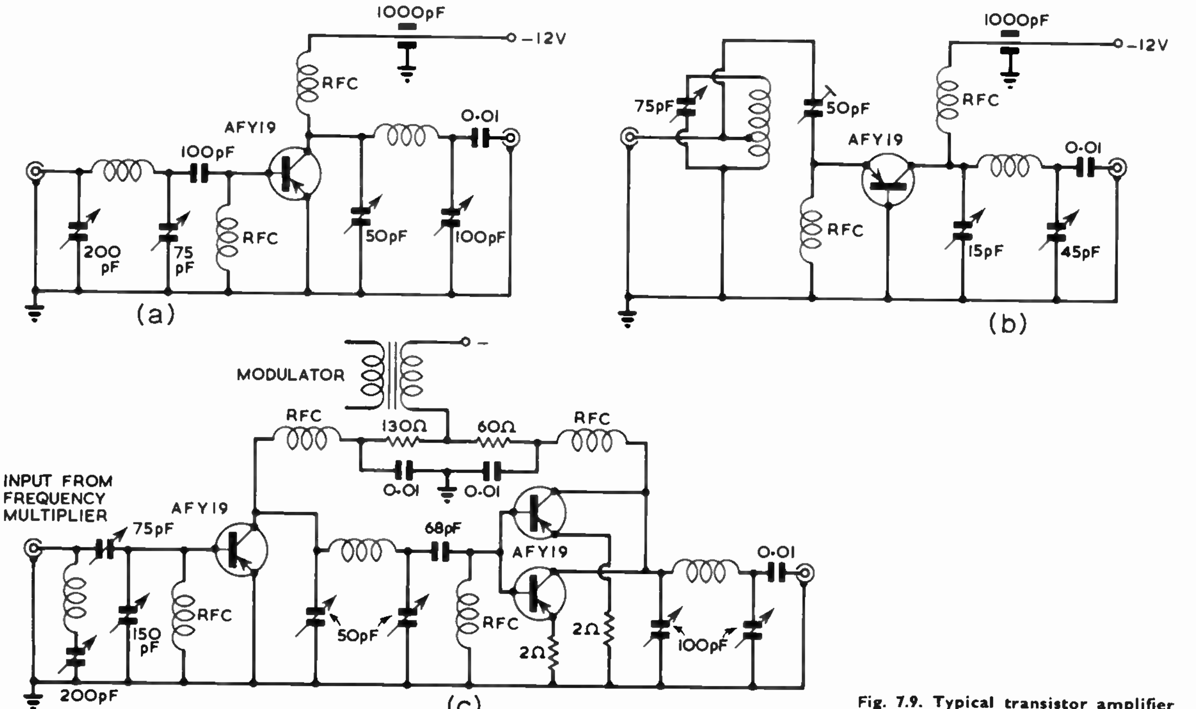


Fig. 7.9. Typical transistor amplifier circuits.

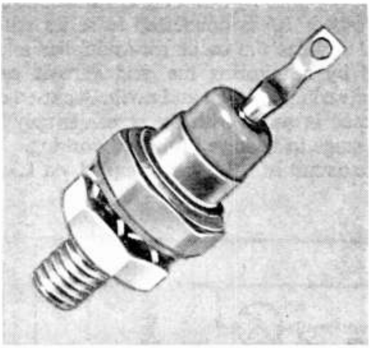
Varactors

Varactor diodes are available for frequency doubler or tripler service at quite high power levels and efficiency of more than 50 per cent. Circuits employing such diodes are entirely r.f. powered, thus providing a simple means of converting the output of a 144 Mc/s transmitter to 432 Mc/s. However, unless a valve or transistor p.a. is used, it is necessary to employ frequency or phase modulation.

While simple tuned circuits may be used with Varactors, both input and output frequencies will appear in the output circuit. It is essential therefore to employ an adequate filter to eliminate all but the wanted frequency.

Practical Varactor Doubler and Tripler Circuits

The most useful application of Varactor diodes in amateur equipment is for frequency multiplication in the range 70 to 1300 Mc/s to drive transistor or valve power amplifiers which can be amplitude modulated.



BAY96 Varactor Diode

In the circuit arrangements shown in **Fig. 7.10**, it is necessary to match the input source and the output load (50 or 75 ohms) to the Varactor diode which has a very low impedance. The circuits are usually of low loaded *Q*, a figure of about 10 being a good compromise between circuit losses and reduction of unwanted harmonics. In the unloaded condition, these circuits should be of high *Q*.

The Varactor diode may be used either in series or shunt connection. The latter is particularly convenient because one side of the diode may be connected directly to the chassis which provides a convenient heat sink. Bias can be readily provided by a suitable resistor connected across the diode. The value of the resistor should be low enough to prevent instability and 15 K to 100 K ohms will generally be found suitable. The higher the drive power the lower will be the value of the resistor required.

Varactors may be operated in parallel for higher power output, the most suitable being the Mullard BAY66 and BAY96 and the Amperex H4A. The BAY66 is the lowest power but is more suitable than the other for operation above 1000 Mc/s.

Typical performance of these Varactors is as follows:

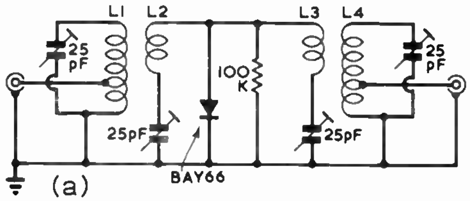
BAY66: An input of 10 watts at 72 Mc/s will give 7·5 watts output at 144 Mc/s. An input of 10 watts at 144 Mc/s will give 6·5 watts output at 432 Mc/s.

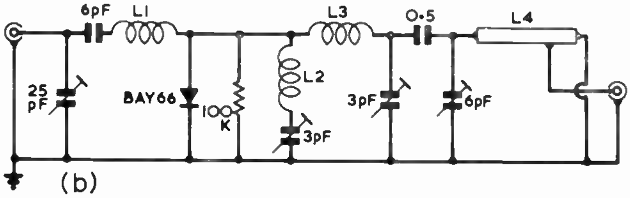
BAY96 and H4A: An input of 25 watts at 144 Mc/s will give an output of 17 watts at 432 Mc/s.

FOUR METRE TRANSMITTERS

For the newcomer to the v.h.f. bands 70 Mc/s offers many interesting possibilities. It is the lowest v.h.f. band allocated to British amateurs and is sufficiently close to that of the high frequency 28 Mc/s band to warrant no great change in transmitter design and construction techniques; in some respects they are simpler because crystal controlled oscillators are usually employed on 70 Mc/s, whereas h.f. transmitters invariably require a v.f.o. It is thus possible for a high efficiency 70 Mc/s transmitter to be built without having to use the more expensive valves and components necessary to achieve optimum results on the other v.h.f. and u.h.f. bands. Provided that short, low impedance connections are made in r.f. circuits, and this is probably the most important factor in v.h.f. transmitter construction, a successful 70 Mc/s transmitter is assured.

From an operating point of view the band has distinct attractions. For local communication the results are similar to those obtained on the 2m band and excellent results can be obtained with transmitters having a power input of 10 to 15 watts. For ranges in excess of say, 100 miles, under poor or average conditions it appears that the transmission path on 4m is better than that obtained on 2m. In addition, the 4m band has the distinct advantage that DX working, up to distances exceeding 1000 miles, is possible under certain conditions by virtue of tropospheric refraction.





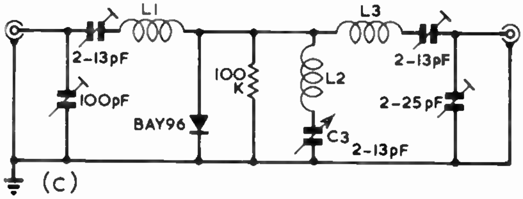


Fig. 7.10. Practical circuits using shunt connected BAY66 and BAY96 varactors.

TABLE 7.3

Crystal frequencies and multiplication factors for 70 Mc/s transmitters.

Crystal Frequency	Multiplication
5·8416 to 5·8916 Mc/s	× 3 × 2 × 2
7·788 to 7·855 Mc/s	× 3 × 3
8·7625 to 8·8375 Mc/s	× 2 × 2 × 2
11·683 to 11·783 Mc/s	× 3 × 2

A range of suitable crystal frequencies and the multiplying factors necessary to give a final frequency between 70·1 and 70·7 Mc/s is listed in Table 7.3.

A 50 WATT PHONE AND C.W. TRANSMITTER

The three stage transmitter shown in **Figs. 7.11 to 7.13** consist of a crystal oscillator multiplier followed by a second frequency multiplier and power amplifier. The final stage, which may be run at the maximum permissible input of 50 watts on both phone and c.w., employs a Mullard QV06-20 (6146) v.h.f. power tetrode which is suitable for operation at frequencies up to 175 Mc/s and makes an economical and efficient power amplifier for a 4m transmitter. At 70 Mc/s an efficiency of approximately 75 per cent may be obtained.

Provision is made for keying the transmitter in the h.t. lead to the screen of the multiplier valve. Protection of the p.a. stage during " key up " periods is achieved by a clamp valve which reduces the screen voltage and thus limits the anode current of this stage. Anode and screen modulation of the final amplifier may be effected by an external modulator with an audio output power of 25 to 30 watts, which can be obtained, for example, from two 25W audio tetrodes operating in class AB1. H.T. and heater supplies of 400 volts at 175 mA. and 6·3 volts at 2·65 amps respectively are required to operate the transmitter.

Circuit

The circuit diagram is shown in **Fig. 7.11.** An EL91 (6AM5) a.f. pentode (V1) functions as a crystal controlled oscillator-multiplier in a modified Pierce circuit; the crystal frequency is in the 7·8 Mc/s range, and the multiplier anode circuit is tuned to the third harmonic, i.e. approximately 23·4 Mc/s. R.F. output from V1 is capacitively coupled to the grid of V2, which is a QV03-12 (5763) tetrode also operating as a frequency tripler, the anode circuit L2, C9 being tuned to the final radiated frequency between 70·1 and 70·7 Mc/s, dependent on the actual frequency of the crystal used. Grid bias for V2 is obtained by flow of grid current through R4: a low value resistor R5 is connected in series with the grid resistor, and enables the oscillator-multiplier anode circuit to be tuned by monitoring the grid current on a meter connected between point TP1 and earth. A closed circuit jack J1 is connected in series with the screen supply to V2 and enables this stage to be keyed for c.w. working.

The anode circuit of V2 is link coupled via L3 and L4 to

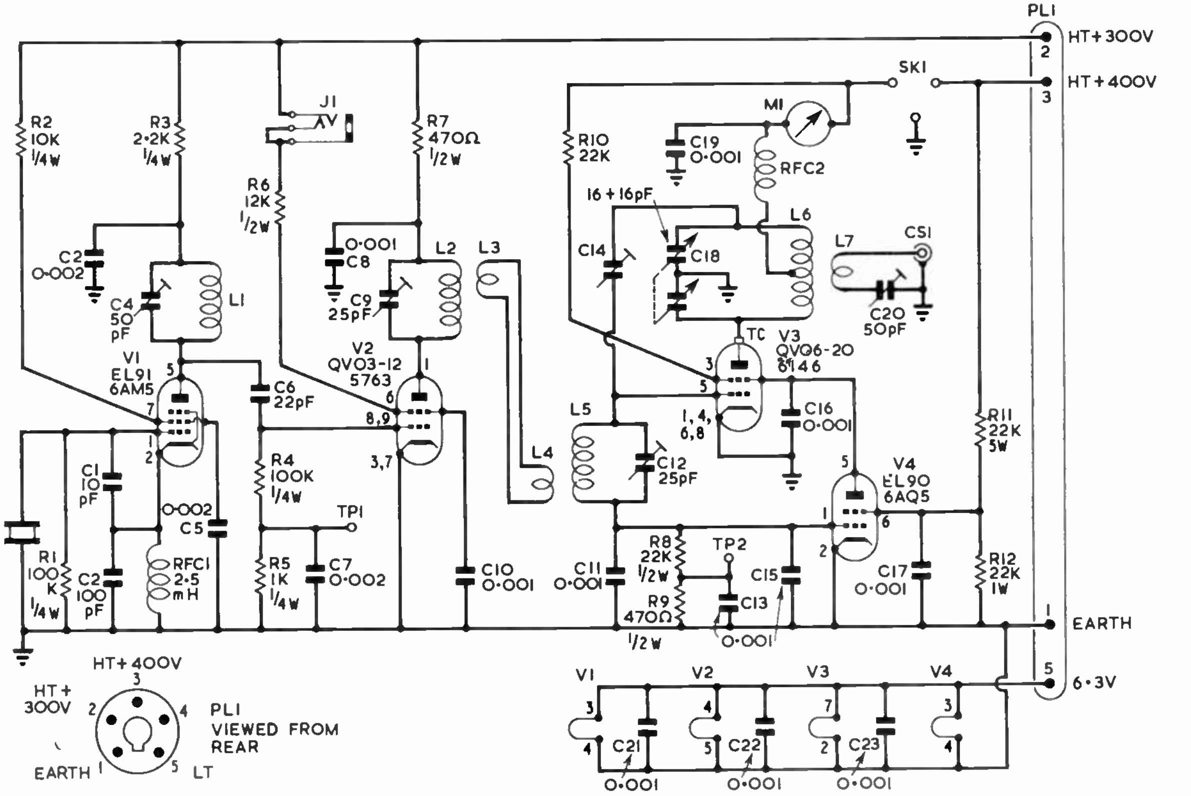
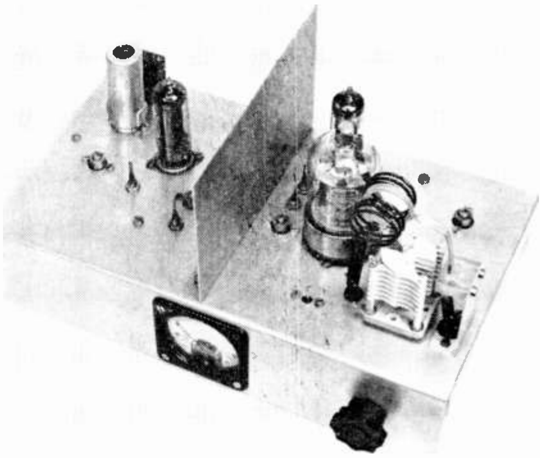


Fig. 7.11. Circuit diagram of the 70 Mc/s transmitter. C7, 8, 10, 11, 13, ceramic feedthroughs; C14, see text; C15, 16, 17, 21, 22, 23, 0·001 μF ceramic; C19, 0·001 μF 1250V d.c. wkg.; L1, 9 turns 20 s.w.g. enam. $\frac{3}{4}$ in. o.d. $\frac{5}{8}$ in. long; L2, L5, 3 turns 16 s.w.g. enam $\frac{3}{8}$ in. o.d. $\frac{1}{2}$ in. long; L3, L4, 1 turn 18 s.w.g. tinned copper $\frac{3}{4}$ in. o.d. insulated with polythene sleeving; L6, 6 turns 16 s.w.g. enam. 1 in. o.d. 1 in. long $\frac{3}{8}$ in. gap at centre for L7; L7, 2 turns 18 s.w.g. tinned copper 1 in. o.d. insulated with polythene sleeving; M1, 0–150 mA d.c. meter; RFC1, 2·5 mH (Eddystone No. 737); RFC2, 50 turns 28 s.w.g. enam. $\frac{1}{4}$ in. diam. $1\frac{1}{4}$ in. long wound on Eddystone former type 685.



A 50 watt phone and c.w. transmitter for 70 Mc/s.

the grid circuit L5, C12 of the tetrode valve V3 QV06-20 (6146). This stage functions as a straight amplifier; r.f. output from the anode circuit L6, parallel-tuned by a split-stator capacitor C18, is inductively coupled through a low impedance loop L7 to the coaxial output socket CS1; C20 is connected in series with the output circuit, and in operation its capacitance is so adjusted that the reactance cancels out the inductive reactance of L7.

A split-stator tuning capacitor C18 is used in the anode circuit of V3 so that an r.f. voltage in antiphase to that at the anode is available to be applied to the grid through C14 to effect neutralization of the p.a. stage. The grid anode capacitance of V3 is low enough for the stage to be stable. It may be operated without neutralization at 70 Mc/s but the inclusion of a neutralizing capacitance is recommended so that optimum working conditions may be obtained. The capacitance required in practice is very small and C14 is formed by a wire connected at one end to the grid of V3, the other end being adjacent to the stator plates of C18 remote from the anode of V3.

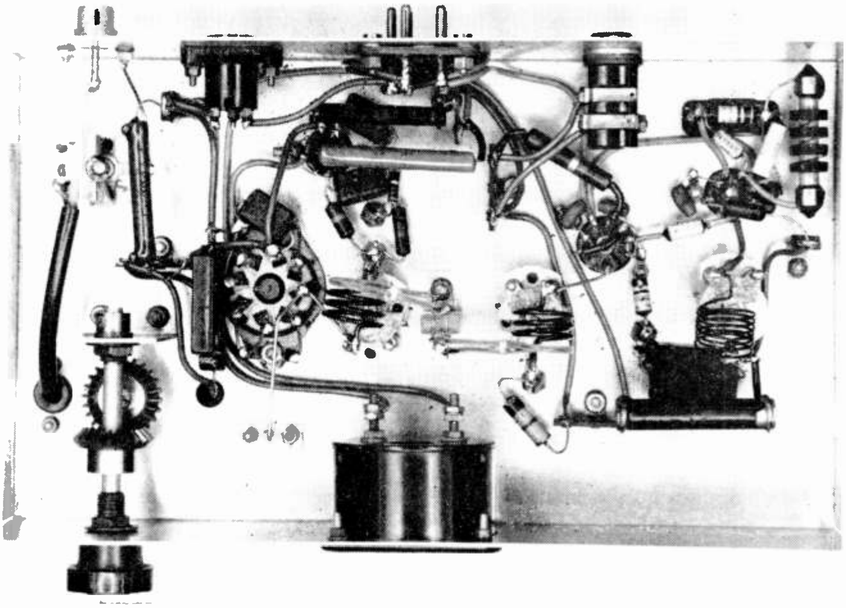
Grid leak bias only is also applied to the p.a. valve and is obtained by the grid current through R8 and the monitor resistor R9 in series; a milliammeter may be connected between test point 2 and chassis to give an indication of resonance of the anode circuit of V2. An EL90 (6AQ5) a.f. pentode (V4) is connected between the screen grid of V3 and earth; the grid of V4 is decoupled by C15 and is connected to the grid return circuit of V3 so that when V3 is being driven the negative bias voltage developed across R8 and R9 in series is applied to the grid of V4. This voltage, which is approximately 90, biases V4 well beyond cut-off and the valve then has no effect on the operation of the circuit.

When the drive to V3 is removed, the negative bias applied to the grids of both V3 and V4 is removed also: the clamp valve conducts heavily, resulting in an increased voltage drop across R10 and a consequent reduction in the screen voltage to a very low value. This reduces the anode current of V3 sufficiently to prevent the maximum permissible dissipation (20 watts) of the valve being exceeded. With 400 volts h.t. applied this is represented by an anode current of 50 mA; in practice the anode current falls to less than 25 mA when the drive is removed.

Anode and screen modulation of the 400 volts h.t. supply to the p.a. is effected by connecting the output of the modulator to SK1. The anode current taken by V3 is continuously monitored by M1.

An h.t. supply line of 300 volts is required for the crystal oscillator and frequency multiplier stages. This may be obtained from a separate power unit capable of delivering approximately 50 mA at 300 volts, or alternatively from the main 400 volts line through a 2K ohm 5 watt series resistor. The heaters of V1, V2 and V3 are decoupled by C21, C22 and C23 respectively, connected directly between the heater tag on each valveholder and chassis.

A view under the chassis of the 70 Mc/s transmitter.



Construction Details

The transmitter is assembled on a chassis measuring 10¾ in. long × 6 in. wide × 2½ in. deep fabricated from 18 s.w.g. aluminium. The layout may clearly be seen in the photographs.

The crystal and the oscillator-multiplier valve V1 are located at the left-hand side towards the rear, with the anode tuning capacitor C4 immediately in front of the valve. Nylon-loaded p.f. (phenolformaldehyde) valveholders are used for this stage and for the frequency multiplier V2. It is essential that a screening can is not fitted to V2 because this would restrict the dissipation of heat from the valve. As the multiplier is unscreened, it is necessary to fit a shield on top of the chassis between this stage and the power amplifier; the shield which is located at the middle and runs from front to rear of the chassis, measures 5¾ in. long × 3½ in. high and is made from 20 s.w.g. aluminium.

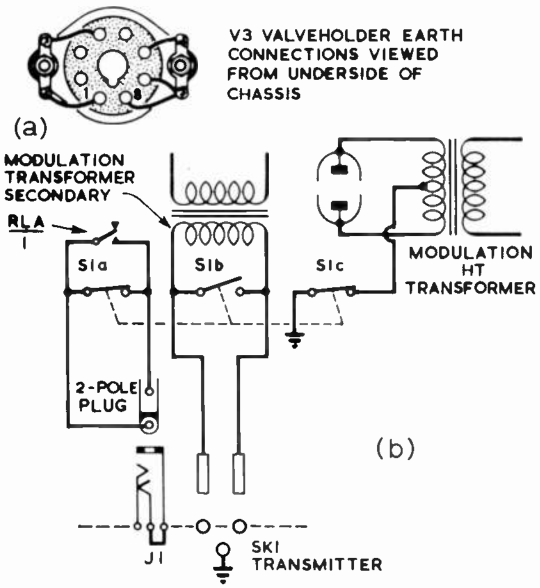


Fig. 7.12. (a) P.A. valveholder earthing. (b) Phone/c.w. switching.

The feedthrough bypass capacitors C7 and C13 enable the grid currents of V2 and V3 to be monitored conveniently from the top side of the chassis. All inductors are self-supporting and are soldered directly across the associated tuning capacitors. If desired the multiplier anode inductor L2 can be made more rigid by cementing two or three thin strips of polystyrene or Perspex along the axis of the inductor on the inside.

The two coupling loops L3 and L4 are supported by a small insulator measuring 1½ in. × ½ in. made from ⅛ in. polystyrene, which carries two ½ in. lengths of 18 s.w.g. tinned copper wire pushed through two holes spaced $\frac{3}{16}$ in. apart near to one end of the block. Each single turn loop is insulated with polythene sleeving and soldered to opposite ends of the pair of wires; the loops are pushed between the turns of L2 and L4 adjacent to the earthy end of each inductor.

A ceramic valveholder is employed for the final stage and is secured to the *underside* of the chassis by its mounting plate. The base of the QV06-20 (6146) valve carries three cathode pins, and each of the relevant tags on the valveholder, together with the base shield and heater return connections are wired directly to the chassis with the shortest possible leads; a separate connection is made between each of the tags and chassis (**Fig. 7.12 (a)**). A heavy gauge conductor should be used for these connections; in the transmitter illustrated 18 s.w.g. tinned copper wire is employed. A woodflour-loaded p.f. valveholder is used for V4 as no r.f. losses can be incurred in the circuit associated with this valve.

The anode tuning capacitor C18 for the final amplifier is mounted vertically above the chassis with the rotor spindle projecting through to the underside. This enables a relatively short connection, made of thin copper strip ⅛ in. wide, to be obtained between the valve anode and the p.a. tank circuit. An additional nut is screwed on to the threaded rotor bush before the capacitor is mounted so that the ceramic end plate is raised $\frac{3}{16}$ in. above the chassis; this is necessary to avoid the possibility of a short circuit between the stator screws of the capacitor and chassis.

The neutralizing capacitor C14 is formed by two 18 s.w.g. wires, one connected to the grid tag of the p.a. valveholder, and projecting through a hole in the chassis immediately in front of the valveholder; this hole is lined with a polystyrene bush to prevent a short circuit and minimize r.f. loss. The other wire is connected to a 6B.A. soldering tag inserted under the stator column of the tuning capacitor in place of one of the existing two washers.

The tuning control for the p.a. tank capacitor is mounted on the front of the chassis and a right-angle drive is obtained through two 1 in. brass bevel gears. The shaft is of ¼ in. diameter mild steel and measures 3¼ in. long; the rear bush is carried by an 18 s.w.g. aluminium bracket located approximately 2½ in. from the front of the chassis. The two bevel gears should be tightly meshed to avoid any backlash between the control knob and rotor of the capacitor.

The output coupling loop L7 is supported by a block of ¼ in. polystyrene mounted on an aluminium bracket. The turns of the loop are insulated with polythene sleeving, the ends being inserted through holes in the block and cemented in position. A short length of 75 ohm coaxial cable is used to connect the loop to the r.f. output socket and the capacitor C20. An alternative p.a. arrangement is shown in **Fig. 7.13.**

Alignment

During initial alignment h.t. voltage should be applied to each stage successively as the preceding one is tuned. V2 and V3 may conveniently be made inoperative by inserting a plug in J1 and leaving SK1 open circuited. The lining-up procedure is then as follows:

(a) Connect a low range (0-5) milliammeter between TP1 and chassis. Adjust C4 for maximum meter reading.

(b) Connect the milliammeter between TP2 and chassis. Apply h.t. to the screen of V2 and adjust C9 for maximum meter reading. Adjust the position of L3 and L4 to obtain optimum grid current in V3.

(c) Connect a dummy load (e.g. a 230 volt 40 watt lamp) to CS1, apply h.t. to V3 and tune C18 to resonance, observed as a sharp dip in the anode current reading of M1. Adjust the neutralizing capacitance C14 by bending the two wires towards or away from each other until maximum r.f. output

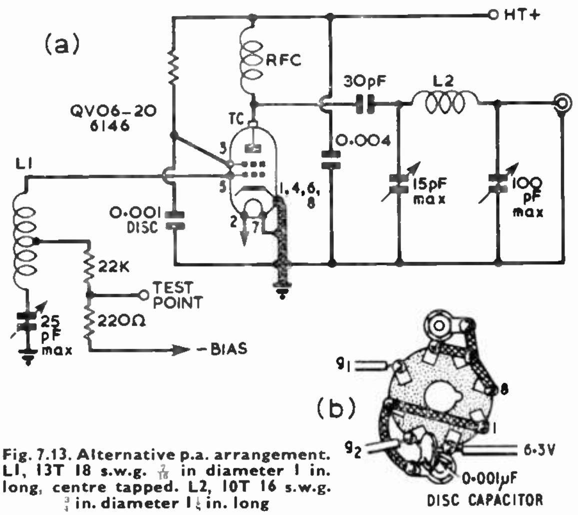


Fig. 7.13. Alternative p.a. arrangement. L1, 13T 18 s.w.g. 5/16 in diameter 1 in. long, centre tapped. L2, 10T 16 s.w.g. 3/8 in. diameter 1¼ in. long

corresponds to minimum anode current. Care should be taken when moving the wire connected to h.t. Adjust the coupling between L6 and L7 and the value of C20 until an anode current of 125 mA, corresponding to an input of 50 watts, is obtained.

(A very convenient method of "hot" neutralization of an r.f. amplifier, i.e. neutralization with h.t. applied to the stage, consists in observing the change in grid current as the anode tuning capacitor is detuned each side of resonance. When the stage is correctly neutralized the condition of minimum anode current at resonance corresponds to maximum grid current. If the grid current falls as the anode circuit is detuned the stage is perfectly balanced; if the grid current rises at either side of the resonant point, the neutralizing capacitor should be adjusted until the correct condition is achieved. It will be found that very accurate neutralization of the p.a. stage in this transmitter can be obtained.)

(d) Check the operation of the clamp valve by removing h.t. from V1 and V2 or inserting a plug into J1. The anode current of V3 should fall to approximately 25 mA.

Operation

When operating on c.w. the transmitter can be run at the maximum permitted input of 50 watts to the final stage (125 mA at 400 volts). *The screen circuit of V2 should be keyed through a relay operated from a low voltage supply to prevent h.t. voltage appearing directly on the key contacts.*

For telephony operation the output of a 25 watt modulator matched into 3225 ohms is connected to SK1, and the keying plug either removed from J1 or short circuited by a switch. The anode current of V3 should be reduced to a value not

TABLE 7.4

Typical current readings for the 4m transmitter

Valve	Grid current	Screen current	Anode current
V1	—	2·5 mA	15 mA
V2	1 mA	6 mA	26 mA
V3	4 mA	12 mA	112 mA

greater than 112 mA at 400 volts (45 watts) by adjusting the position of L7. If this is not done, the rating of the valve at this frequency will be exceeded.

A simple switching system for changing over from c.w. to telephony operation is shown in **Fig. 7.12 (b)**. The contacts of the switch S1 are shown in the telephony position. The use of the relay RLA avoids operation with h.t. being applied to the key contacts. This method can also be used with other transmitters.

TRANSISTOR LOW POWER TRANSMITTER FOR 70 Mc/s

The transmitter shown in Fig. **7.14** was designed around the Texas Instruments Type 2G110 as a power amplifier. This is a germanium *p-n-p* diffused-alloy mesa transistor having a *ft* of 350 Mc/s. It is rated at a maximum collector dissipation of 250 mW at 25° C. ambient temperature and as an r.f. amplifier delivers an output of 100 mW at 100 Mc/s with an efficiency of 35 per cent. In the present application the transistor is operated at an input of 250 mW (18 mA approximately at 15 volts).

The crystal oscillator and frequency multiplying stages of the transmitter employ v.h.f. receiving-type transistors. For the oscillator and first multiplier either Texas 2G401 or Mullard OC170 transistors are suitable. The second multiplier uses a Texas 2G402 or Mullard OC171.

Another type of transistor suitable for the power amplifier stage of a v.h.f. transmitter is the Texas 2S131. This is a silicon *n-p-n* epitaxial transistor having a collector dissipation of 300 mW at an ambient temperature of 25° C. This transistor will deliver an r.f. output power of 150 mW at 100 Mc/s, with a positive collector supply of 18 volts. The Texas 2N715 and 2N716 are rated for operation up to 200 Mc/s, and will give an output at 70 Mc/s of 400 and 600 mW respectively.

Circuit Description

The circuit diagram is shown in Fig. **7.14.** TR1 functions as a crystal oscillator; feedback at the resonant frequency of the crystal is applied between collector and base of the transistor. The collector tuned circuit L1, C1, C2 resonates at the fundamental crystal frequency, which is between 7·788 and 7·855 Mc/s to produce final frequencies from 70·1 to 70·7 Mc/s.

Positive emitter bias to ensure the oscillator starting is provided by connecting the base to a negative potential derived from R1 and R2 across the d.c. supply. R3, in conjunction with R1 and R2, provides d.c. stabilization of the transistor collector current.

The r.f. output from TR1 is applied to the base of the grounded emitter frequency tripler TR2 via L2. L1, L2, form a step-down transformer to match the impedance of TR1 collector circuit to the lower impedance of TR2 base circuit. The third harmonic of the crystal frequency appearing across L3, C5 is similarly applied to the second frequency tripler TR3. Emitter resistors R4 and R5 provide d.c. stabilization for the frequency multiplying stages; the value of these resistors is selected during initial alignments to give maximum r.f. output within the limits of the collector dissipation of the transistors.

The output from TR3 at 70 Mc/s is coupled to the grounded-base amplifier TR4 via L5, C8 and L6, C11. Although the input impedance of TR4 is 75 ohms only, it has been found that greater drive to the emitter results when L6 is tuned. TR4 operates in class B, with a slight forward emitter bias of the order of 0·1 volt, obtained from the junction of R6 and R7, R6 being adjusted to give maximum r.f. output from the stage. The collector circuit of the amplifier is coupled to a 75 ohm unbalanced load through the pi-network C13, L7, C14.

For c.w. working the collector supply to TR3 is keyed. TR2 is thus operating continuously which maintains a constant load on the oscillator. In this service the secondary of the modulation transformer is short-circuited by a contact S1a of a d.p.c.o. switch. For phone operation S1a is changed over, which removes the short-circuit from the modulation transformer secondary in the collector feed to TR4, and short-circuits the key; the second contact S1b of the switch is now closed to complete the power supply circuit of the modulator.

When operating from a 15 volt negative power supply the oscillator draws a collector current of approximately 4 mA. Each of the frequency multipliers draws 8 to 10 mA, which provides sufficient r.f. to drive the p.a. to 25 mA collector current.

Constructional Details

Normal techniques are adopted for the construction of the transmitter, which is assembled on a chassis measuring 5 in. square and 1½ in. high. All components with the exception of the crystal and output transistor are mounted on the underside of the chassis. TR4 is mounted on a small strip of polystrene with the three leads projecting through a hole to the underside of the chassis. The emitter and collector circuits of this stage are shielded by a screen, but this is not essential provided that L5, L6 and L7 are mounted at right-angles and are not closely spaced to each other.

Alignment

Alignment is best carried out by connecting the transistors in circuit one at a time, commencing with TR1, and tuning the appropriate collector circuit for maximum output as indicated on an r.f. voltmeter across the emitter winding of the succeeding stage. The circuit of a suitable meter is shown in Fig. 7.15. When the transistor is substituted for the meter it may be necessary to re-peak the preceding tuned circuit. It is essential that the frequency of each tuned circuit is checked by an absorption wavemeter during alignment as many frequencies are generated by the transistors, possibly due to leakage through the base-to-collector capacitance, together with the diode formed by the emitter-to-base junction acting as a mixer. It should be remembered that until drive from the oscillator is applied, TR2, TR3 and TR4 are in the cut-off condition.

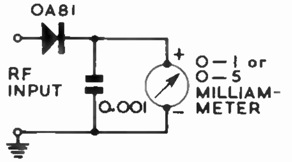


Fig. 7.15. R.F. voltmeter used for aligning transmitter in Fig. 7.14.

The second frequency multiplier is tuned by adjusting C8 and C11 to drive TR4 to maximum collector current with R6 set to maximum resistance. To tune the collector circuit of the amplifier stage, a 75 ohm resistor is connected across the coaxial socket CS1, and with C14 at maximum capacitance. C13 is tuned to resonance. The capacitance of C14 is then reduced in steps and C13 re-adjusted each time until maximum r.f. output is obtained. After the stage has been tuned R6 is adjusted to give maximum output.

A convenient method of measuring the r.f. output is by a transistorized field-strength meter similar to that described in Chapter 9 (*Measurements*). R.F. pick-up may be by a short length of wire located adjacent to the output socket of the transmitter. *When adjusting each stage it is essential that the collector current of the transistor is monitored to ensure that the rated input is not exceeded.*

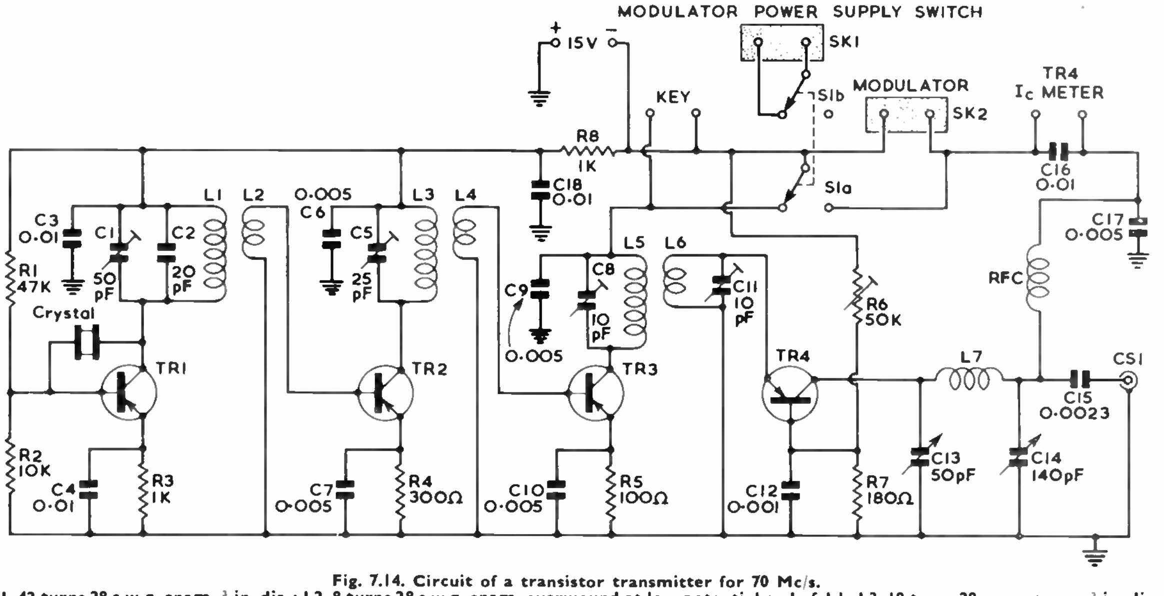


Fig. 7.14. Circuit of a transistor transmitter for 70 Mc/s.
L1, 42 turns 28 s.w.g. enam. ⅜ in. dia.; L2, 8 turns 28 s.w.g. enam. overwound at low-potential end of L1; L3, 18 turns 28 s.w.g. enam. ⅜ in. dia.; L4, 5 turns 28 s.w.g. enam. overwound at low-potential end of L3; L5, 12½ turns 22 s.w.g. enam. 0·3 in. dia.; L6, 5 turns 26 s.w.g. silk-covered interwound at low-potential end of L5; L7, 8 turns 16 s.w.g. silver-plated $\frac{7}{16}$ in. dia., ⅞ in. long (0·3 μH); RFC, 42 in. 28 s.w.g. enam. wound on ⅜ in. dia. former; TR1, Texas 2G401, Mullard OC170; TR2, TR3, Texas 2G402, Mullard OC171; TR4, Texas 2G110.

TWO METRE TRANSMITTERS

A range of crystal frequencies suitable for use in 2m transmitters, and the required multiplication factors are given in Table 7.5.

TABLE 7.5

Crystal frequencies and multiplication factors to give frequencies between 144 and 146 Mc/s.

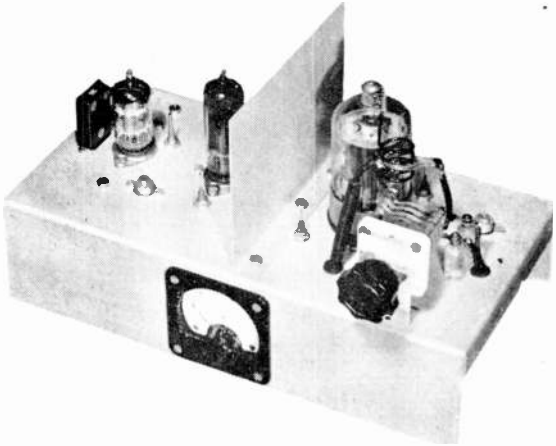
Crystal frequency (Mc/s)	Multiplication factor
6·0 to 6·083	3 × 2 × 2 × 2 4 × 3 × 2
7·2 to 7·3	5 × 2 × 2
8·0 to 8·111	3 × 3 × 2
9·0 to 9·222	2 × 2 × 2 × 2 4 × 2 × 2
12·0 to 12·166	3 × 2 × 2
24·0 to 24·33	3 × 2

A SIMPLE PHONE AND C.W. TRANSMITTER

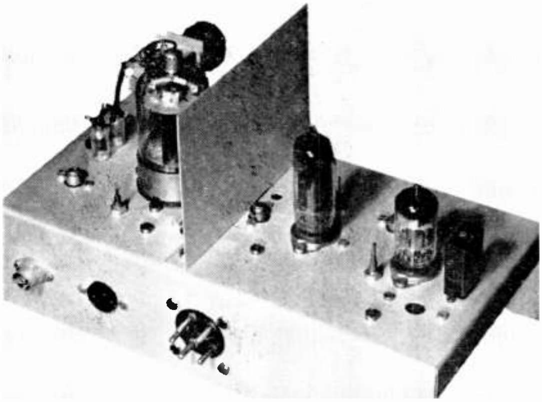
The three valve, four stage transmitter to be described employs a QV06–20 (6146) tetrode valve as a single-ended amplifier, which can be run at an input of up to 35 watts in telephony, and up to 45 watts in c.w. service. In these conditions the power requirements are an h.t. supply of 320 volts at 200 mA, and a heater supply of 6.3 volts at 2·3 amps. Approximately 20 watts of audio power are required for full anode and screen modulation, the impedance presented by the p.a. stage to the modulator being 2700 ohms.

Circuit Description

In the circuit diagram shown in **Fig. 7.16**, V1a, one half of an ECC81 (12AT7) double triode valve functions as a crystal oscillator and uses an 8 Mc/s type FT243 crystal in a Squier overtone circuit. The slug tuned inductor L1 is resonated at the third overtone of the crystal frequency. Capacitive coupling is employed between the anode of V1a and the grid of the second section of the double triode, V1b, which operates as a frequency tripler. The anode circuit L2, C6 is tuned to approximately 72·5 Mc/s, the exact frequency being dependent on the frequency of the crystal being used.



A simple phone and c.w. transmitter for 144 Mc/s employing a QV06-20/6146 valve in the p.a.



A rear view of the simple 144 Mc/s transmitter.

The output of V1b is capacitively coupled to the input of V2, a QV03–12 (5763) tetrode operating as a frequency doubler. The anode circuit of this valve, which comprises L3 and C11, is tuned to approximately 145 Mc/s, and is inductively coupled to the series-tuned grid circuit of the QV06–20 (6146) tetrode power amplifier V3. Grid resistors R11 and R12 of V3 are connected between the centre-tap of L4 and earth.

The anode circuit of V3 is also series-tuned and is formed by L5 and C16. H.T. is fed to the anode via the meter M1, choke RFC2 and the centre-tap of L5. R.F. output is fed to the coaxial socket CS1 by the single turn link L6, which is coupled to the centre of L5, and the tuning capacitor C18.

Screen neutralizing of the amplifier stage is effected by the pre-set capacitor C14 connected between the screen of V3 and earth. C14 is adjusted so that its capacitance, in conjunction with the screen lead inductance and stray capacitance, form a series-tuned resonant circuit from screen to earth. This places the screen of the valve at earth potential with respect to r.f. and effectively eliminates any feedback between anode and grid circuits through the valve.

Modulation is applied to the anode and screen of the p.a. via SK1. Use of a three pole socket for this purpose allows a screened connecting cable to be earthed to the chassis of the transmitter if desired.

Power supplies are connected to the transmitter through the five pole plug PL1, only three pins of which are normally used. The resistor R10 reduces the voltage on the h.t. line feeding V1 and V2 from 320 to 300 volts.

Resonance of the oscillator and multiplier anode circuits and the grid circuit of the amplifier stage is indicated by maximum grid current in the following stage; the grid current is read by connecting a milliammeter between the relevant test point (TP1, TP2 or TP3) and chassis. Resonance of the p.a. anode circuit is indicated by minimum anode current shown on the meter M1.

Construction Details

The transmitter is built on an 18 s.w.g. aluminium chassis measuring 10 in. long × 5 in. wide × 2½ in. deep with a ⅝ in. lip turned down at each end to increase its rigidity. The

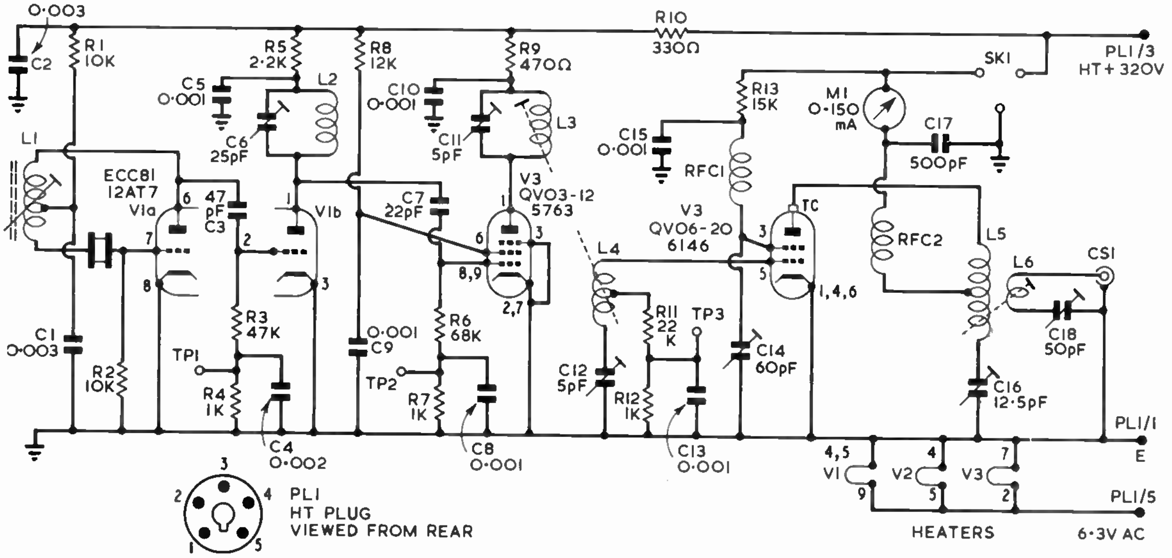


Fig. 7.16. Circuit diagram of the 144 Mc/s transmitter. C1, 2, 3, 5, 7, 9, 10, ceramic; C4, 8, 13, 15, 0·001 μF ceramic feedthroughs; C17, 500 pF 1000V d.c. wkg., mica; L1, 24 turns s.w.g. d.s.c., tapped 5 turns from grid end, wound on ⅜ in. diam. former, slug-tuned; L2, 4 turns 16 s.w.g. enam. ⅝ in. o.d., ½ in. long, self-supporting; L3, 2 turns 16 s.w.g. enam. ⅝ in. o.d., ⅜ in. long, self-supporting; L4, 3½ turns 16 s.w.g. enam. ⅝ in. o.d. ⅝ in. long, self-supporting; L5, 4½ turns 16 s.w.g. enam. ⅝ in. o.d., ⅞ in. long, self-supporting, with ⅞ in. lead to the anode cap of V3; L6, 1 turn 20 s.w.g. enam. ⅝ in. o.d.; M1, 0–150 mA d.c. meter; RFC1, 50 turns 34 s.w.g. enam. close wound on ⅛ in. diam. former; RFC2, 28 turns 26 s.w.g. d.s.c. ¼ in. diam. 1¼ in. long wound on Eddystone former type 863.

position of components can be located from the photographs and the drilling plan (Fig. 7.17).

Shielding between the p.a. valve anode circuit and the doubler valve V2 is effected by a vertical screen across the chassis at a distance of 4¾ in. from the right hand end. The screen is of 20 s.w.g. aluminium and measures 5 in. wide × 3½ in. high with the addition of a ⅜ in. wide lip by which it is secured to the chassis by three 4B.A. screws and nuts. The p.a. anode tuning capacitor C16 is mounted on an 18 s.w.g. aluminium bracket at a height of 1¼ in. above the chassis.

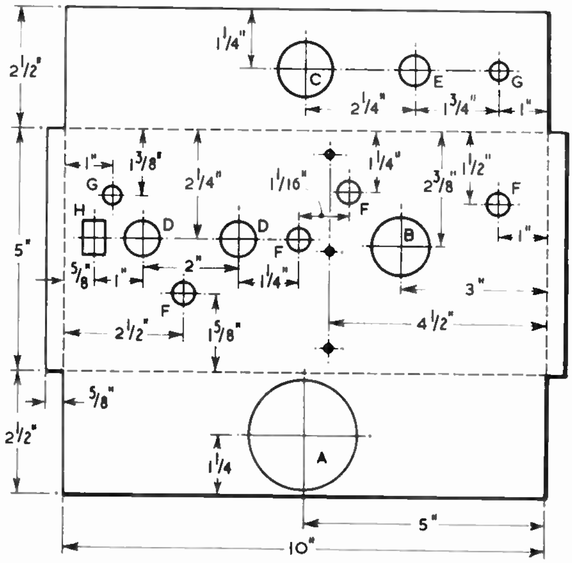
All bypass capacitors, with the exception of C17 are of the hi-K ceramic type; C17 is a 1000 volt working, silvered mica type. These capacitors are connected into the circuit with the shortest possible leads. The grid circuit capacitors are of the feedthrough type; one lead of each capacitor projects through the chassis to form the test points TP1, TP2 and TP3.

The aerial coupling loop L6 consists of a single turn of 18 s.w.g. enamelled copper wire, insulated with polythene sleeving and inserted at the centre of L5. This loop is supported by two ⅜ in. polystyrene stand-off insulators and is connected to the coaxial output socket via a short length of 75 ohm twin feeder, which is lead through a grommeted hole in the chassis. The output coupling capacitor is connected between one leg of the twin feeder and chassis.

Nylon-loaded valveholders are employed for V1 and V2. The holder for V3 is a ceramic type, and is mounted on the underside of the chassis; the tags on the valveholder that connect to the three cathode pins and brass sleeve of V3 and the heater return (pins 1, 4, 6, 7 and 8) are bent over and earthed by individual connections to soldering tags located under the two fixing nuts of the valveholder. These connections should be as short as possible and made with a heavy gauge conductor, not less than 18 s.w.g. tinned copper wire.

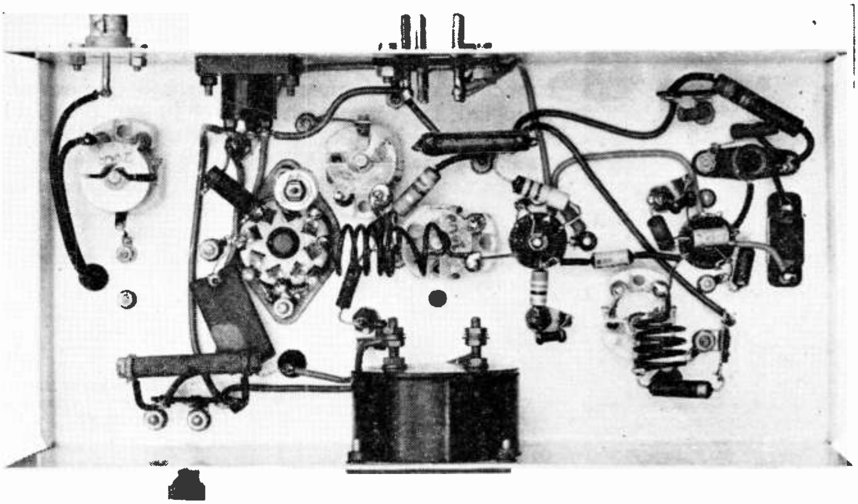
Setting up

Before power is applied to the transmitter it is recommended that the resonant frequencies of the tuned circuits are checked with a grid dip oscillator. L1 should be tuned to approximately 24 Mc/s; L2, C6 should be tuned to 72·5 Mc/s; L3, C11, L4, C12 and L5, C16 should be tuned to 145 Mc/s. As no protective grid bias is applied to any



HOLE DIAMETERS - A - 2⅛" B - 1¼" C - 1⅛"
D - ¾" E - ⅝" F - ½" G - ⅜" H - ¾" x 7/16"

Fig. 7.17. Drilling plan (above-chassis view).



Under-chassis view of the transmitter of Fig. 7.16.

stage of the transmitter, it is recommended that initial adjustments are carried out with a low h.t. voltage, not greater than 200 volts, applied to V1 and V2. For the same reason h.t. should not be connected to the anode and screen of V3 until the oscillator and multiplier stages have been adjusted and bias is being developed across the p.a. grid resistor. The most convenient method of removing h.t. from the p.a. stage is to leave the modulator socket SK1 open-circuited.

The procedure to be adopted in tuning the transmitter is as follows:

(a) Connect a 0–5 mA meter between TP1 and chassis (positive to chassis). Adjust the dust-iron core of L1 to obtain maximum current reading, then de-tune the circuit slightly to the high-frequency side of this position; this will result in a small drop in grid current.

(b) Connect the meter to TP2 and chassis and tune C6 for maximum grid current in V2. (A better indication is obtained if a 0–1 milliammeter is used in this position.)

(c) Connect the meter between TP3 and chassis; tune C11 and C12 in that order for maximum grid current in V3. Re-adjust C11.

(d) Check the neutralization of V3 by tuning C16 through resonance. Adjust C14 until the minimum change in grid current results as the p.a. anode circuit is swung through resonance.

(e) Short-circuit the modulator socket. Connect a dummy load (such as that described on page 7.50) to the aerial socket. With the low voltage still applied to the transmitter, tune C16 to resonance.

(f) Apply 320 volts to the transmitter and load the p.a. by increasing the coupling between L5 and L6 and adjusting C18 until an anode current at resonance of 110 mA is obtained. Check that the neutralizing is correct, i.e., that maximum r.f. output corresponds to minimum anode current; re-adjust C14 if necessary.

The loosest coupling between L5 and L6 consistent with obtaining the required power input should be the aim. It may be necessary to adjust the inductance of L3, L4 and L5 to enable these circuits to be peaked at the actual frequency being used. The coupling between L3 and L4 should be adjusted to give maximum grid current in V3 and it will be found that these circuits must be tightly coupled to obtain sufficient drive to the amplifier.

Typical values of anode and grid current readings are given in Table 7.6.

TABLE 7.6

Typical current readings for the 35 watt transmitter of Fig. 7.16.

Valve	Anode current	Screen current	Grid current
V1a	8·5 mA	—	—
V1b	11 mA	—	1·2 mA
V2	32 mA	4 mA	0·6 mA
V3	110 mA (loaded) (telephony) 140 mA (loaded) (c.w.)	8 mA	2·25 mA

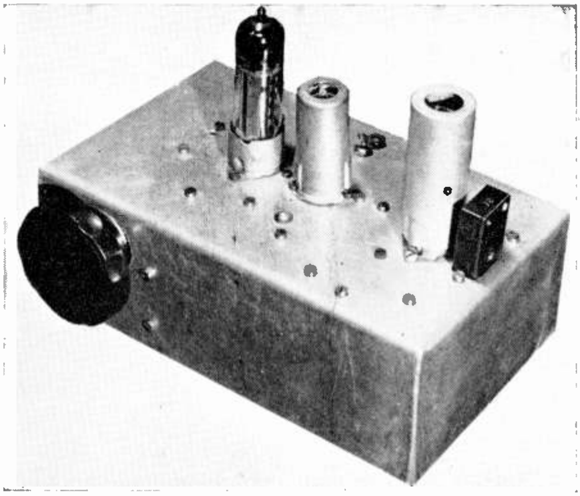
For c.w. operation of the transmitter it is recommended that the screen of the doubler valve V2 is keyed, and a clamp valve incorporated in the p.a. circuit to reduce the screen voltage of V3 in the spacing condition. The clamp circuit may be identical to that employed in the 4m transmitter shown in **Fig. 7.11.**

Unless the h.t. supply to the transmitter has good regulation, anode voltage for the oscillator and first multiplier should be derived from a separate source in order to avoid excessive variation in the anode voltage applied to these stages when the transmitter is keyed.

A 15 WATT PHONE TRANSMITTER

This three valve transmitter employs a double tetrode valve in the output stage which can be operated at a maximum input of 15 watts; in this condition it delivers approximately 6 watts r.f. into an aerial. The power supply requirements are an h.t. voltage of 200 to 250 volts at 120 mA and a heater supply of 6·3 volts at 1·33 amps.

The circuit of the transmitter is shown in **Fig. 7.18.** V1 is an ECL80 (6AB8) triode-pentode valve, the triode section of which functions as an overtone crystal oscillator, employing an 8 Mc/s fundamental frequency type FT243 crystal. The dust cored inductor L1 is tuned to the third overtone of the crystal frequency, approximately 24 Mc/s. Capacitive coupling is employed between the anode of V1a and the



A 15 watt 144 Mc/s transmitter for phone operation.

grid of the pentode section V1b, which operates as a frequency doubler. The anode circuit L2, C3 is tuned to 48 Mc/s.

V2 is an EL91 (6AM5) pentode operating as a frequency tripler. The anode circuit of this stage consists of L3, parallel-tuned to approximately 145 Mc/s by C9, which is connected between the anode of the valve and earth. Link-coupling is used to couple the tripler anode circuit to the grid circuit of the QQV03-10 (6360) push-pull power amplifier. V3 both link windings L4 and L5 are coupled to L3 and L6, respectively at points of low r.f. potential, that is, at the h.t. end of L3 and the centre of L6. The grid inductor L6 is self-resonant at 145 Mc/s with the input capacitance of V3 and stray circuit capacitances.

The p.a. tank circuit consists of L7 parallel-tuned by the split-stator capacitor C14. R.F. is fed to the coaxial output socket CS1 via a low impedance winding L8 coupled to the centre of L7. Modulated h.t. is applied to the screen and anode of V3 through R9 and R10 respectively.

Grid bias for all stages is derived from the flow of grid current through the respective grid resistor when excitation is applied, i.e. when V1 is oscillating. Test points TP1, TP2 and TP3 permit the grid current of V1b, V2 and V3 to be monitored when the transmitter is being initially tuned.

The transmitter is built on an 18 s.w.g. aluminium chassis measuring $7\frac{1}{2}$ in. long $\times$ $4\frac{1}{2}$ in. wide $\times$ $2\frac{1}{2}$ in. deep. All components, with the exception of the three valves and the crystal are mounted below the chassis. The dust-iron

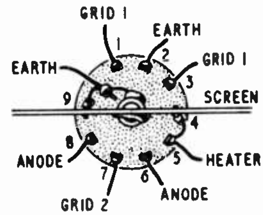


Fig. 7.19. Location of the screen across the valveholder for V3.

core of L1 and the two tuning capacitors C3 and C9 are accessible for adjustment from the underside of the chassis. C9 is spaced 1 in. from the chassis, being mounted on a 2BA threaded stud to which the rotor connection of the capacitor is soldered.

The link inductors L4 and L5 are connected directly to the tags of a small ceramic standoff insulator. L4 is interwound with the low-potential (h.t.) end of L3; L4 should be insulated with thin sleeving to prevent a short-circuit between the windings should this be considered necessary.

The grid resistors R3, R5 and R8 are returned to chassis via single-point tag strips, which form the three test points. The bypass capacitors C5, C8 and C12 are connected between the relevant tag strip and earth; each capacitor is bridged by a low value resistor, one end of which is disconnected when it is desired to monitor the grid current during

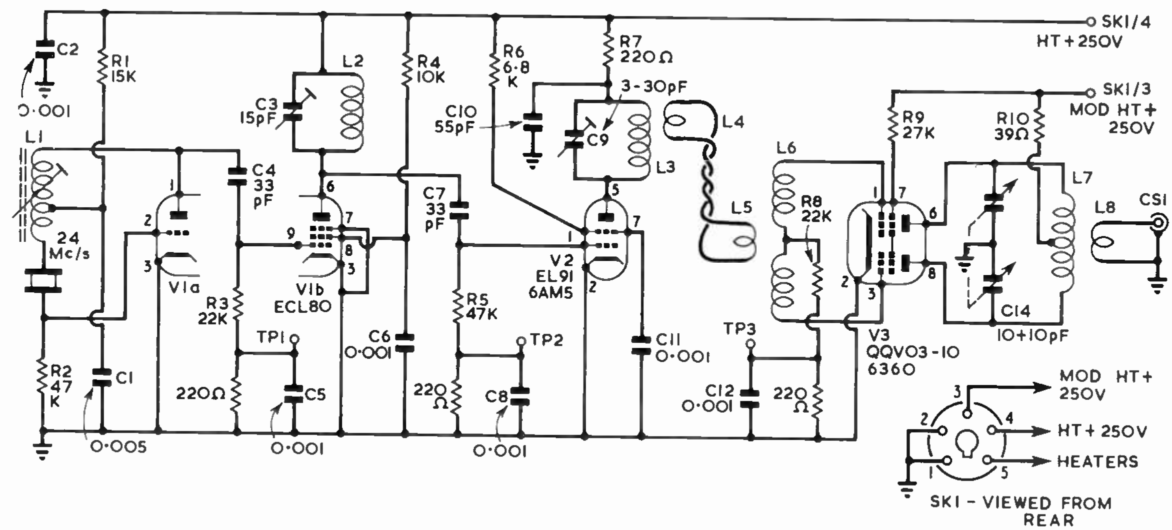
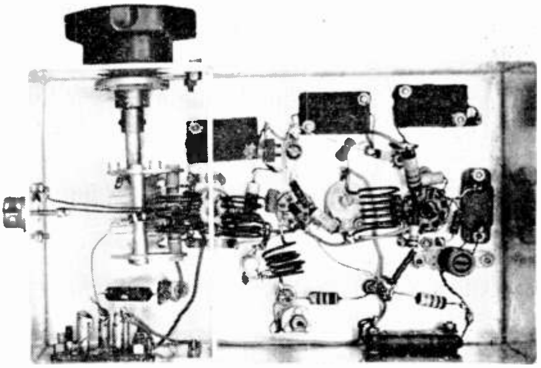


Fig. 7.18. Circuit diagram of the 15 watt 144 Mc/s transmitter. C2, 6, 10, 11, hi-K ceramic; C4, 7, ceramic; L1, 22 turns 22 s.w.g. enam. tapped $7\frac{1}{2}$ turns from grid end wound on $\frac{3}{8}$ in. diam. former slug tuned; L2, 6 turns 18 s.w.g. enam. $\frac{1}{2}$ in. diam. $\frac{1}{2}$ in. long; L3, 4 turns 18 s.w.g. enam. $\frac{1}{2}$ in. diam. $\frac{1}{2}$ in. long; L4, 2 turns 18 s.w.g. enam. $\frac{1}{2}$ in. diam. interwound with earthy end of L3, L5, 2 turns 18 s.w.g. enam. $\frac{1}{2}$ in. diam; L6, 4 turns 18 s.w.g. enam. $\frac{1}{2}$ in. diam. $\frac{5}{8}$ in. long split at centre for L5; L7, 4 turns 18 s.w.g. enam. $\frac{3}{4}$ in. diam. $\frac{7}{8}$ in. long split at centre for L8,; L8, 2 turns 20 s.w.g. enam. $\frac{3}{8}$ in. diam.



Under-chassis view showing the layout of the components. Note the screen across the p.a. valveholder.

initial tune-up. A switch can then be arranged to switch a low reading milliammeter across each test point.

Underneath the chassis the p.a. grid and anode circuits are isolated by a screen mounted across the chassis. The screen is 2 in. deep, to provide clearance between the upper edge and the p.a. valveholder contacts. Fig. 7.19 shows how the screen is located across the diameter of the valveholder which is positioned so that pins 4 and 9 are in line with it.

In the transmitter illustrated the p.a. anode tuning capacitor C14 is mounted by two 6BA screws which engage in threaded holes in one of the bars forming the capacitor frame. Each screw carries a ¼ in. diameter spacer located between the frame of the capacitor and the chassis. Alternatively C14 may be mounted on a small aluminium bracket for which three hole fixing is provided on the capacitor. A small epicyclic slow-motion drive is fitted to C14.

Screening cans are fitted to V1 and V2; the can fitted to V2 is essential as the capacitance between the earthed can and the anode of V2 forms part of the anode circuit tuning capacitance. Owing to the amount of heat generated by V3 no screening should be used.

Alignment

Initial adjustments should be carried out with a low h.t. voltage, not exceeding 200 volts, applied to V1 and V2 only. H.T. is most conveniently removed from the anode and screen of V3 by leaving SK1/3 disconnected.

The procedure to be adopted in tuning the transmitter is as follows:

(a) Disconnect the jumper from TP1 and connect a low-range milliammeter across the test point. Adjust the dust-iron core of L1 for maximum grid current in V1b, then detune slightly to the h.f. side of resonance. Remove the meter and reconnect the jumper.

(b) Tune the anode circuits of V1b and V2 by adjusting C3 and C9 for maximum grid current in the meter when connected to TP2 and TP3 respectively. Adjust the coupling of the link windings L4 and L5, and the inductance of L6 so that maximum grid current in V3 is obtained.

(c) Connect a dummy load across CS1, apply h.t. to the anode and screen of V3 by connecting a 0–100 milliammeter between SK1/3 and SK1/4. Tune C14 to resonance as indicated by the dip in anode current. (A suitable dummy load consists of two 12 volt 0·3 amp. MES bulbs connected in parallel. These lamps should light to full brilliance when an h.t. voltage of 250 is applied to the transmitter.)

Typical current readings are shown in Table 7.7.

The transmitter is modulated by connecting the secondary winding of the modulation transformer between SK1/3 and SK1/4. With the aerial connected the coupling between L7 and L8 should be adjusted to give the required anode current to the p.a. stage, up to the maximum permissible limit of approximately 60 mA. In this condition, with 250 volts h.t. applied to V3, the impedance presented by the p.a. to the modulator is 4200 ohms; approximately 7 watts of audio are required for 100 per cent modulation.

TABLE 7.7

Typical current readings for the 15 watt 2m transmitter of Fig. 7.18

Valve	Anode current	Screen current	Grid current
V1a	5 mA	—	—
V1b	26 mA	5 mA	0·6 mA
V2	22 mA	3 mA	1·25 mA
V3 (loaded)	60 mA	3 mA	2 mA

A COMPACT TRANSMITTER FOR 144 Mc/s

The circuit of a compact transmitter for 144 Mc/s based on the popular QQV03-20A (6252) double tetrode power amplifier is shown in Fig. 7.21. The complete unit is built on a standard Eddystone die-cast box type 903 measuring 7¼ in. × 4½ in. × 3 in. A drilling plan is given in Fig. 7.20.

As may be seen from Fig. 7.21 the circuit is a five stage arrangement in which the crystal oscillator V1 (EF91, 6AM6) employs six switched 6 Mc/s crystals. The oscillator is a standard Colpitts with the anode circuit tuned to twice the crystal frequency (12 Mc/s). The second stage, V2, another EF91/6AM6, operates as a frequency doubler with its anode circuit (L2) tuned to 24 Mc/s. It would be possible to replace these two stages by a single valve using 8 Mc/s crystals and an anode circuit tuned to 24 Mc/s, but the arrangement shown is an almost fool-proof method of ensuring there is sufficient drive to subsequent stages.

The third stage, V3 (QV03-12, 5763) is a frequency tripler with its anode circuit tuned to 72 Mc/s. V4, also a QV03-12, is a doubler to the final frequency of 144 Mc/s. The anode circuit of this stage, unlike those for V2 and V3, is series tuned to give greater output than is possible with the conventional parallel-tuned arrangement. In this configuration, C2 should be as nearly equal to the output capacitance of V4 as possible. L4 must be adjusted to achieve this condition.

The output from the anode circuit of V4 is link coupled to the input circuit of V5, the QQV03-20A (6252) p.a. stage. This double tetrode is internally neutralized and provided there is adequate isolation between the input and output circuits no external neutralization should be required.

The anode circuit of the p.a. stage, C3, L6, is a simple coil parallel tuned by a split stator capacitor. H.T. is fed to the anodes of the valve through RFC2 connected to the centre tap of L6. This r.f. choke comprises a $\lambda/3$ winding of 26 s.w.g. enamelled wire on a ¼ in. diameter former. Coupling to the output circuit is by L7, a single turn coil series tuned by C4.

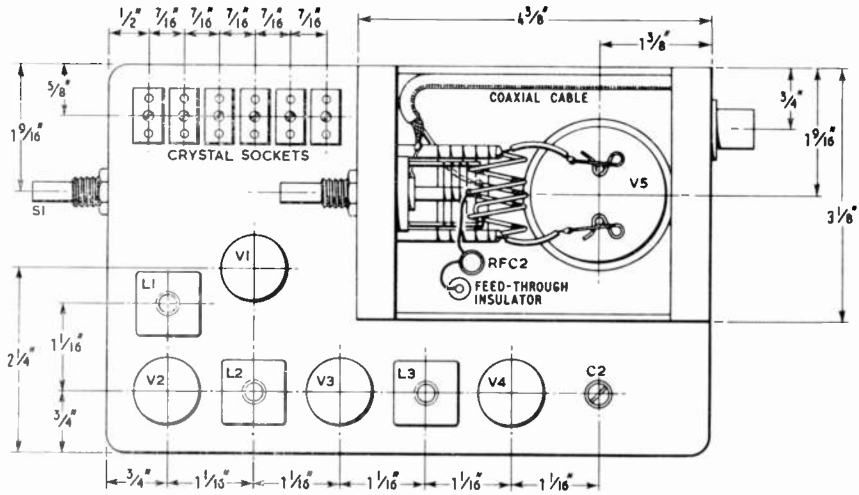


Fig. 7.20. Diagram giving above-chassis component layout together with drilling and other important dimensions.

Test points are provided in each of the anode supply leads and in the grid circuit of all stages except the crystal oscillator.

TABLE 7.8

Operating conditions for the QQV03-20A (6252) as a class C anode and screen modulated amplifier up to 200 Mc/s

Anode Voltage	300	500	600
Screen Voltage	250	250	250
Grid Bias	−50V	−80V	−80V
Anode current (total)	80 mA	80 mA	80 mA
Screen Current (total)	8 mA	8 mA	8 mA
Grid Current (total)	2 mA	2 mA	2 mA
Screen Dropping Resistor (Rs)	4·25 K ohms	32 K ohms	42 K ohms
Grid Bias Resistor (Rg)	25 K ohms	40 K ohms	40 K ohms
Power Input (anode)	24 watts	40 watts	48 watts
Driver Power Output	1·5 watts	3 watts	3 watts
Power Output in Load	14 watts	24 watts	32 watts
Audio Power for 100 per cent Modulation	13 watts	21 watts	25 watts

Constructional Details

The complete unit is constructed on an Eddystone die-cast box with the power amplifier and its output circuit fitted into an additional screening box mounted on top of the chassis. The layout of the p.a. tank circuit is shown in **Fig. 7.23.**

The layout is straight-forward and **Fig. 7.22,** in conjunction with Fig. 7.20, shows the location of all the major components. Starting at the six-way crystal socket, the various stages are arranged along the front and side of the die-cast box.

A tag strip runs alongside V2, V3 and V4 and provides for the location of small components and test points. Unused tags in the bases of the coil formers are used as solder points to avoid joining two components together without a tiepoint.

The output from the final frequency multiplier is taken from a single turn coil via a short piece of 75 ohm twin feeder to another single turn coil, coupled to the grid circuit of the p.a. valve. This arrangement is shown clearly in Fig. 7.22. Stand-off insulators are used to provide suitable fixing points for the link coupling circuit which runs about 1 in. below the chassis.

The p.a. anode and aerial coupling circuit are housed in a 4⅜ in. × 3⅛ in. × 3½ in. box which is mounted above the chassis.

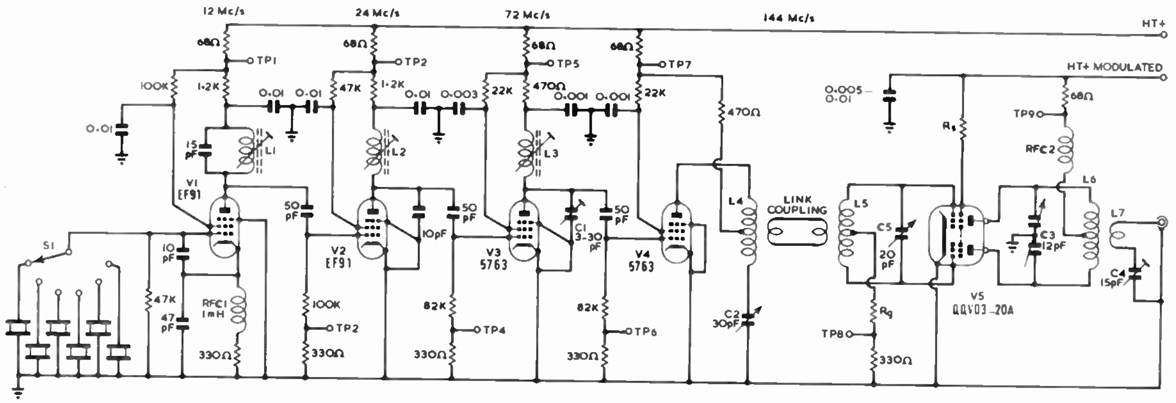


Fig. 7.21. Circuit diagram of the compact transmitter for 144 Mc/s. C1, C5, 3-30pF Philips trimmers; C2, 30 pF air spaced trimmer; C3, 12 pF + 12 pF slit stator capacitor (Eddystone type 587); C4, 15 pF air spaced trimmer; L1, ½ in. long winding, of 36 s.w.g. enam. wire on ¼ in. diam. former; L2, 15 turns 30 s.w.g. enam. ¼ in. diam. former, turns spaced 1 wire diam.; L3, 7 turns 16 s.w.g. ½ in. diam. ¾ in. long; L4, 5 turns 16 s.w.g. enam. ¼ in. diam. close wound; L5, 5 turns 16 s.w.g. enam. ⅜ in. diam. spaced 1 wire diam.; L6, 3 turns 14 s.w.g. enam. ⅜ in. diam. ¾ in. long; L7, 1 turn 18 s.w.g. insulated wire. Values of Rg and Rs may be obtained from Table 7.8.

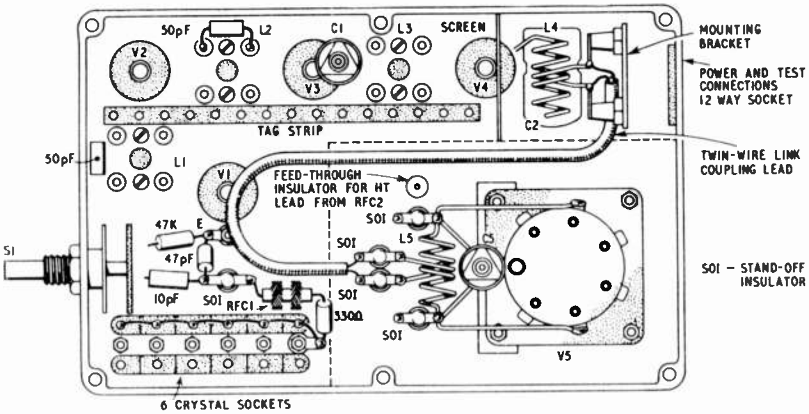


Fig. 7.22. Under-chassis component layout. Position of 75 ohm feeder cable link between final multiplier and p.a. is also clearly shown.

The test points may be fitted to the chassis at the most convenient place and tests made with a flying lead. Alternatively meters and switches may be incorporated on a front panel of suitable size.

The valveholder for the power amplifier V5 is submounted so that the valve's internal screen is approximately in line with the top of the chassis. A hole 2 in. in diameter is made in the chassis to allow the p.a. valve to be plugged into the socket and to leave clearance around it for ventilation.

H.T. for the p.a. is fed through an insulator and together with the other power supply leads is taken from a multiple connector socket.

Alignment

The alignment is straightforward and there is no neutralization to adjust. As with most v.h.f. transmitters, the precise adjustment of the p.a. tuning and coupling circuits is best carried out with the aid of a suitable r.f. indicator such as a simple r.f. voltmeter connected to the feeder or, better still, a s.w.r. meter of the type described in Chapter 19—*Measurements*. In this way, false tuning indications can be avoided. It should be remembered that maximum dip in the anode current of a tetrode valve seldom coincides with maximum power output.

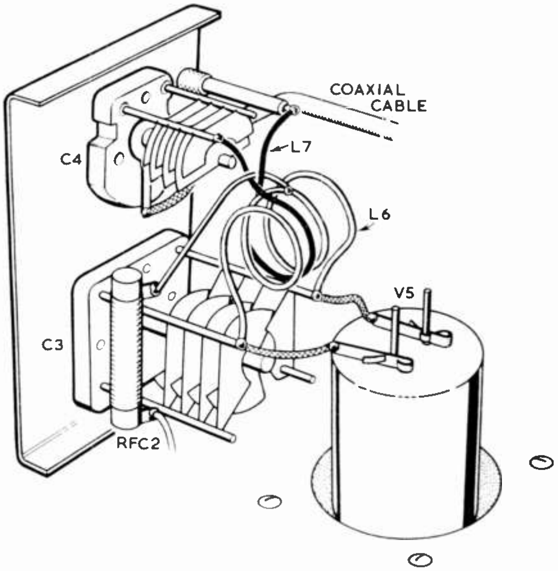


Fig. 7.23. Layout of the p.a. tank circuit.

LOW POWER TRANSMITTER FOR PORTABLE OR MOBILE STATIONS

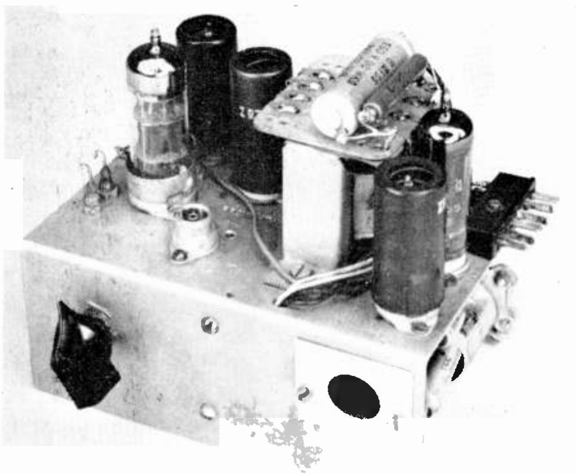
The small telephony transmitter illustrated is primarily intended for portable or mobile operation but can also be used in a fixed station. A Mullard QQV02–6 miniature v.h.f. double tetrode is employed in the power amplifier stage, which may be run at a maximum input of 7·5 watts; with this input an r.f. power output of approximately 5 watts can be obtained. A two-valve modulator is incorporated, which provides anode and screen modulation of the p.a. stage. The circuit is shown in **Fig. 7.24.**

Power requirements are an h.t. voltage of 220 volts at approximately 115 mA, and a 12 volts d.c. supply which is used to heat the valves and operate a receive-to-transmit changeover relay. When used as a fixed station transmitter an a.c. heater supply of either 6 or 12 volts may be used by connecting the heaters in parallel or series-parallel respectively but a suitable d.c. voltage must be available to operate the aerial change-over relay.

In the transmitter illustrated a 4·5 volt battery is used to energize the carbon microphone, but an alternative method whereby the microphone current is derived from the h.t. supply may be substituted if desired.

Circuit Description

The circuitry of the crystal oscillator and first multiplier stages is similar to the corresponding stages in the 35 watt transmitter of Fig. 7.16. In this design an ECC91/6J6 double triode valve (V1) is employed, using an FT243 type crystal having a frequency in the 8 Mc/s range; the anode circuit of the Squier oscillator (V1a) and the multiplier (V1b) are tuned to frequencies of approximately 24 and 72 Mc/s respectively. The oscillator anode inductor L1 is slug-tuned and the doubler anode inductor L2 is parallel-tuned by the air trimmer C4. Capacitance coupling is used between the anode circuit of V1a and the grid circuit of V1b and in all interstage coupling circuits throughout the transmitter. A second ECC91/6J6 (V2) with the electrodes of each section connected in parallel to form a single triode, functions as a doubler from 72 to 144 Mc/s. H.T. supply to the anodes of V2 is applied via a centre-tap on L3, parallel-tuned by C6.



Low power 144 Mc/s transmitter for portable or mobile service.

The p.a. valve V3 is a QQV02–6 double-tetrode; as in the earlier stages of the transmitter, the cathodes of this valve are earthed directly and grid leak bias only is applied. Grid circuit decoupling is effected by a feed-through capacitor C9; R7 is located so that the earthy end is readily detached from the chassis to permit insertion of a meter in series with the resistor when initially tuning the transmitter. Anode and screen circuits are decoupled by R8, C10 and R9, C11 respectively; the screen resistor is not bypassed directly to earth. R.F. output from the anode circuit of the p.a. is taken from a link winding L5 inductively coupled to the centre of L4.

A simple two stage resistance coupled amplifier consisting of an EC90/6C4 triode (V4) and EL90/6AQ5 pentode (V5) is employed which provides 100 per cent modulation when used in conjunction with a single-button carbon microphone. Operation of the send-receive switch on the microphone S1 energizes the coil of the changeover relay RLA/2 from the 12 volt d.c. supply; RLA1 disconnects the aerial socket (CS1) from the receiver socket (CS2) and connects CS1 to the output of the transmitter; RLA2 disconnects the h.t. positive supply line from the transmitter and modulator, and extends the supply to PL1/2, where it is available to operate a receiver.

If desired the microphone current may be obtained from the h.t. supply, thus eliminating the need for a separate battery. A suitable arrangement which may be substituted for the microphone circuit in Fig. 7.24 is shown in Fig. 7.25.

Coupling between the modulator and p.a. valves is effected by the centre-tapped primary winding of T2, which acts as a 1 : 1 ratio auto-transformer. A resistor R15 is included in series with the modulated h.t. supply to the anode and screen of the p.a. to reduce the h.t. applied to this valve to 180 volts.

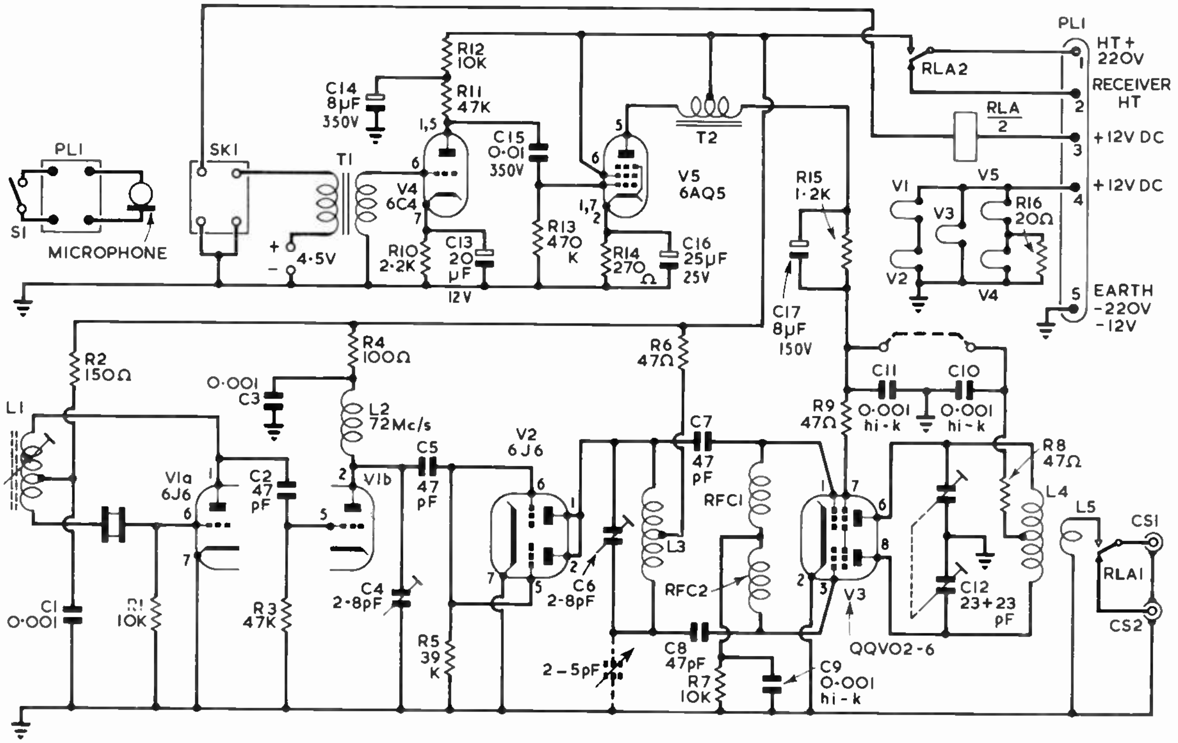
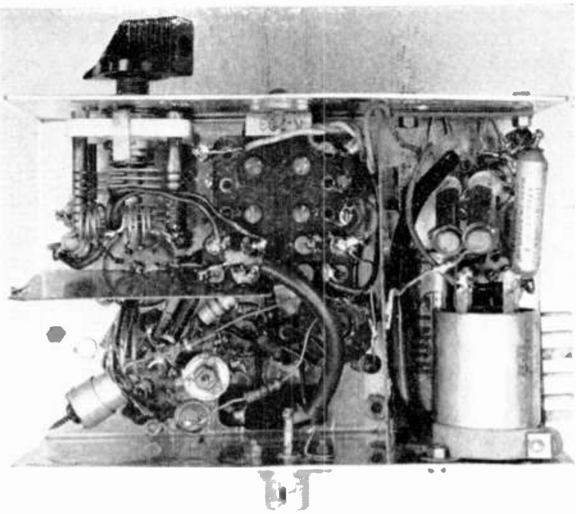


Fig. 7.24. Circuit diagram of the low power transmitter. C1, 3, 0·001 μF ceramic; C9, 10, 11, 0·001 μF hi-K ceramic feed throughs; L1, 20 turns 24 s.w.g. enam. tapped 5 turns from grid end, wound on 3/8 in. diam. former, slug-tuned; L2, 6 turns 20 s.w.g. tinned copper 1/2 in. o.d. 1/2 in. long; L3, 3 turns 18 s.w.g. tinned copper 1/2 in o.d. 1/2 in. long; L4, 3 + 3 turns 20 s.w.g. tinned copper 3/4 in. long, 1/4 in. gap at centre for L5, ; L5, 2 turns 22 s.w.g. tinned copper 3/8 in. o.d., p.v.c. insulated; RFC1, 2, 35 turns 30 s.w.g. enam. close wound on 3/16 in. diam. 470K ohm 1/2 watt resistor; T1, microphone transformer ratio 1 : 60 (Wearite type 207 suitable); T2, push-pull speaker output transformer, primary only used. The relay RL1 is a miniature 12 volt type with double pole changeover contacts.



Under-chassis view of the low power phone transmitter. The microphone transformer is at the lower right and the p.a. tuning capacitor at the top left.

R15 is bypassed by C17 to prevent attenuation of the audio frequencies by the resistor.

Valve heaters are connected in series-parallel for operation from a 12 volt battery. R16 is shunted across the heater of V4 to equalize the current flowing through the heater of V5.

Construction

The transmitter is constructed on an 18 s.w.g. U-shaped aluminium chassis measuring 5½ in. long × 3¾ in. wide × 2½ in. deep. Layout of components is clearly shown in the photographs. The aerial socket CS1 is located on the top of the chassis in front of the oscillator-multiplier valve V1. On the rear drop of the chassis are mounted the receiver socket CS2 and microphone transformer T1. At the extreme left-hand side adjacent to V3 are the p.a. anode and screen feed through capacitors C10 and C11;

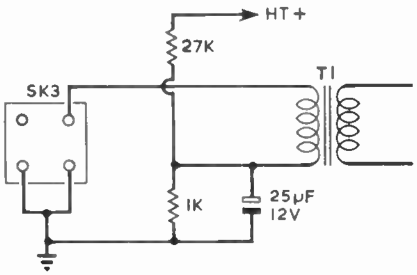


Fig. 7.25. Alternative microphone circuit for obtaining energizing voltage from the h.t. line.

these provide a convenient point for connecting an anode current meter to the p.a. stage during initial alignment, and may be either short-circuited or connected to a monitoring meter when the transmitter is subsequently being operated. R7, the grid resistor for V3, is located on top of the chassis behind the p.a. valve and is connected between the feedthrough capacitor C9 and a soldering tag on the chassis. This arrangement permits grid current to be monitored by disconnecting the resistor from the tag and inserting a low-range milliammeter at this point. R15 and C17 are wired in parallel and connected between one end of the primary of T2 and an unused tag on the transformer panel.

All tuning controls are adjusted from the underside of the chassis. Each of the two concentric trimmers C4 and C6 is connected directly across and supported by inductor L2 or L3 respectively.

The receive-to-transmit changeover relay is mounted towards the front of the chassis adjacent to C12. If the transmitter is built on a chassis of the size specified, the actual relay used must not occupy a space exceeding 1½ in. cube; if the relay is larger than this it will be necessary to increase the chassis dimensions. A vertical shield made of 22 s.w.g. tinplate running the full width of the chassis screens the modulator from the r.f. section. Screening between amplifier grid and anode circuits is effected by a tinplate shield measuring 2½ in. long × 2¼ in. deep. This shield has a cut-out to accommodate the valveholder and is soldered to the centre spill and to tags 2 and 5 of the valveholder; the shield is earthed by soldering it to tags mounted under the fixing nuts of the valveholder (Fig. 7.26(b)).

Alignment and Operation

Initial adjustments should be carried out with h.t. removed from the anode and screen of the power amplifier valve.

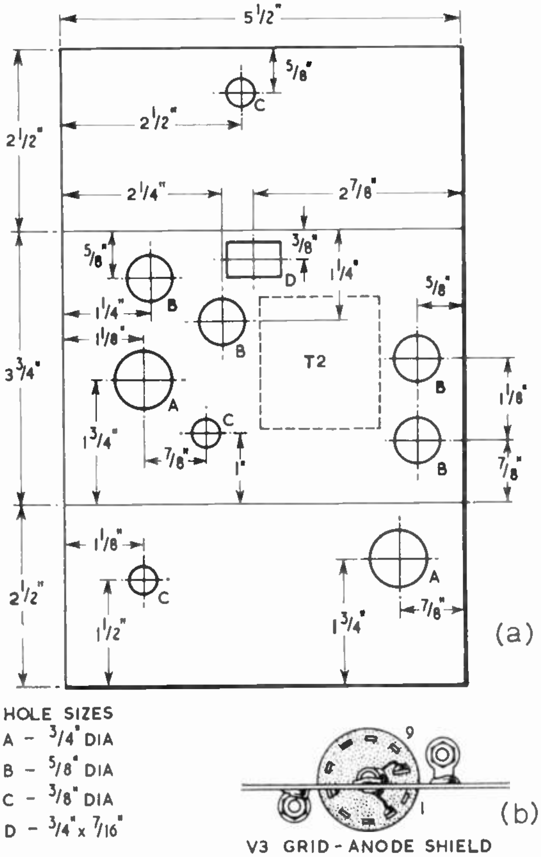


Fig. 7.26. (a) Drilling plan. (b) Position of V3 screen.

This can be done by disconnecting one side of R15 and C17 from the modulation transformer. For alignment on the bench or for fixed station operation the 12 volt heater supply, connected between PL1/1 and PL1/5 (earth) may be either d.c. or a.c.; the total heater current is 1·2 amps. A 12 volt d.c. supply must be available for operating the changeover relay, and is connected between PL1/3 and earth. Alternatively, a small silicon rectifier may be wired in series with the relay coil and a 100–500 μF 25 volts working capacitor used for smoothing. With such an arrangement, the transmitter can be used on d.c. or a.c. without modification. An h.t. supply of between 200 and 220 volts should be applied between PL1/1 and earth.

A crystal having a frequency between 8·0 and 8·11 Mc/s is required. After power supplies have been connected the most convenient method of aligning the anode circuits of V1a and V1b is to disconnect one end of R2 and R4 respectively, insert a milliammeter in series and tune for maximum dip in anode current. The anode circuit of V2 may then be tuned by disconnecting the earthy end of R7 from the chassis and inserting a low range milliammeter (0–1 mA) at this point. C6 is then adjusted for maximum grid current (approximately 1 mA).

A 0–100 milliammeter should be connected between C10 and C11, R15 and C17 reconnected, and h.t. applied to V3. A dummy load, for example a 6 volt 0·6 amp. bulb, may be connected to CS1. C12 should then be tuned for maximum dip in anode current, and the coupling between L4 and L5 adjusted so that the anode current does not exceed 40 mA. Speaking closely into the microphone should result in increased bulb brilliancy on peaks of speech.

When the station is set up for operation R15 should be adjusted so that the h.t. voltage applied to V3 is 180 volts. It is important that this figure is not exceeded.

A MEDIUM POWER TWO METRE AMPLIFIER

The 2m power amplifier uses a QQV06–40A/5894 double-tetrode valve in conjunction with a quarter wavelength linear anode circuit. It may be run at a maximum input of 90 watts on phone or 120 watts on c.w. with natural cooling; in these conditions an h.t. supply of 600 volts at 175 mA or 220 mA respectively, is necessary. A heater supply of 6·3 volts at 2·25 amps. is also required. The efficiency of the stage is approximately 70 per cent which decreases only slightly when operated at a lower anode input, i.e. at a reduced h.t. voltage. An r.f. drive power of approximately 5 watts is required to obtain full output from the stage.

For high level modulation an audio output of 50 to 60 watts is necessary, which can be obtained, for example from two 807's operating in class AB2 or KT66 or KT88 in class AB1. The modulating impedance depends on the power input to the stage, and representative values are 2400 ohms at 60 watts input (400 volts at 150 mA) and 4000 ohms at 90 watts input (600 volts at 150 mA). For c.w. transmission the driver may be keyed and to this end a clamp valve is included which limits the anode dissipation of the p.a. valve to an acceptable value in the " key up " condition.

Circuit Description

The circuit is shown in Fig. 7.27. R.F. drive is applied to the push-pull grid circuit L2, C1. Grid bias is obtained solely from current flow through R2 when excitation is applied. For this reason a tetrode clamp valve V2 (6V6) or equivalent is connected between the amplifier screen and earth, so that when the p.a. is driven the negative bias developed across R2 is also applied to the grid of the clamp valve, biasing it to well beyond cut-off; it has thus no effect on the operation of the circuit. Removal of the drive, and hence removal of the bias from V2 grid, causes this valve to conduct heavily, resulting in a greatly increased voltage drop across R4. The screen voltage of V2 thus drops to a value such that the anode current is reduced to a figure that is well within the rated dissipation (40 watts) of the valve.

Use of a tetrode clamp valve is imperative with a valve such as the QQV06–40A. A triode-connected clamp valve will not reduce the screen voltage to a sufficiently low value to protect the p.a. valve.

The anode circuit L3 is a quarter wave line, tuned by C3 which is tapped down the line a short distance from the

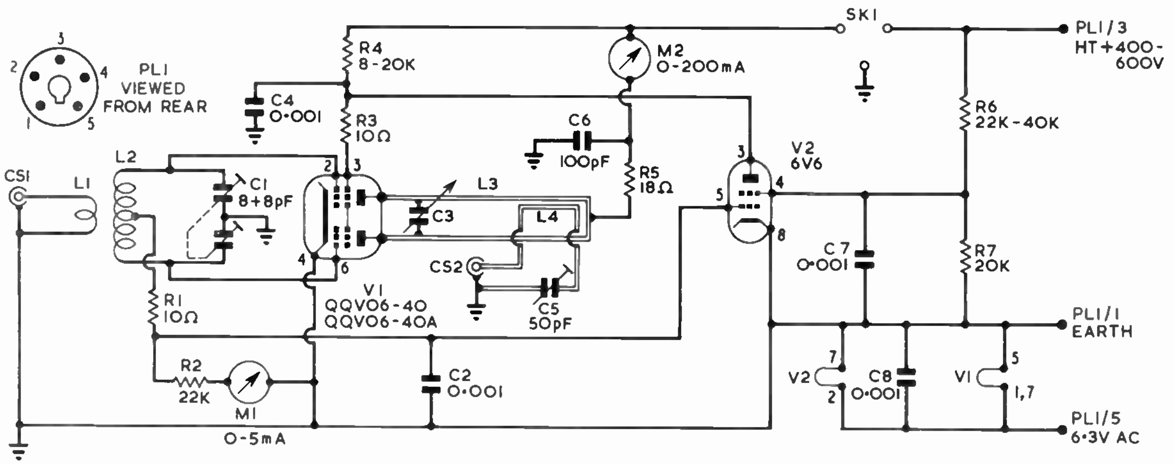
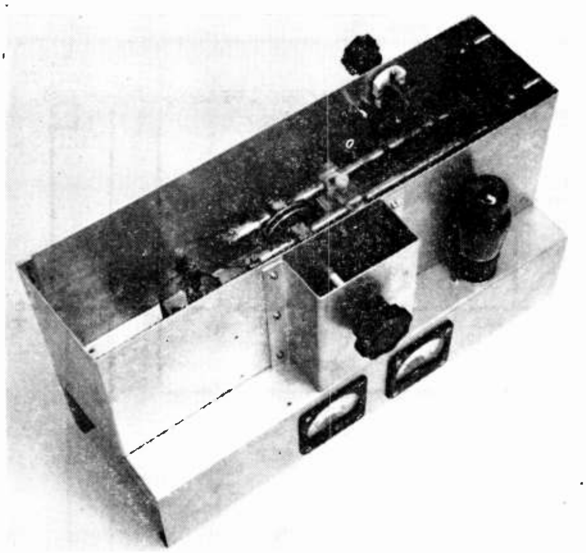


Fig. 7.27. Circuit diagram of the 144 Mc/s p.a. C2, 4, 7, 8, 0·001 μF mica; C3, see text; C6, 100 pF 2500 V d.c. wkg., mica; L1, 1 turn 20 s.w.g. enam ⅛ in. diam. insulated with polythene sleeving, at centre of L2; L2, L4, 4 turns 16 s.w.g. enam. ⅜ in. i.d. 1¼ in. long; L3, parallel-line anode circuit consisting of two ⅜ in. diam. copper tubes, 8 in. long, silver plated; L4, output coupling loop, 16 s.w.g. enam. 2¾ in. long, 1 in. wide; R1, 3 10 ohms wire wound.



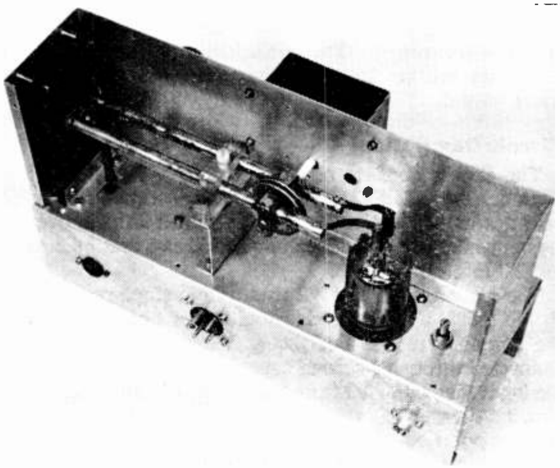
144 Mc/s power amplifier using a QQV06–40A (5894).

anode end of the line. R.F. output is taken from a loop L4 which is coupled to L3 adjacent to the short-circuited end.

The value of the screen dropping resistors for V1 and V2 (R4 and R6 respectively) is chosen according to the h.t. voltage so that the screen dissipation of the valves is not exceeded. The range of values shown is suitable for voltages between 400 and 600. Resistors R1 and R3 in the grid and screen circuits respectively are wirewound to reduce any tendency to parasitic oscillation in these circuits.

Construction

The amplifier illustrated is built on an aluminium chassis 14 in. long × 5 in. wide × 3 in. deep; the output circuit compartment is $3\frac{1}{2}$ in. wide and $4\frac{1}{2}$ in. high. Dimensions are not critical and may be chosen to suit any particular installation. Complete isolation between grid and anode circuits is achieved by mounting the valveholder below a 2 in. diameter hole in the chassis on $\frac{5}{8}$ in. pillars, so that the horizontal screening disc inside the valve is level with the top of the chassis.



Linear tank circuit for the 144 Mc/s p.a.

The anode lines can be easily fabricated and consist of two lengths of $\frac{3}{8}$ in. diameter brass tubing 8 in. long, spaced 1 in. apart. At the short circuited end the two tubes are joined by a strip of heavy gauge (e.g. 12 s.w.g.) brass, in which are drilled two $\frac{3}{8}$ in. diameter holes centred 1 in. apart, into which the tubes are pushed after filing to fit and then soldered. H.T. (via R5) is connected to the mid-point of the short circuiting strip. At a distance of $1\frac{5}{8}$ in. from the open circuited end of the lines the anode tuning capacitor C3 is mounted; each plate of this capacitor consists of a brass disc $1\frac{3}{4}$ in. in diameter attached to a threaded stud which engages in a tapped hole in a brass block soldered to the lines. A fine thread (e.g. 40 t.p.i.) should be cut on the studs to ensure smooth movement of the capacitor plates. A slot should be cut at the end of one stud to engage with a small metal blade fixed to the end of a $\frac{1}{4}$ in. diameter polystyrene rod which forms the tuning control; this rod may be spring-loaded to avoid any tendency of the blade to jump out of the slot as the control is rotated.

It will be found that resonance occurs when the two plates are approximately $\frac{1}{8}$ in. apart and they should be adjusted so that this distance is obtained when the front plate is at the mean position of its travel. The threaded studs should be a reasonably tight fit in the tapped holes to ensure that there is no electrical discontinuity between the capacitor and the lines. It is advisable for the rear plate to be locked in position after resonance has been determined.

Anode connectors must be used and can conveniently be made of $\frac{5}{16}$ in. brass rod $\frac{3}{4}$ in. long, similar to those employed for the 70cm transmitter (Fig. 7.39(d)). Copper strip $\frac{1}{4}$ in. wide and $2\frac{1}{4}$ in. long is used to join the end of the lines to each anode connector. The anode circuit lines are supported at the end by a $\frac{1}{4}$ in. thick Paxolin plate to which the short-circuiting strip is screwed, and by two polystyrene blocks mounted on an aluminium bracket located $4\frac{1}{2}$ in. from the end of the lines.

Operation

To set up the unit apply heater voltage only and connect a load to CS2. Excitation may then be applied to CS1 and C1 tuned to resonance. The output of the exciter and coupling between L1 and L2 should be adjusted to give a grid current of 4 mA. H.T. may then be applied to the amplifier and C3 tuned; the dip in anode current at resonance should correspond to maximum r.f. output. Coupling between L3 and L4, and the capacitances of C5 should then be adjusted to load the stage to the required rating. Finally, the value of R2 and R4 should be adjusted to obtain maximum r.f. output for a given d.c. input to the stage, without exceeding the rated anode and screen dissipation of the valve. After the anode circuit has been tuned, h.t. voltage may be applied without grid drive, as the clamp valve effectively reduces the input to anode and screen of the valve to a very low value.

A LOW POWER TRANSISTOR TRANSMITTER FOR 144 MC/S

The simple low power transistor transmitter circuit shown in **Fig. 7.28** is typical of the type likely to become popular for portable use during the next few years. Although the power output is quite low—about 15 to 20 mW—it can be raised by means of a higher power amplifier.

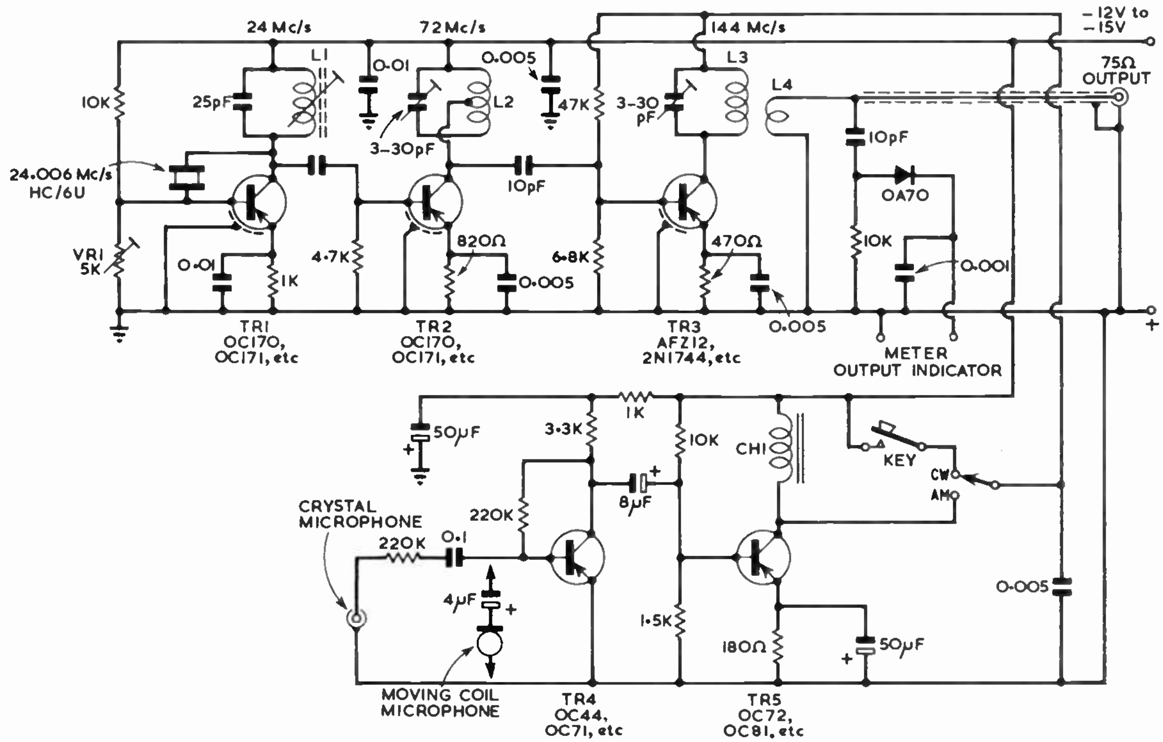


Fig. 7.28. Circuit diagram of a low power transistor transmitter for 144 Mc/s. L1, 25 turns 30 s.w.g. enam. close wound on ¼ in. diam. former with dust-iron core; L2, 8 turns 26 s.w.g. enamelled ¼ in. diam. self-supporting, centre tapped; L3, 5 turns 26 s.w.g. enam. ¼ in. diam. turn spaced one wire diameter; L4, 2 turns 26 s.w.g. enam. wound over L3 at supply end. Value of coupling capacitor between TR1 and TR2 is 47pF.

TR1 is a crystal oscillator operating at 24 Mc/s which materially simplifies the transmitter. There is of course no reason why the crystal should not be of a lower frequency but it might then be necessary to use an additional frequency multiplier. The second stage, is tuned to 72 Mc/s, TR2 operating as a tripler from 24 Mc/s. TR3 is a frequency doubler from 72 Mc/s to 144 Mc/s and feeds the aerial via coupling coil L4.

Modulation is provided by TR4 (speech amplifier) and TR5 (Heising modulator). It is important that the audio choke CH1 has a low resistance.

The power output may be raised by adding several amplifying stages ending with one of the following transistors, all of which are capable of about 12 watts output at 144 Mc/s: 2N3478, 40280, 40281 and 40288. A higher power modulator would be required and it would be desirable to modulate the driver and possibly earlier stages for the best results.

SEVENTY CENTIMETRE TRANSMITTERS

The portion of the 70cm band generally occupied by crystal controlled transmissions ranges from 432 to 436 Mc/s. This has been chosen because it is in harmonic relationship with the 2m band, and thus an existing 2m transmitter may be used to drive a tripler stage to give output in the 70cm band without any duplication of existing equipment. In practice the majority of crystal controlled transmissions take place in the range 432 to 434 Mc/s.

Although some transmitters make use of a tripler stage only following the 2m transmitter or exciter, the majority consist of a tripler followed by a straight amplifier on 70cm. In addition to the increased efficiency obtained from an amplifier as compared to a frequency multiplying stage, its use is desirable to minimize radiation at 145 Mc/s which results if a tripler stage feeds the aerial directly.

Low Power 70cm Transmitter

The transmitter to be described is a typical low power transmitter for 70cm using double tetrodes in the final triple and amplifier stages.

The complete circuit given in Fig. 7.29 shows the exciter stages and tripler/amplifier as two separate units sharing a common h.t. supply. The tripler/amplifier may be used to follow an existing 2m transmitter giving between 1 and 2 watts output.

Circuit Description

The exciter consists of three stages, V1, uses a triode-pentode, type ECF82 (6U8), the triode section of which operates as a " Squier " oscillator with a 24Mc/s overtone crystal while the pentode section is used as a multiplier tuned to 72 Mc/s.

The second multiplier V2 uses a high slope pentode type E180F (6688) with its anode circuit series tuned to 144 Mc/s the coupling to the final stage V3, a QV03–12(5763) is connected through a second capacitor to reduce the effect of the input loading of V3 from excessive damping of the tuned circuit. Alternatively the drive may be taken from a suitable tapping point on L3.

V3 operates as an amplifier on 144 Mc/s, the anode circuit is series tuned, with anode supply connected to the centre of

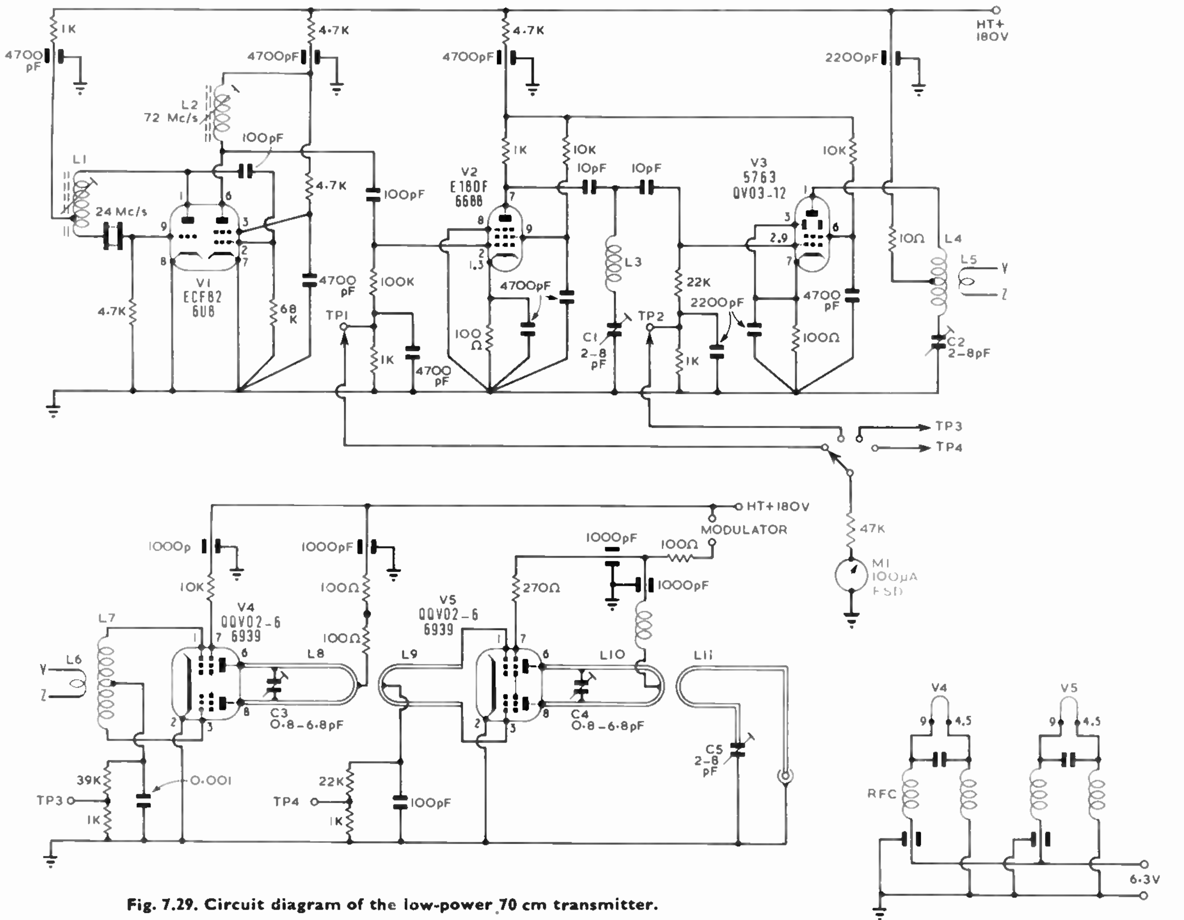


Fig. 7.29. Circuit diagram of the low-power 70 cm transmitter.

L4 ideally the value of the tuning capacitor should be equal to the output capacitance of the valve when hot. Coupling between V3 and V4 is provided by inductors L5 and L6 both of which are single turn coils.

In a complete unit this coupling should take the form of a link coupling, with neither side connected to earth.

If however, the tripler amplifier is built as a separate unit, then the input from an existing transmitter or exciter is most likely to be fed through a coaxial socket to the coupling coil. The use of screened twin feeder would assist in preserving a balanced input to the push-pull tripler, but single core coaxial cable can of course be used.

Both the V4 tripler and V5 amplifier stages use low power double tetrodes type QQV02–6 (6939) in a push-pull arrangement. V4 is operated at a considerably higher grid bias voltage than the amplifier to give good harmonic performance. The amplifier V5 will, under the operating conditions given, provide an output in the region of 5–6 watts when the h.t. supply is increased to about 220 volts.

All the circuits operating at 70cm are of strip line type and it is important that these are made mechanically similar to the details given in the drawings Fig. 7.32.

Grid bias for each of the stages is obtained from the grid current, through the individual grid resistors. Provision is made for a test point on each of these resistors for tuning and test purposes. If desired each test point may be connected to a meter through a suitable selector switch if this facility is required. Alternatively the test points, which are brought through feedthrough capacitors or insulators can be metered from above the chassis.

Output coupling is through a conventional stripline coupling loop with a capacitor in series with the earth lead, to a coaxial socket.

Construction

The complete transmitter can be built on a chassis 16 in. × 4 in. × 2 in. deep. If only the tripler amplifier section is required for use with an existing 144 Mc/s exciter or transmitter, this can be built on a chassis 9 in. × 4 in. × 2 in. deep, see Fig. 7.31.

The layout of the complete unit is shown in diagram Fig. 7.30. It will be seen that the prototype crystal was mounted in an octal based oven, this of course may be replaced by a standard type of crystal and socket but a lower standard of frequency stability must be expected.

The oscillator and the first frequency doubler inductors L1 and L2 are wound on 7·5mm diameter formers and mounted above the chassis. Standard screening cans are fitted to both coils.

Both L3 and L4 are self supporting and tuned by 2–8 pF

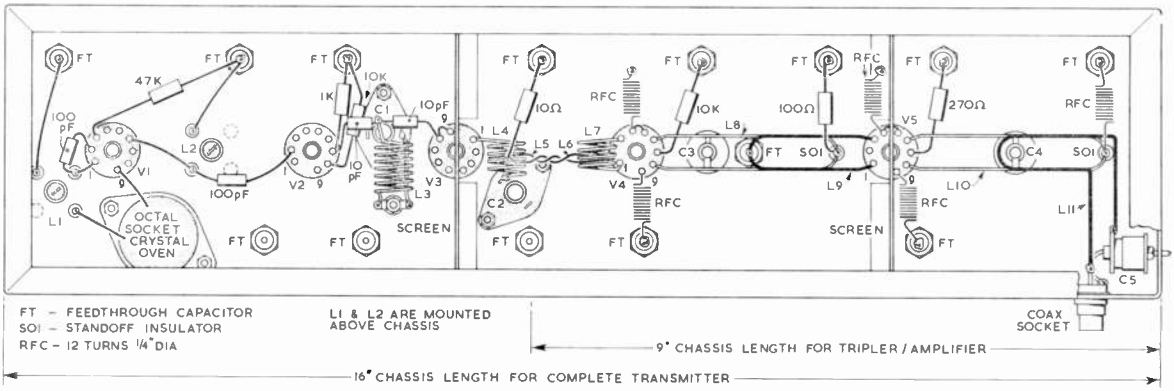


Fig. 7.30. Under-chassis layout of 70 cm transmitter.

pot type Philips trimmers. The trimmers are mounted below holes in the chassis so that they may be adjusted from above. One of the standard hexagonal trimming tools should be used for adjustment, alternatively, a hollow hexagonal insulating material is marketed by the trimmer makers.

The coupling coil L5 is a single turn of insulated wire and should be coupled into the centre of L4. This coupling coil is connected directly to a similar input coil in the grid twin-coil L7 of the 144 to 432 Mc/s tripler V4.

The anode circuit of V4 is tuned to 70cm and consists of a " U " shaped stripline using ¾ in. wide strip bent to ½ in. between sides. The dimensions of this circuit are shown in Fig. 7.32. As shown the tuning capacitor is soldered by its solder tags to a point 1⅝ in. from the closed end of the line.

The tuning capacitors C3 and C4 are both Philips type COO4/EA. This type has been chosen because it is more convenient to fit into the circuit than older versions and it also has a lower minimum capacitance.

To support this circuit which is mounted ½ in. below the chassis, the closed end of the strip loop is fixed to a ½ in. ceramic stand-off insulator as shown in the side view general arrangement Fig. 7.32.

The amplifier V5 grid circuit is mounted, ¼ in. below the tripler anode circuit and is a similar strip line loop made of ¼ in. wide material. No tuning is provided but small adjustments to the inductance are possible by adjusting the solder points and thus varying the length of the loop. The connection from centre point at the closed end of this loop is fixed to a feed through capacitor which makes it rigid and at the same time provides the connecting point for the bias resistor.

The anode circuit of V5 is very similar to that used for the tripler anode circuit, differing only in the connecting points for the tuning capacitor which in this case are made 1¼ in. from the closed end.

The output coupling L11 is another strip loop of ¼ in. wide strip mounted ¼ in. below the anode circuit. It is shaped to connect to the co-axial socket mounted on the side wall of the chassis at one end and the series pot type Philips trimmer at the other. Fig. 7.30 shows this quite clearly.

Screens are fitted across the chassis and are located centrally across the valve sockets V3 and V5. They should be fixed to the main chassis by means of screws each side of the valve socket. A suitably shaped cut out must be made for them to fit closely over the valve socket.

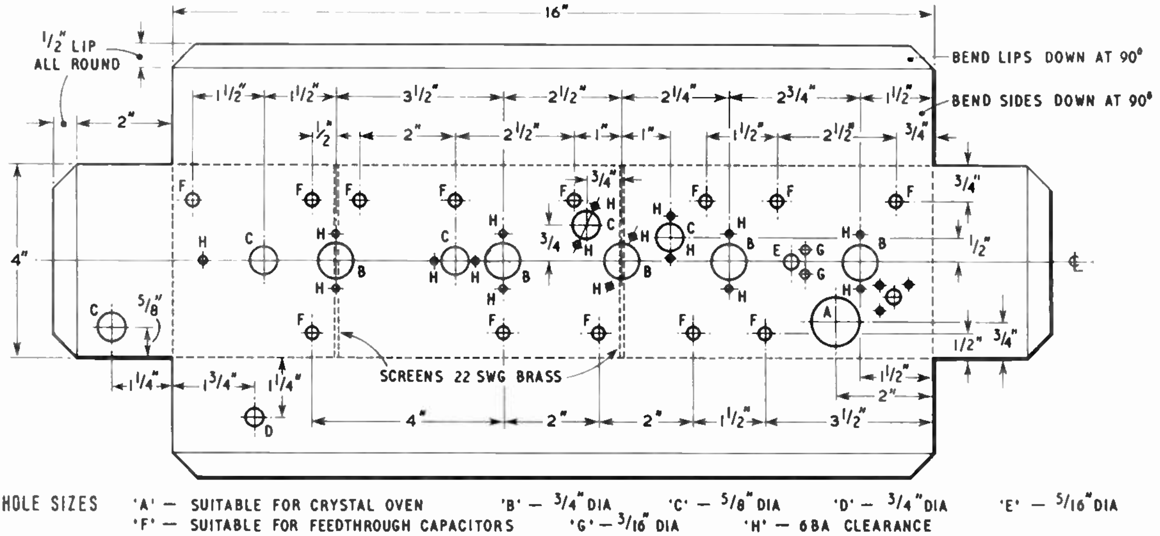
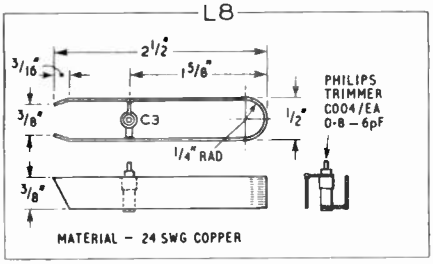
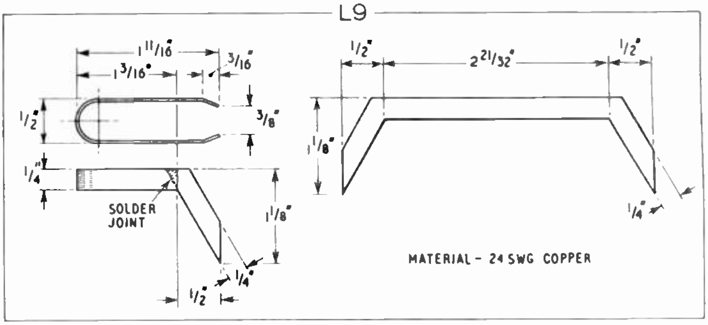
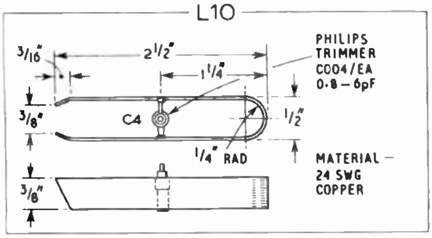
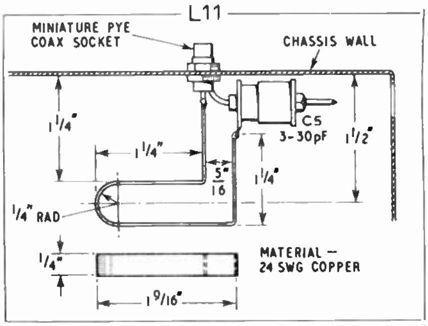


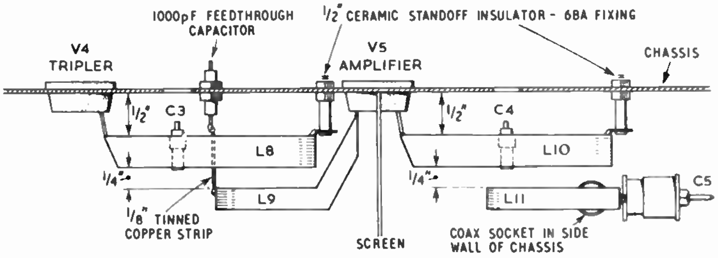
Fig. 7.31. Drilling and bending diagram for low power 70 cm transmitter.











SIDE VIEW SHOWING POSITIONING OF INDUCTORS

Fig. 7.32. Constructional details of 70 cm transmitter stripline inductors. Positioning of tuning capacitors is also critical and should be closely followed.

Adjustment and Operation

With the crystal and V1 and V2 in position, the screen of V2 should be temporarily open circuited. Adjust both L1 and L2 for maximum output as indicated by a meter connected to TP1 in the grid circuit of V2. A check should be made to see that the crystal controls the oscillator and the valve is not self-oscillating. A check should then be made to ensure that L2 is tuned to 72 Mc/s.

The succeeding stages should be adjusted using the following stage valve with its screen disconnected as a tuning indicator. This procedure is particularly useful when the link coupling between V3 and V4 is being set up.

Checks should also be made to ensure that neither V3 or V5 are self-oscillating.

Preliminary output testing can be carried out using a 12 v. 0·3 amp lamp as a load. When plugged into the output socket it should light up to full brilliance when the transmitter is correctly adjusted. It should be noted however that the value of C5 will be different when set up with a lamp load, to that which is required for a normal co-axial feed line to the aerial.

The use of an r.f. volt meter as shown on page 7.52 or a SWR Bridge, reading forward power is recommended, to enable the proper adjustment to be made when feeding into the aerial line.

With the meter (100μA full scale) and its series 47 K resistor, the grid metering is by measurement of the voltage across the resistors connected to the various test points.

Typical values are:

TP1	grid of	V 2	0·5 V
2		V 3	1·0 V
3		V 4	3·0 V
4		V 5	0·8 V

Alternatively a simple 0·5 mA meter may be used to read current. In this case some small change in bias will occur when the meter is connected across the resistor concerned.

Typical anode and screen currents for the 70cm Tripler and Amplifier are shown in Table 7.9.

TABLE 7.9

Valve	Grid Current	Screen current	Anode current
V4	20·0 mA	5 mA	26 mA
V5	1·25 mA	9 mA	40 mA

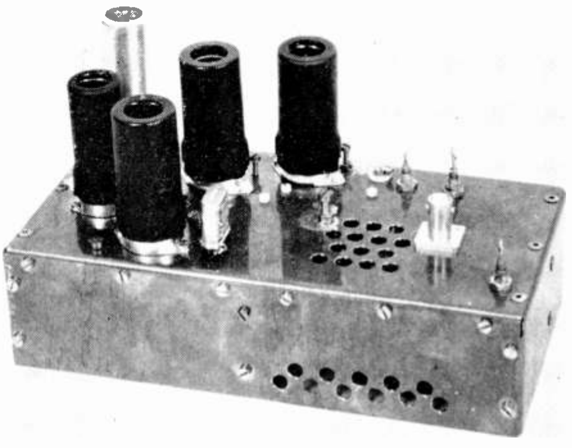
Note that the input power to the tripler is much lower than that to the amplifier and the proportions given should be readily obtained. If there is a tendency for needing higher inputs to this stage the trouble can be corrected by improving the circuit efficiency—this is achieved by reducing the coupling between V4 anode and V5 grid circuit. Similarly the output coupling should not be too close.

A 15 WATT TRANSMITTER FOR 430 Mc/s

The transmitter shown in Fig. 7.33 employs grounded grid final multiplier and power amplifier stages and is capable of providing r.f. power outputs of 15 watts c.w. or 10 watts phone. It can be used as the r.f. section of a complete 430 Mc/s transmitter or as a driver for a higher power stage such as a 4X150A or 4CX250B.

Circuit

The crystal controlled multiplier chain uses three valves to



The complete 430 Mc/s transmitter. Components may be identified by reference to Fig. 7.34.

produce 750 mW at a frequency of 144 Mc/s from an 8 Mc/s crystal oscillator. Series tuning of the N78 (V3) anode allows more latitude in the construction of the coil than would be possible with parallel tuning. The anode coil of the N78 and the cathode coil of the A2521 tripler (V4) are coupled by a single turn of p.v.c.-covered wire pushed into each coil. The anode coil of the A2521 trebler and amplifier stages are identical single turn loops and each is coupled to the following cathode by an untuned loop adjacent to the anode coil. The DET24 (V6) anode circuit consists of a square loop from which the output is taken through a coupling capacitor close to the anode. All three grounded grid stages employ cathode bias; as the heater and cathode of the DET24 are common, this necessitates a separate 6·3 volts heater supply for this valve.

Since some of the power output reaching the aerial is fed through from the A2521 amplifier, it is necessary to anode modulate both the DET24 and the A2521 amplifier (V5) and

TABLE 7.10

COMPONENT INFORMATION FOR FIG. 7.33

* *CT1, 2, 3/8 in. length of 1/8 in. bore copper tube soldered to anode loop. The tube is internally lined with 1½ turns of 0·005 in. p.t.f.e. tape; an earthed 6BA screw is inserted for tuning.
* *CT3, ½ in. length of 3/16 in. bore copper tube soldered to anode mount. The tube is internally lined with 3½ turns of 0·005 in. p.t.f.e. tape, and an earthed 4BA screw inserted.
* *CT4, 3/16 in. length of ¼ in. bore copper tube mounted on a 3/16 × 3/16 in. copper bracket and soldered to the anode mount. The tube is internally lined with 2½ turns of 0·005 in. p.t.f.e. tape, and a 3/16 in. copper slug threaded on 6BA studding inserted. The studding is soldered directly to the centre pin of a chassis mounted BNC socket.

L1, 26 turns, 26 s.w.g. enam., 3/8 in. diam. former with dust core.
L2, 15 turns, 20 s.w.g. enam., 3/8 in. diam. former with dust core.
L3, 6 turns, 20 s.w.g. enam., 3/8 in. diam. self-supporting (close wound on 3/8 in. diam. mandrel).
L4, 5 turns, 16 s.w.g. bare copper, ½ in. diam., air-spaced to 5/16 in.
L5, 2½ turns, 20 s.w.g. enam., 3/8 in. diam., air spaced to ½ in.
L6, 8, 1 turn, 10 s.w.g., ¾ in. diam., with one end attached to a 1 in. × ¾ in. 20 s.w.g. copper plate to form a capacity of 150 pF to earth using 0·001 in mica dielectric.
L7, 9, 1 turn, 16 s.w.g., enam., 3/8 in. diam.
L10, Square form single turn loop of 1 in. × 1⅛ in. internal dimensions with one end integral with anode mounting and the lower horizontal surface spaced from the chassis with 0·001 in. mica. The loop is fabricated from 1¼ in. width × 16 s.w.g. copper strip.
RFC1, 2, 3, 26 s.w.g. enam. wound to cover a 100 K ohm Erie type 8 resistor.
RFC4, 5, 11 turns, 26 s.w.g. enam., 3/16 in. diam., self supporting.
RFC6, 7, 8½ turns, 20 s.w.g. enam., 3/16 in. diam., air spaced to ½ in.

* Commercial types available in Mullard COO4/EA range.

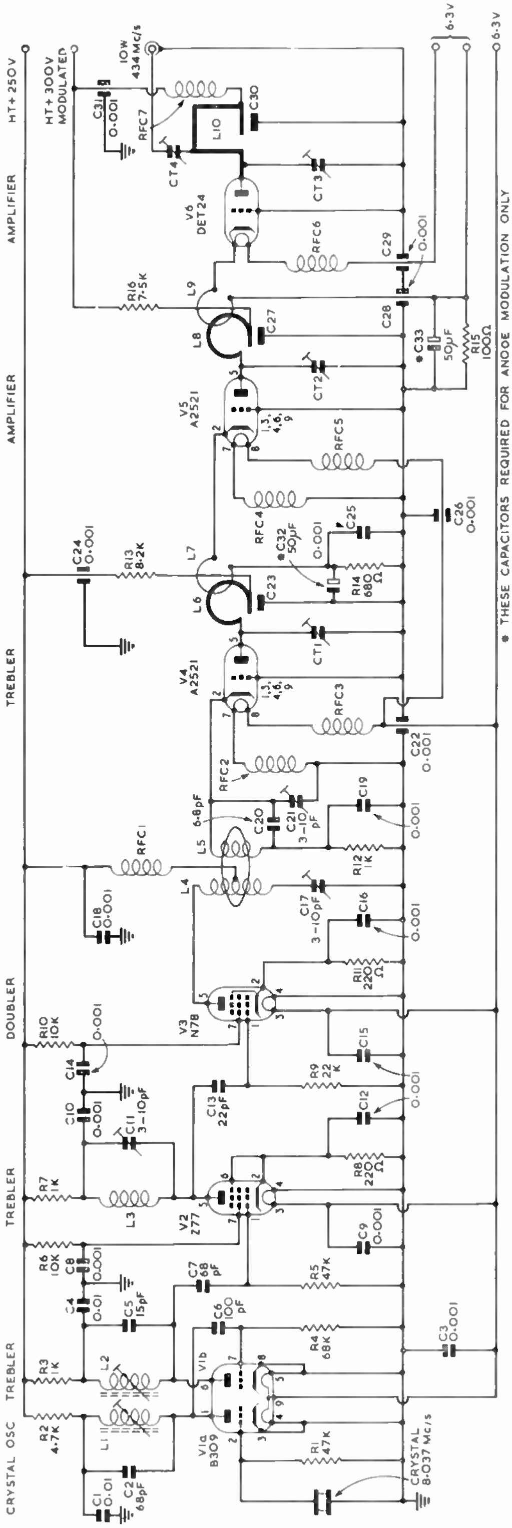


Fig. 7.33. Circuit diagram for the 430 Mc/s 10 watt transmitter. An alternative valve for V1 is the 12AT7, and for V2 the EF91 or 6AM6.

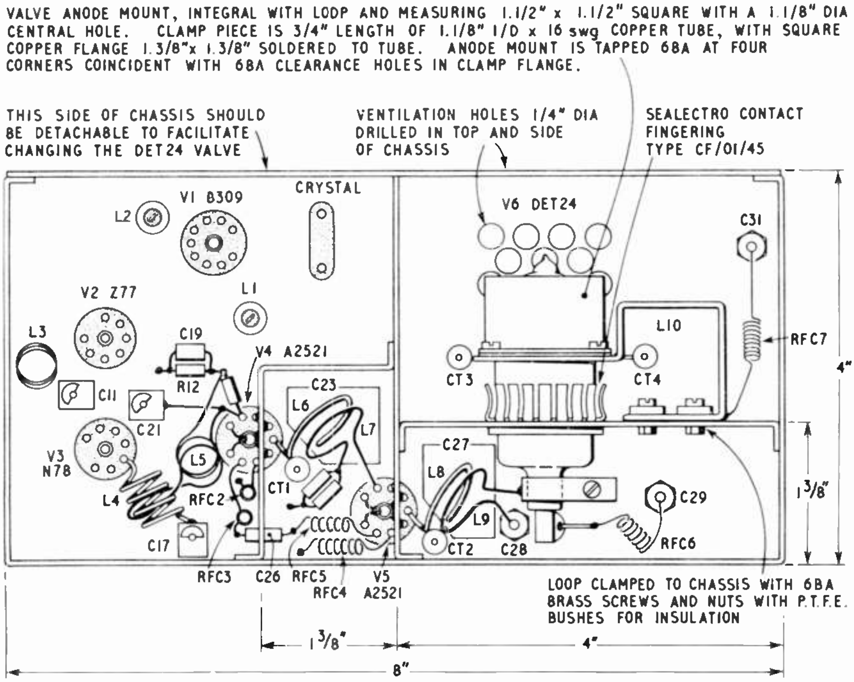
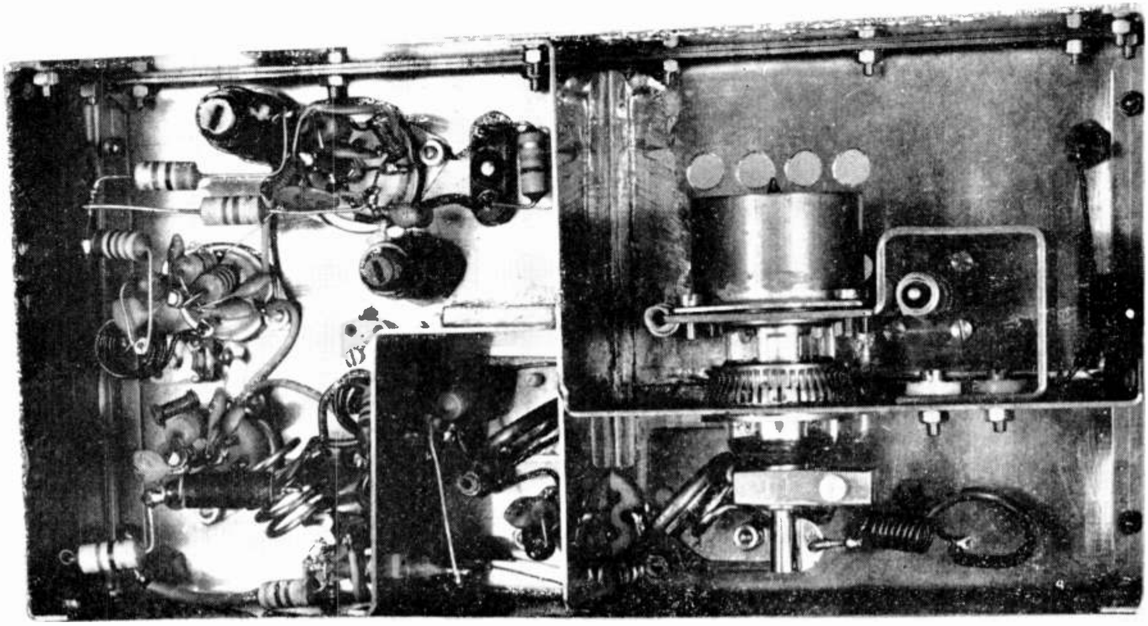


Fig. 7.34. Layout of the underside of the complete transmitter, showing the positions fcr coils, associated capacitors and resistors.

to bypass the cathode resistors of both stages to audio as well as radio frequencies. The anode supply to the A2521 from the modulated h.t. line should include a resistor to reduce the voltage to 160 volts at 18 mA. If anode modulation is not applied, then the A2521 may be fed directly from the 250 volts h.t. line and the DET24 anode supply may be increased to 400 volts. The A2521 anode current should never exceed 18 mA, nor the grid current 6 mA.

Construction

The crystal controlled multiplier is conventional and simple to construct. Copper screens are soldered in the positions shown in Fig. 7.34, and pins 1, 3, 4, 6 and 9 of the two A2521 valveholders should be soldered directly to the appropriate screen. The anode circuits of the A2521 tripler and amplifier are identical. A length of 10 s.w.g. (⅛ in. diameter) copper ware is bent around a ¾ in. mindrel; one end is straightened over a length of approximately ¼ in., and this is soldered to the top surface of a small copper plate measuring ¾ in. × 1 in.; the other end of the loop is cranked to allow connection to the anode pin of the valve base and at the same time allow a small copper tube (CT1 and CT2) to be soft soldered to the loop as near as possible to the anode but projecting over the chassis, clear of the valveholder. A



An under-chassis view of the transmitter. The main components may be identified by reference to Fig. 7.34.

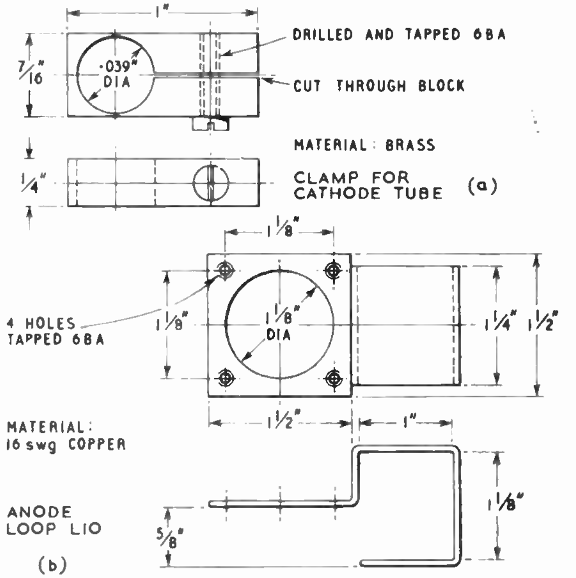


Fig. 7.35. (a) Cooling clamp for the cathode of V6. (b) Method of bending metal to form L10, the anode loop for V6.

hole is then drilled and tapped 6B.A. through the top of the chassis in line with the axis of the copper tube; a 6B.A. brass nut is soldered on to the top surface of the chassis, over the tapped hole, to give increased stability to the 6B.A. screw which forms the tuning control.

The copper plate which forms the bypass capacity to earth (C23 and C27) should be " tailored " to provide clearance of the valveholder and tuning control. The plate is insulated by a thin mica sheet and fixed to the chassis by two 6B.A. nylon screws (Radiospares). It is necessary to countersink the 6B.A. clearance holes on the underside of the chassis to prevent voltage flashover around the edges of the screw holes. In the case of the A2521 amplifier, one nylon screw is replaced by a brass screw and p.t.f.e. bush (Polypenco, Welwyn Garden City) in the chassis to provide an external connection for the anode supply to that stage.

The DET24 anode circuit is made from a strip of 16 s.w.g. copper (Fig. 7.35). The anode mount is 1½ in. wide and the strip is trimmed over the length of the loop to 1¼ in. wide. The anode flange of the valve seats on to the anode mount and the anode clamp, which acts as an additional heat sink, is fixed to the mount by 6B.A. screws at each corner. It is essential to keep the temperature of the valve anode seal below 140°C and it is necessary to provide good heat conduction to the chassis; the integral mount and loop of copper together with the use of thin mica for insulation, meet this requirement. Nylon screws are not suitable for clamping the loop to the chassis screens due to the temperature of the chassis, so that it is necessary to use p.t.f.e. bushed holes and 6B.A. brass screws and nuts. The grid contact is formed by soldering contact fingering directly to the chassis screen and arranged to give firm contact with the grid ring of the valve.

A clamp on the cathode tube of the valve assists in keeping it cool and provides a convenient anchorage for the coupling loop.

It is essential to provide a well-fitting base plate to the chassis in order to prevent direct radiation from the output circuit. One side of the chassis should be made removable so that the anode clamp of the DET24 can be unscrewed if the valve has to be changed.

TABLE 7.11

Typical performance figures for the DET24 output stage of Fig. 7.33.

Class C Telegraphy			
Va	350V	300V	250V
Ia	92mA	81mA	73mA
Ig	38mA	39mA	40mA
P load	15·0W	12·8W	10·3W
Class C Telephony anode modulated (carrier conditions)			
Va	300V		250V
Ia	77mA		68mA
Ig	30mA		34mA
P load	10.9W		8·8W
P mod	11.6W		8·5W
Z mod	3·9 K ohms		3·7 K ohms

A 24 WATT TRIPLER-AMPLIFIER

The illustrations show a higher power 70cm tripler-amplifier having an r.f. output of not less than 5 watts. It is intended for use in conjunction with a 2m transmitter. With an input of 24 watts to the amplifier stage this unit will deliver a power output of 12 watts at 435 Mc/s. Mullard QQV03-20A (6252) double-tetrode valves are employed in both tripler and amplifier stages.

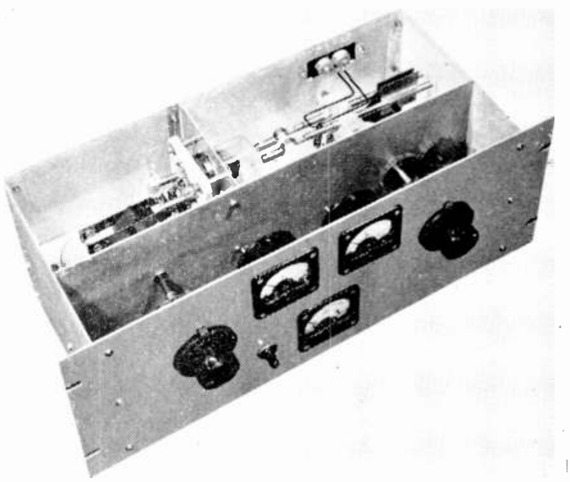
The unit requires an h.t. supply of 300 volts at 200 mA and a 230 volts a.c. mains supply; the latter is necessary to obtain 6·3 volts for the valve heaters, for which purpose a transformer is mounted underneath the chassis.

All the 70cm tuned circuits are formed by parallel lines which are shielded to minimise radiation and thus increase their efficiency. Forced-air cooling of the valves is unnecessary unless the amplifier stage is run at an input greater than 24 watts. Anode and screen modulation of the p.a. stage may be effected by a modulator having an audio output of 15 to 20 watts; with an input of 24 watts to the power amplifier the impedance of the modulated stage is 3750 ohms.

Circuit Description

The circuit of the tripler-amplifier is shown in **Fig. 7.36.** R.F. drive from the 2m exciter is applied to CS1 and via L1, L2 to the push-pull grids of the tripler valve V1. L2 is tuned by the valve input capacitance and stray circuit capacitances to a frequency of approximately 145 Mc/s; coupling between L1 and L2 is variable and can be adjusted to obtain the rated grid current for V1. Grid-leak bias for the tripler valve is derived from flow of grid current through R1, which is returned to chassis through the monitoring resistor R2, bypassed by C1.

The anode circuit of V1 is tuned to 435 Mc/s and consists of a quarter wave parallel strip line L3 tuned by C2 which consists of the capacitance between opposite legs of the strip line. This capacitance is made variable by gradually introducing a block of insulating material, in this instance Keramot, between the line, so that the permittivity of the dielectric and hence the effective capacitance of the circuit is changed. This method of tuning, which is also employed in the p.a. anode circuit has been adopted to maintain balance between each section of the valve. It has commonly been found that the



Front view of the 24 watt tripler-amplifier. The tripler tuning is on the left and the p.a. tuning on the right. The anode meters M2 and M3 are at the top of the panel.

more normal method of tuning push-pull circuits by using a split-stator capacitor having the rotor either earthed or left floating, results in unbalance of the circuits when employed at frequencies of the order of 435 Mc/s. A material such as polystyrene having a lower power factor than Keramot, may be used but even with the less efficient dielectric employed the r.f. drive to the amplifier grids is more than adequate.

H.T. supply to the anode of V1 is decoupled by RFC1 and the feed-through capacitor C3. The screen supply is reduced from 300 volts to a maximum of 250 volts by R3, the value of which should be adjusted during initial setting-up of the unit so that maximum r.f. output is obtained from the valve when operated within the permissible rating. The screen of V1 is un-bypassed.

Inductive coupling is employed between V1 anode circuit and the grid circuit of V2. The tuned grid circuit L4 of the power amplifier is a half-wave long when loaded by the input capacitance of the valve, and RFC2 and RFC3 are connected at the approximate electrical centre of the line. The two grid circuit chokes are returned to earth through the bias resistor R4 and monitoring resistor R5 in series; these resistors are bypassed by C5, C6 and C7.

Another half-wave line L5, tuned by C9 forms the anode circuit of V2. In this instance the anode tuning capacitor is formed by the capacitance between two copper plates soldered to the line; the dielectric of this capacitor in the transmitter illustrated is also of Keramot but it is recommended that polystyrene be used so that the highest possible efficiency is obtained in this circuit. H.T. is fed to the p.a. anodes through RFC4 and RFC5, connected to the line at the voltage node. The anode circuit is decoupled by C8, which is specially constructed to provide a low impedance bypass at the operating frequency. As with the tripler circuit, the screen of V2 is un-bypassed.

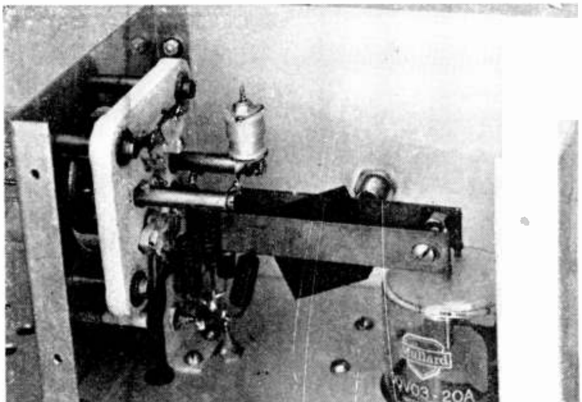
Anode and screen modulation of the power amplifier is obtained by connecting the output of the modulator to SK1. A screened lead between modulator and p.a. may be employed, with the screen connected to the chassis of the r.f. unit at SK1 if necessary.

A balanced output loop L6 is coupled to L5 adjacent to the point corresponding to the voltage node on the line, and is connected to two polystyrene feedthrough insulators mounted on a strip of paxolin fixed to the rear screen. The paxolin strip ensures a low capacitance between the insulators and the earthed screen. If a coaxial cable is used to feed the aerial array, it is recommended that a balance-to-unbalance transformer be interposed between L6 and the cable; the coupling loop shown is suitable for this purpose.

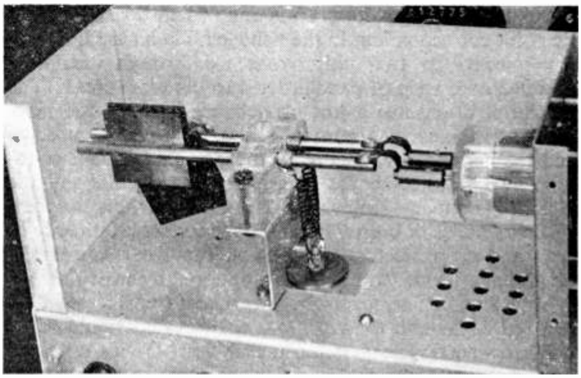
The cathodes of both V1 and V2 are connected directly to earth and no protective grid bias is employed. Current in the grid circuits of V1 and V2 is monitored by switching M1 across R2 and R5 respectively. The heater transformer T1 is included in the unit to avoid voltage drop that would occur in the connecting lead if the heater voltage was obtained from a separate power unit. For the same reason connections between the secondary of T1 and heater tags on the valveholders should be made with a heavy gauge conductor, preferably not less than 16 s.w.g. For efficient operation of the valves it is essential that the voltage across the actual valve heaters should be maintained at 6·3 volts. The heaters of V1 and V2 are bypassed at the valveholders by C10 and C11 respectively.

Construction

Views of the tripler and amplifier stages, and the underside of the chassis are amply illustrated, chassis drilling details are shown in **Fig. 7.37**. Panel, chassis, front and rear screens are made from 16 s.w.g. aluminium; the side screens and vertical shield on which the tripler valveholder is mounted are of 18 s.w.g. aluminium. The panel fitted to the transmitter illustrated measures 17 in. long × 7 in. high, but the length may be increased to 19 in. if it is desired to mount the unit in a standard rack. The U-shaped chassis on which the two stages are mounted measures 15 in. long × 4 in. wide × 2½ in. deep, and the side screens are secured to it by 2½ in. × ½ in. aluminium angle brackets at each corner on the underside. When making the two side screens, the rear flange on each should be formed so that it is flush with the edge of the screen and hence with the



Close-up view of the tripler anode circuit and p.a. grid circuit. Note the variable dielectric between the lines forming L3.



The anode circuit of the power amplifier.

rear drop of the chassis when fitted in position. Cheese-head 6B.A. screws and nuts are used in the assembly of the unit, with the exception of the front panel, which uses six 4B.A. round-head screws and nuts.

At the extreme left of the unit the tripler stage valveholder is mounted underneath the chassis on four spacers ½ in. long, made from ¼ in. diameter brass rod drilled and tapped 4BA. A small fibre washer is inserted between the valveholder and each spacer and under the head of each fixing screw to prevent fracture of the ceramic holder as the screws are tightened.

The input coupling loop L1 is mounted on a thin strip of polystyrene attached to a paxolin support by a small angle bracket on which it is pivoted by a single 8B.A. screw; this permits adjustment in the position of L1 relative to L2. The paxolin support is secured to the underside of the tripler valveholder by two of the 4B.A. securing screws. L1 is insulated with polythene sleeving; the ends of the wire are bent at right-angles and passed through two small holes in the polystyrene to which it is secured with cement.

Decoupling capacitors C1 and C10 are mounted on an aluminium bracket (Fig. 7.37) fixed to the underside of the chassis adjacent to the tripler stage valveholder; the cathode is connected to this bracket through a short length of ½ in. wide copper strip.

The amplifier valveholder is mounted on the interstage screen in a similar manner to that employed for the tripler valveholder. Before the holder is fitted in position it is necessary to remove the two control grid contacts so that the p.a. grid lines may be fitted directly on to the valve pins. This is done by drilling out the rivets which secure the spring contacts to the base; the holes in the base are then enlarged to a diameter of not less than 3/16 in., using a tungsten-carbide tipped drill (Mason Master or Rawlplug) with turpentine as a lubricant. Extreme care must be exercised when enlarging

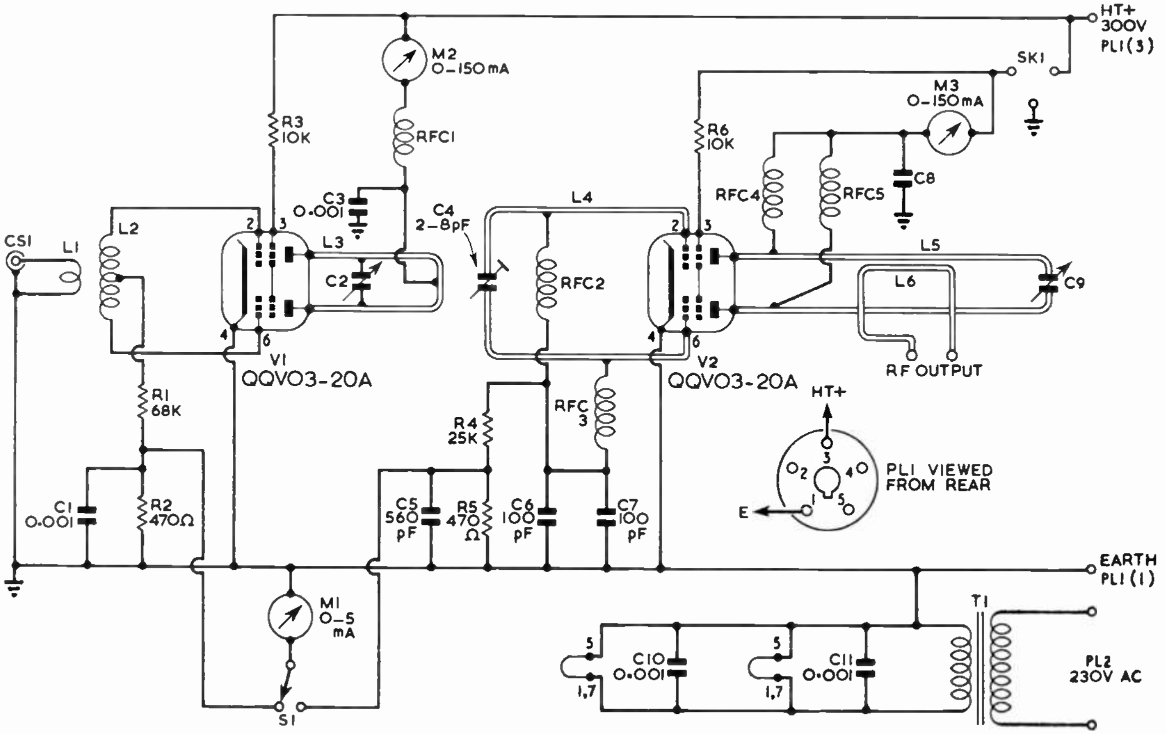
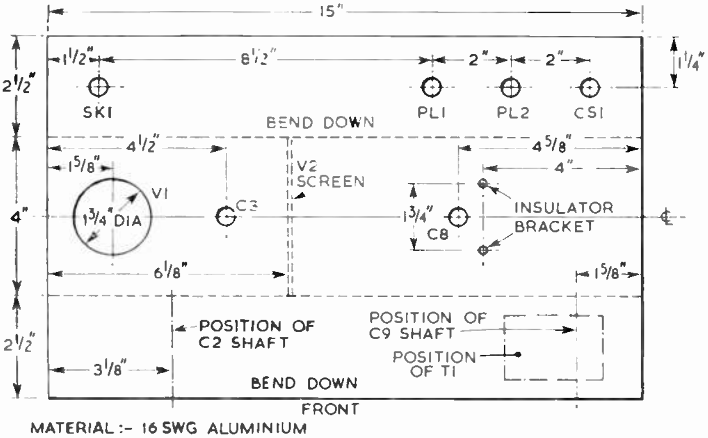


Fig. 7.36. Circuit diagram of the 24 watt tripler-amplifier for 435 Mc/s. C1, 2, 3, 11, 12, 0·001 μF ceramic stand-off; C2, C6, 7, 100 pF ceramic feedthrough; C8, approximately 220 pF (see text and Fig. 7.38(d)); C9, see text; L1, 1 turn 20 s.w.g. enam ⅜ in. o.d. illustration in text ; L2, 4 turns 16 s.w.g. enam. ⅜ in. o.d. approximately ½ in. long, centre tapped; L3, copper strip 18 s.w.g. ½ in. wide total length 6¾ in. bent into U form (see Fig. 7.39(a)); L4, two ¼ in. diam. brass rods 1⅞ in. long drilled and tapped as shown in Fig. 7.39 (b); L5, two ¼ in. diam. brass rods 5½ in. long plus anode connectors, see text and Fig. 7.39(c); L6, loop 2 in. long ½ in. wide of 16 s.w.g. enam.; M1, 0-5 mA m.c. meter; M2, 3, 0-150 mA m.c. meter; RFC1, 2, 3, 4, 5, 12 turns 20 s.w.g. enam. ¼ in. o.d. approximately 1 in. long; T1 230V 50 c/s primary. The valveholders for V1 and V2 are modified as described in the text.



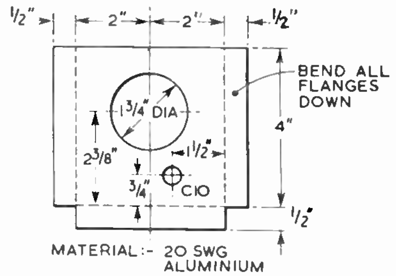


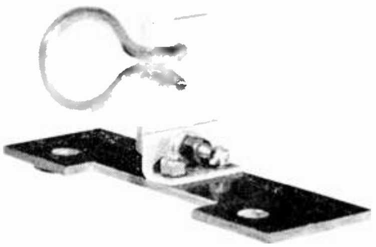
Fig. 7.37. Drilling and bending instructions for the 24 watt tripler-amplifier chassis.

the holes to avoid fracture of the ceramic base. Alternatively p.t.f.e. sachets made by Henry and Thomas can be used.

The valveholder is positioned on the screen so that the cathode contact is nearest to the chassis; this permits a short, low inductance cathode connection formed by a 20 s.w.g. aluminium bracket (**Fig. 7.38 (b)**) to be obtained. At the top of the bracket a tag is located which is soldered directly to the cathode tag on the valveholder. This bracket also carries the two grid circuit bypass capacitors C6 and C7.

Details of the mounting for the p.a. anode lines are also shown in **Fig. 7.39**. The ¼ in. diameter lines are supported between two polystyrene blocks fixed to an aluminium bracket by two 2B.A. screws and nuts; the cut-out at the base of the bracket is to accommodate the anode circuit bypass capacitor C8. To form the semi-circular slots in which the lines are seated in the blocks, clamp both blocks together, and with centres marked 14mm apart, drill through the blocks with a ¼ in. drill at the junction between them. When the blocks are fitted in the transmitter they are clamped together by a 2B.A. nylon screw (Radiospares) passing through a clearance hole in the upper block and engaging in a corresponding tapped hole in the lower one. A nylon screw is used to avoid any effects due to the presence of metal in the r.f. field between the two lines.

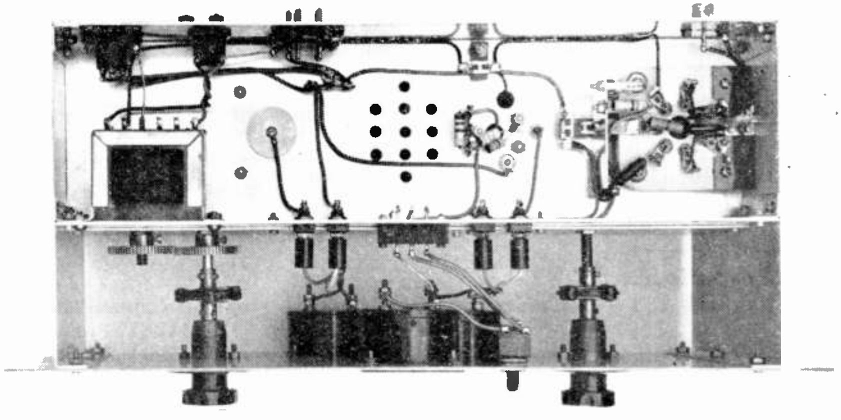
Constructional details of the anode circuit bypass capacitor C8 are shown in **Fig. 7.38 (d)**. The capacitor consists of two circular brass plates, 1¼ in. in diameter, made from 16 s.w.g. brass sheet; a heavier gauge metal may be used with



Coupling loop from the 144 Mc/s exciter to the tripler grid circuit.

advantage. One of the plates is mounted above and the other below the chassis, from which each plate is insulated by a 1½ in. square sheet of 0·004 in. mica. A 3/16 in. diameter hole is drilled in the chassis to clear the 2B.A. screw by which the plates are clamped together and a paxolin washer

Under-chassis view. The tripler stage is on the right. The transformer at the left is for the heaters.



having a thickness corresponding to the gauge of metal used for the chassis, which in the present instance is $\frac{1}{16}$ in. The two active surfaces of the brass plates should preferably be faced in a lathe to ensure their flatness. A capacitor made as described has a capacitance of approximately 300 pF.

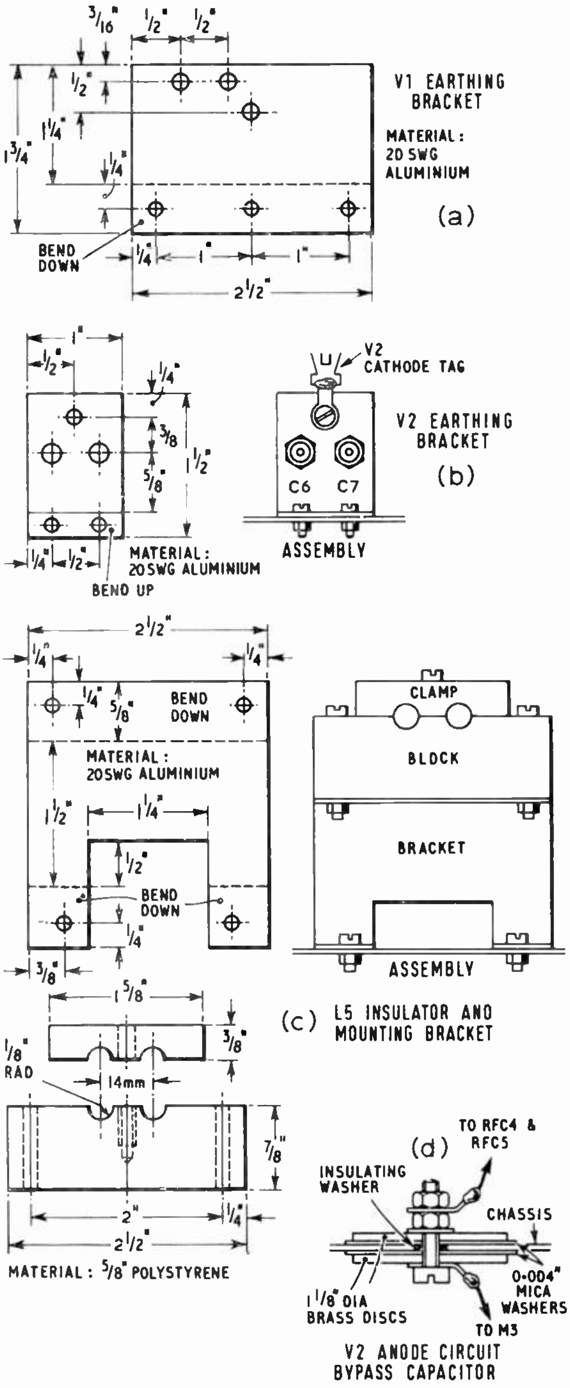


Fig. 7.38. Constructional details of mechanical sub-assemblies.

Details of the 435 Mc/s tuned circuits are shown in **Fig. 7.39.** The tripler anode line is attached to the valve anode pins by two small spring clips made from 24 s.w.g. beryllium copper secured to the line with 4B.A. nuts and screws. The block of dielectric used for tuning the circuit measures 1 in. square × $\frac{3}{8}$ in. thick, and has a $\frac{1}{4}$ in. diameter hole drilled at a distance of $\frac{1}{4}$ in. from the bottom edge; the $\frac{1}{4}$ in. diameter Keramot spindle is filed down slightly to be a push fit in this hole.

Across one end of the amplifier grid lines the tuning capacitor C4 is mounted on soldering tags secured to the line by 6B.A. screws; the other end of the line is pushed over the grid pins of the valve and secured by 6B.A. grub screws. In the unit illustrated the grid circuit chokes RFC2 and RFC3 are connected to soldering tags which are attached to the lines by 8B.A. screws; a better method is to use small anode top cap connectors similar to those used to connect the anode circuit chokes in this transmitter.

Each of the amplifier anode lines (L5) is joined to the valve anode connector through a $1\frac{1}{4}$ in. length of thin copper strip $\frac{1}{4}$ in. wide. This provides a flexible connection between the rigidly mounted lines and the valve anodes to prevent fracture of the valve envelope due to unequal expansion of the glass and metal. Considerable dissipation of heat from the anodes is effected due to the use of heavy connectors fitted to the anode pins. The plates of the tuning capacitor C9 measure $1\frac{1}{2}$ in. square and are made from 16 s.w.g. copper soldered at a centre distance of $1\frac{1}{4}$ in. from the end of the line. The variable dielectric is $\frac{1}{8}$ in. thick and $1\frac{1}{2}$ in. square and is fitted to the insulated control spindle in a manner similar to that employed for the tripler stage. Anode chokes RFC4 and RFC5 are attached to the line through two small valve anode top cap connectors; this enables the point of connection of the chokes to the line to be adjusted.

Both tuning controls are fitted with slow motion dials; the tripler stage capacitor is driven directly, but the amplifier capacitor is driven from the slow motion head through two $1\frac{1}{2}$ in. gear wheels, which enables a symmetrical front panel layout to be obtained.

In the unit illustrated the two anode current meters are connected into circuit through plugs and sockets mounted on the front drop of the chassis. This is not essential but prevents leakage of air if the transmitter has to be blown. The heater transformer T1 is mounted underneath the chassis at the extreme right hand end.

Adjustment and Operation

The output of a 2m transmitter or exciter should be coupled to CS1 through a short length of coaxial cable, and a 230 volts a.c. mains supply connected to PL2. The tripler and power amplifier tuned circuits should then be adjusted in that order by adopting the following procedure:

(a) Set the meter switch S1 to read tripler grid current and with no h.t. applied to the unit, adjust the turns spacing of L2 for maximum grid current in V1. Adjust the coupling between L1 and L2 or the output of the 2m exciter to obtain optimum current (see **Table 7.12**).

(b) Switch S1 to read p.a. grid current, and apply approximately 200 volts h.t. to the tripler stage only. (The most convenient method of doing this is to leave SK1 open circuited.) Then tune C2 and C4 to resonance as indicated by maximum grid current in V2. At resonance about half the effective capacitance of C2 should be in circuit;

(c) Connect a dummy load, which may consist of a 230 volts 15 watts lamp, to the output terminals of the transmitter. Apply h.t. to both stages, then tune C9 and adjust the coupling between L5 and L6 for maximum brilliance of the lamp load. It will generally be found that quite a loose coupling is required.

An aerial may now be connected to the output terminals and 300 volts h.t. applied. It will be found that the anode currents of V1 and V2 do not dip as the associated circuits are tuned and therefore some form of r.f. output indicator will be required to show resonance of the p.a. anode circuit. The ideal arrangement is to make use of a reflectometer (see Chapter 19, *Measurements*) connected in the aerial feeder but failing this, a simple r.f. output monitor located adjacent to the feeder will be suitable.

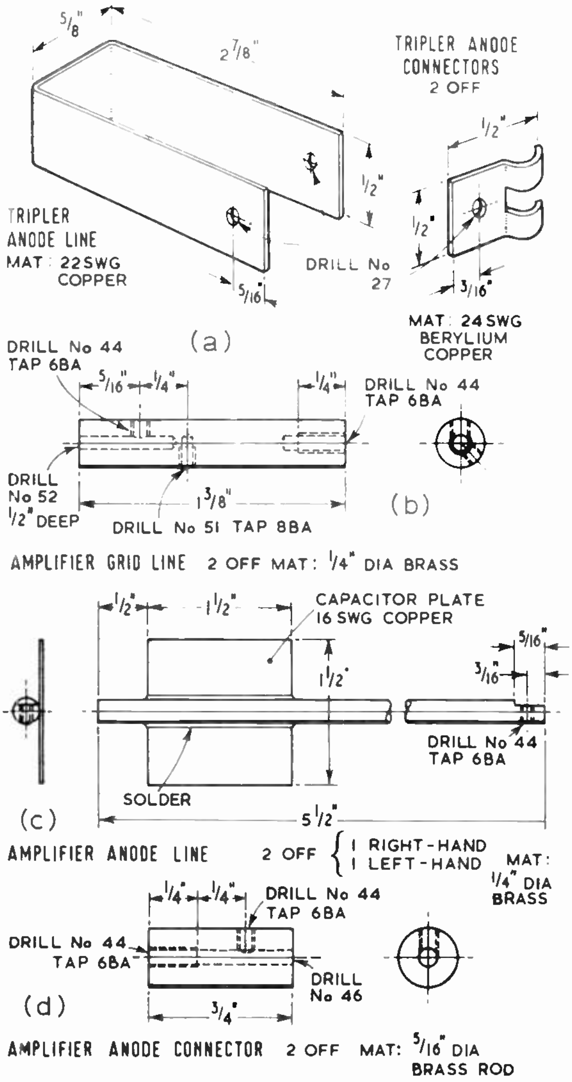


Fig. 7.39. Constructional details of 430 Mc/s tuned circuits.

TABLE 7.12

Typical current readings for the 70cm transmitter of Fig. 7.36.

Valve	Anode current	Screen current	Grid current
V1	26 mA	5·0 mA	2·5 mA
V2	80 mA	9·0 mA	2·0 mA

The value of the two screen resistors R3 and R6 and the value of the p.a. grid resistor R4 should be adjusted for optimum results within the rating of the valves. The position of RFC4 and RFC5 on the p.a. anode line should be adjusted to correspond to the voltage node on the line. A convenient way of doing this is to line up the transmitter into a lamp load, and then repeatedly tap a pencil along each line until a point is found such that when the pencil is touched on the line no diminution in r.f. output results; the clip attached to the choke is then slid along the line to this point.

Table 7.12 shows typical current readings obtained with an h.t. supply of 300 volts; it is important that the anode current of the two valves does not exceed the figures shown unless forced air cooling is employed. If the d.c. power input to either valve exceeds 24 watts it will be necessary for such cooling to be used. In this event the two stages must be completely enclosed by fitting a base and top cover to the section in which the valves are located. A motor blower can conveniently be mounted at the rear of the chassis. Holes must be drilled in the chassis beneath the anodes of the p.a. as shown in the under-chassis view to allow the passage of air past this valve, and in the top cover immediately above both valves; enough clearance for cooling purposes already exists between the envelope of the tripler valve and the chassis. If cooling is resorted to, the r.f. section must be sufficiently airtight to minimize leakage through joints between the chassis and screens; sealing may be effected by applying polystyrene cement to all joints with a small brush. The blower should have sufficient capacity to maintain the pin temperature of the valves at not more than 180°C.

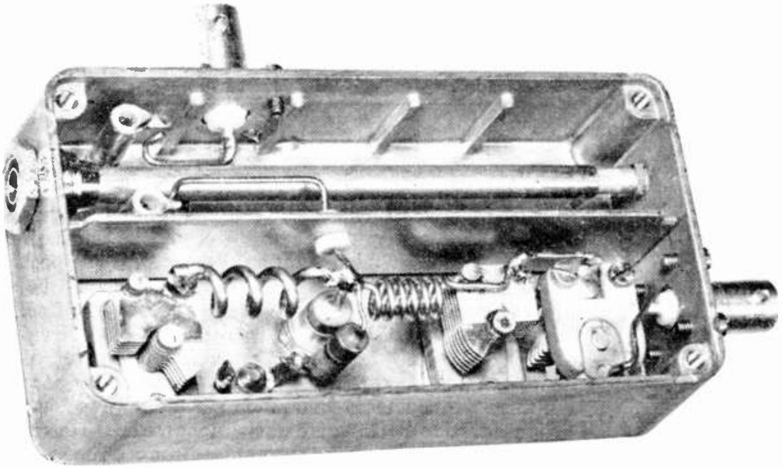
A Varactor Tripler for 70cm

This section describes a varactor tripler that will give an output of more than 10 watts on 70cm. The heart of this unit is a Mullard BAY96 varactor diode; this and all the other components are built into a small diecast box, which also acts as an adequate heat sink. Higher powers, up to about 25 watts r.f. input at 144 Mc/s can be arranged provided that additional cooling is added to prevent overheating.

Although frequency or similar modulation is the preferred method of modulation, certain types of varactor diode, of which the BAY96 is an example, will give good speech quality at the 70cm output when driven by an amplitude modulated 144 Mc/s input. It should, however, not be assumed that all varactor diodes will give an equal performance in this respect, for many will not and hence frequency or similar modulation will have to be used.

Operational tests with the circuit of Fig. 7.40 have found 144 Mc/s a.m. drive satisfactory, and this good performance has been confirmed by the manufacturers so that a tripler using this particular diode can be relied on to be suitable for either frequency or amplitude modulated drive.

It can be seen in Fig. 7.40 that the diode is used in a shunt arrangement with an auto-bias resistor connected directly



Under-chassis view of the 70cm. tripler showing positions of varactor and tuned circuits. Note position of screen between the 432 Mc/s and the input circuits.

across it. This system is convenient as it allows one side of the diode to be fixed directly to the diecast box.

The Circuit

The three tuned circuits are tuned to the input frequency, 144 Mc/s, the idler (second harmonic) frequency, 288 Mc/s, and the wanted third harmonic, 432 Mc/s; the varactor diode is connected to the junction of these circuits see Fig. 7.40. The input tuned circuit consists of C1, C2 and L1, the idler circuit is composed of C3, L2, and C4, L5, both are tuned to 432 Mc/s. As all these circuits have a common point at the diode there is naturally some interaction between them, and thus any adjustment to one circuit must be followed by readjustment of the others.

As can be seen from the photograph and layout diagram the 432 Mc/s circuit is screened from the input and idler circuits by a screen running the length of the diecast box. The connection from the common junction of the three circuits passes through the screen, via a small p.t.f.e. insulator, to the tuning capacitator C4. This circuit, comprising L3, C4, is coupled to the high *Q* circuit L4, C5, and the output circuit L5, C6 is in turn coupled to this high *Q* circuit.

Use of the high *Q* circuit enables a considerable degree of filtering to be attained and ensures that both input and idler frequencies are suitably attenuated in comparison to the wanted 432 Mc/s (third harmonic) output. Examination of the output on a spectrum analyser shows that the presence of the input (144 Mc/s) and idler (288 Mc/s) frequencies are barely noticed when compared to the 432 Mc/s signal, likewise, the outputs at fourth and fifth harmonics are very low, and the only output present at any significant magnitude is the sixth harmonic of the input frequency (twice the output frequency) at 864 Mc/s.

Construction

The construction is really self-evident from the illustrations, but the following information covers those details which may not be apparent.

(a) C2 and C3 are standard chassis fixing capacitors.

(b) C5 is made from a Mullard type COO4EA/12E tubular trimmer by removing the outer plate (fixed tube) and inserting the remaining insulator and movable inner plate into the end of the copper tube which forms L4.

(c) The input and output connections used in the prototype are BNC 50 ohm types. The pattern is not critical.

(d) Care should be taken to ensure that the inductors L1 L2 and L3 are terminated in a common point at the varactor terminal. It is convenient to include the lead of the auto bias resistor.

(e) The varactor diode should be firmly fixed to the diecast box to ensure good thermal contact.

(f) L4 is fixed to the end wall of the diecast box by a spindle lock collet, adjusted to be a tight fit wth the ¼ in. copper tube.

Alignment

The individual circuits should be adjusted for maximum power output at the wanted frequency, but any adjustment of one circuit must be accompanied by readjustment of the other circuits. Where maximum efficiency is required,

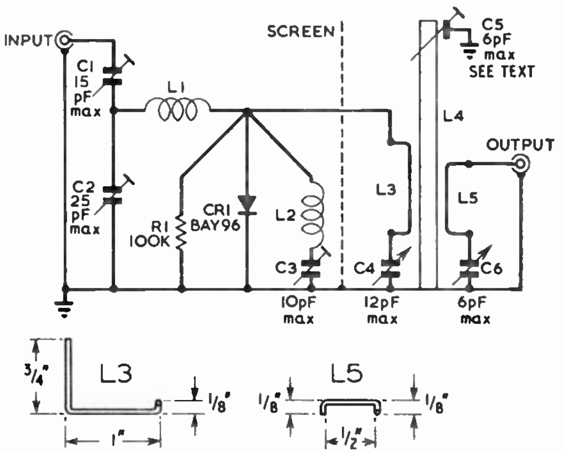


Fig. 7.40. The 70cm tripler circuit. L1, 6 turns, 18 s.w.g., 3/16 in. diam., 1/2 in. long; L2, 3 turns, 14 s.w.g., 3/16 in. diam., 3/4 in. long; L3, 18 s.w.g., shaped as shown, and spaced 3/32 in. from L4; L4, 1/4 in. in o.d., 3/16 in. i.d. copper table, 4 1/8 in. long; L5, 18 s.w.g., as drawing, spaced 3/16 in from L4.

adjustments must be made for optimum output whenever the drive power level is changed.

When correctly adjusted, an output power of 10 watts at 432 Mc/s should be obtained for a drive power of 15 watts at 144 Mc/s. The spurious output should be low enough for the unit to be connected directly to an aerial.

When adjusting the coupling between the input and output 432 Mc/s circuits and the high Q circuit, care should be taken to avoid too tight coupling which would reduce the effectiveness of the high Q circuit as a filter.

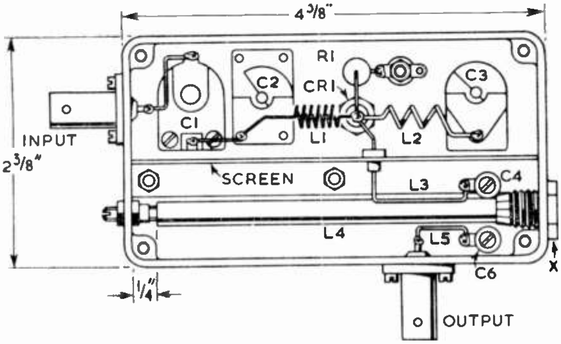


Fig. 7.41. Layout of the tripler in the diecast box. The varactor diode is bolted directly to the base.

Modulation

As mentioned earlier, it is possible to obtain good modulation output when driven by an amplitude modulated 144 Mc/s signal. The actual level of the modulation should, however, be restricted to 80 per cent to avoid the introduction of distortion.

Although the varactor diode is called in this and similar applications a multiplier—in this case a tripler—it does not, however, function in the same manner as a valve multiplier. For the purpose of understanding how this " tripler " can be used and produce good amplitude modulation, it should instead be regarded as a parametric converter. An indication of the suitability of a varactor diode to give a good audio modulation performance is indicated by the linearity of the input/output power characteristic, which in the case of the BAY96 is particularly good.

Operation

The diecast box will reach a steady-state temperature of 40-45 C when the diode is dissipating 5 watts (15 watts input, 10 watts output). If it is intended to operate the unit at a higher power level, up to the maximum rating of the diode, it may be necessary to increase the cooling area by adding fins to the box in the region of the diode fixing stud.

Conclusion

This unit makes an attractive add-on unit to enable operation on 70cm from a 144 Mc/s transmitter, for field and portable operation, or as a driver for a high power 70cm amplifier.

A varactor can be employed as a doubler, which, although unsuitable for use with an existing 2m transmitter, dispenses with the need for an idler circuit. This configuration is obviously less convenient, but there are probably several amateur applications where the improved efficiency would be advantageous.

HIGH POWER CAVITY AMPLIFIER FOR 430 MC/S

Cavity Design

The interest in using a cavity for one valve lies mainly in the simplicity of construction, since the cavity has no advantage in either efficiency or physical size over the alternative coaxial line circuit.

It may be shown that an unloaded cavity, square in shape, has a resonant frequency equal to $\frac{C}{a\sqrt{2}}$ where $C = 300 \times 10^6$ and a = length of the cavity in metres. Using this formula as a starting point, the actual cavity size when loaded with a valve is determined by " cut-and-try " methods. The cavity to be described is based on the 4X150A or 4X250B valves. Its size is fixed to resonate at 440 Mc/s, and a trimmer capacitance has been added which allows the cavity to be tuned from 430 Mc/s to 440 Mc/s, **Fig. 7.42.** A d.c. blocking capacitor is fitted to the top surface of the cavity, and includes a spring finger assembly to make contact with the valve anode. An Eimac type SK-610 air system socket is clamped by a ring to the inside lower surface of the cavity. With the valve in place the cavity is then effectively between anode and screen (see 7.45(d)). A loop attached to a type N socket is arranged to rotate within the cavity to couple the output feeder.

The input circuit is a simple trough line of half-wave electrical length, tuned by a trimming capacitor at the end remote from the valve. Input matching is achieved with a fabricated trimming capacitor in series with the input feeder. The whole grid circuit is conveniently housed in an Eddystone die-cast aluminium box, the bottom of which is fixed to the skirt of the air system socket. The space between the die-cast box and the lower surface of the cavity is enclosed, and becomes an r.f. " dead space." Heater and d.c. screen connections to the valveholder are made within this space.

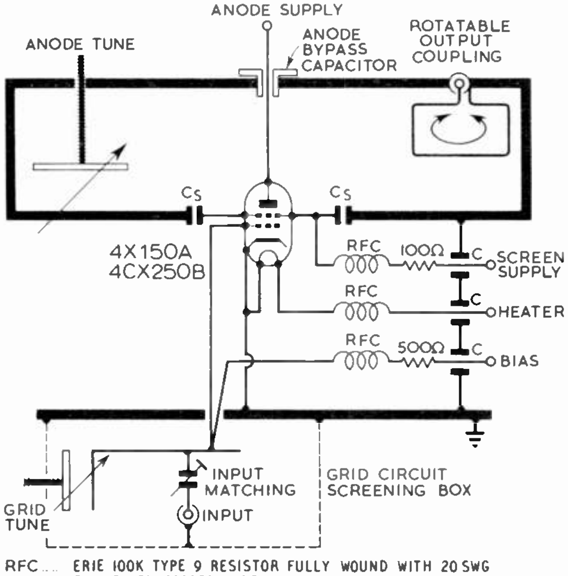


Fig. 7.42. Circuit diagram of high power 430 Mc/s cavity amplifier

Construction

Fig. 7.43 shows the basic construction of the anode cavity resonator, which measures 13$\frac{1}{2}$ in. square by 1 in. deep internally. The internal dimensions control the resonant frequency of the cavity, and this should be borne in mind if it is desired to replace the 1 in. × $\frac{1}{4}$ in. brass bars specified by aluminium U section, which may be either fabricated from sheet or extruded. The trimming capacitor mounting is placed as close as possible to the valve for maximum effect, and the output loop assembly is placed as near the outer edge of the cavity as possible. The reason is that the voltage and current distribution within the cavity is such that the voltage is a maximum at the centre of the cavity, where capacitative loading has maximum effect. The current is a maximum at the outer edge of the cavity where inductive loop coupling will be a maximum. Figs. 7.45 (a), (b) and (c) show the construction of the valveholder clamp and trimming capacitor. The valveholder clamp is shown as a complete ring which ensures intimate contact between the socket and the cavity. In the event of difficulty in fabricating this ring, an alternative would be to use the three small clamp pieces supplied with the valveholder. It may be necessary, however, to fabricate three additional clamp pieces to ensure good contact around the periphery of the valveholder.

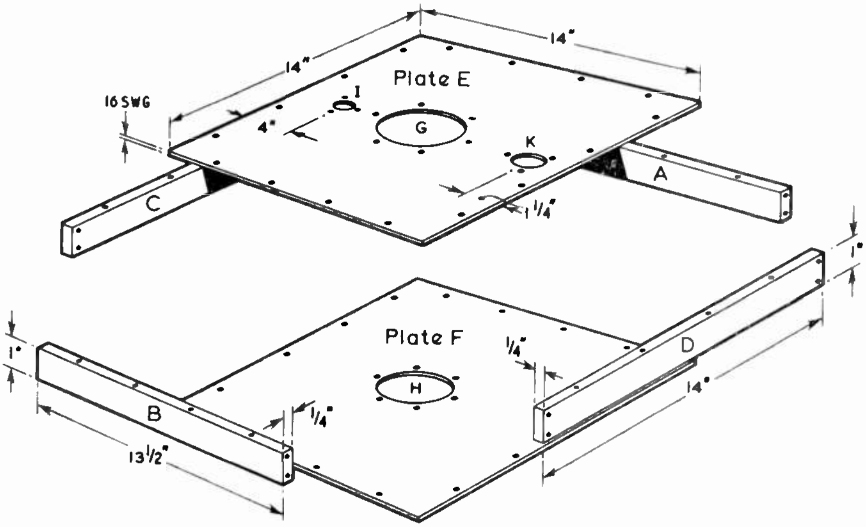


Fig. 7.43. Construction of the main cavity for the high power 430 Mc/s amplifier.

A, B, C, D	1 in. × $\frac{1}{4}$ in. brass bars making the side wall of the cavity.	I on E	$\frac{1}{2}$ in. diameter hole spaced 4 in. from edge of plate with three 6B.A. clearance holes equally spaced on $\frac{3}{4}$ in. P.C.D. to permit fixing of the trimming capacitor support.
A, B	13$\frac{1}{2}$ in. long.	K on E	Output coupling hole 1 in. diameter spaced at 1$\frac{1}{4}$ in. from edge of plate with three 4B.A. clearance holes equally spaced on 1$\frac{1}{2}$ in. P.C.D.
C, D	14 in. long.	H on F	2$\frac{1}{4}$ in. diameter hole centrally placed with six 4B.A. clearance holes equally spaced on 3 in. P.C.D. to permit the valveholder to be clamped on the inside of the plate.
E, F	Top and bottom plates of cavity made of $\frac{1}{16}$ in. aluminium, 14 in. × 14 in.		
G on E	3 in. diameter hole centrally placed with six 4B.A. clearance holes equally spaced on 4 in. P.C.D. to enable the d.c. blocking capacitor to be clamped on the outside of the plate.		

The main requirement as far as the trimming capacitor is concerned is that the disc face should rotate parallel with the surface of the cavity.

Fig. 7.45 (d) shows the output loop assembly which should be fitted to the cavity with the loop in line with the valve for maximum coupling.

The anode capacitor plate with $\frac{3}{8}$ in. diameter fixing holes to facilitate fitting of the insulating bushes is illustrated in Fig. 7.46. This assembly may be simplified by using a square form rather than circular, but should be of the same effective area, e.g. 4$\frac{1}{2}$ in. × 4$\frac{1}{2}$ in. A recess is provided for fitting spring finger contacts which make connection to the anode. A recess on the underside is provided to locate the ceramic chimney, part of the air system socket.

The insulator for the d.c. blocking capacitor and the insulator bushes are shown in Fig. 7.44. Mica was used on the prototype cavity, but p.t.f.e. sheet would be equally satisfactory. Similarly, the insulator bushes were p.t.f.e. on the prototype, but since no r.f. voltage appears across these bushes other materials may be used, e.g. Tufnol, Perspex or polystyrene.

Fig. 7.47. shows the layout of the grid circuit within the die-cast aluminium box. The grid line is fabricated from $\frac{1}{2}$ in. wide by 18 s.w.g. strip and one end is bent through a right-angle to form one plate of the trimmer capacitor. The line is screwed to the grid connector of the air system socket with a short tube to space the line $\frac{1}{4}$ in. from the bottom of the box, and a $\frac{1}{4}$ in. p.t.f.e. spacer serves to support the other end of the line. The type N input socket is fitted with a type N hood on the inside of the box which, in conjunction with a short length of 50 ohm coaxial cable, preserves the continuity of the input feeder to the matching capacitor. The d.c. grid connection consists of an r.f. choke and wire

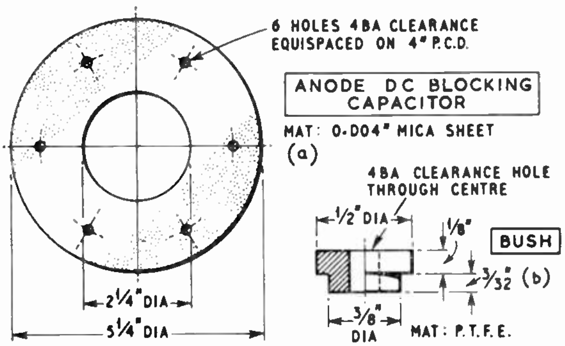


Fig. 7.44. (a) Anode d.c. blocking capacitor. (b) P.T.F.E. bush.

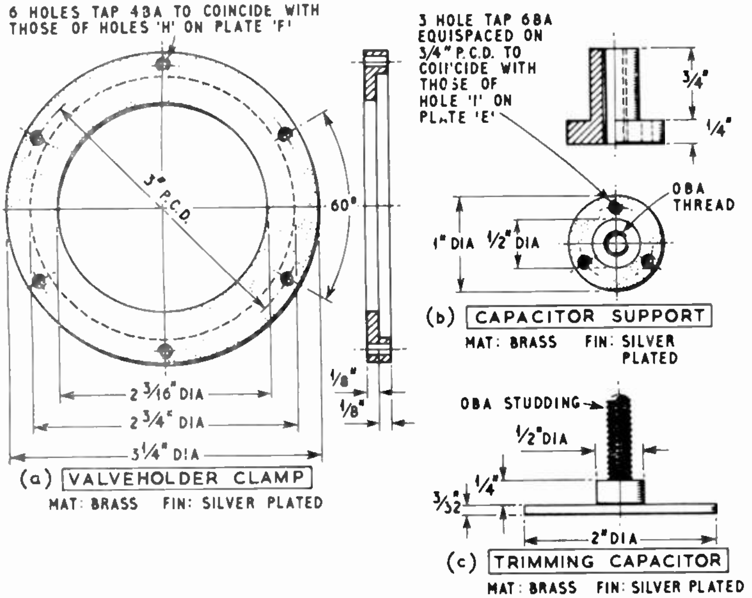


Fig. 7.45 (a) Valveholder clamp. (b) Capacitor support. (c) Trimming capacitor. (d) Output loop.

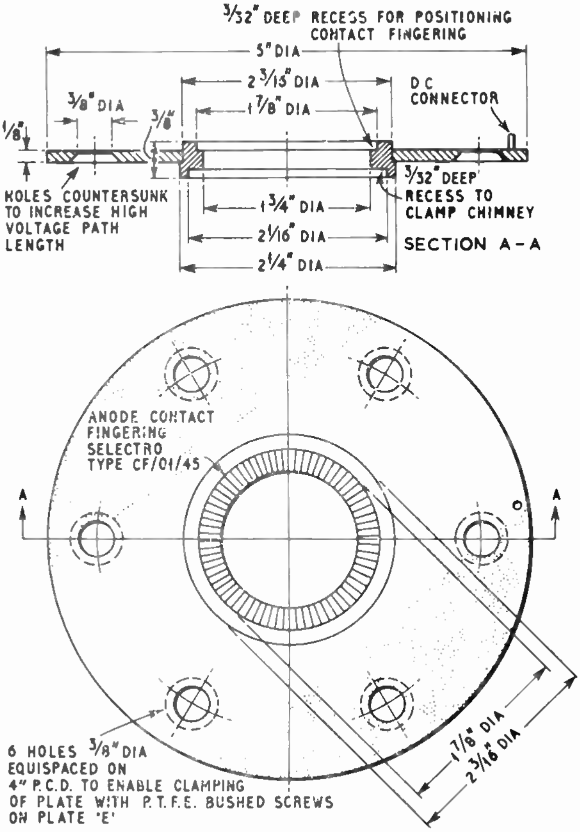


Fig. 7.46. Anode capacitor plate.

wound resistor in series between the grid of the valve and a 1000 pF feed-through type capacitor.

The complete amplifier is illustrated in **Fig. 7.48.** The r.f. dead space between the cavity and the grid circuit is enclosed by $\frac{7}{8}$ in. × $\frac{1}{4}$ in. brass strip and fitted with two 1000pF feed-through capacitors for the screen and heater leads. An r.f. choke is fitted in the heater lead, and another r.f. choke with a 100 ohms wirewound resistor in series connected in the screen lead.

Operation

It is necessary to monitor the anode, screen and grid currents to the valve, and meters should be connected in circuit. A forced air supply should be connected to give a minimum flow of 7·5 c.f.m. through the anode cooler even with only the heater supply switched on. After allowing at least 30 seconds for the heater to warm up, the grid bias, anode and screen voltages should be applied, in that order.

The grid circuit may then be tuned and matched to the drive source, using the anode current meter as an indicator. Due to secondary emission effects, grid current may not be indicated even when full drive is applied, and may even be negative. In this respect it is necessary to fit a bleed resistor across the negative bias supply drawing approximately 20 mA to stabilize the supply under negative grid current conditions. These remarks also apply to the screen supply. The bias voltage should be –80 volts and the screen voltage 250 volts. The drive requirement is 15 to 20 watts.

With an anode voltage of 1000 volts and current of 150 mA, an r.f. output of approximately 75 watts may be obtained.

70 cm. Coaxial Cavity

As an alternative to the box type cavity a tubular (coaxial) type cavity offers a considerable space saving, but of course a significant amount of machine work will be needed.

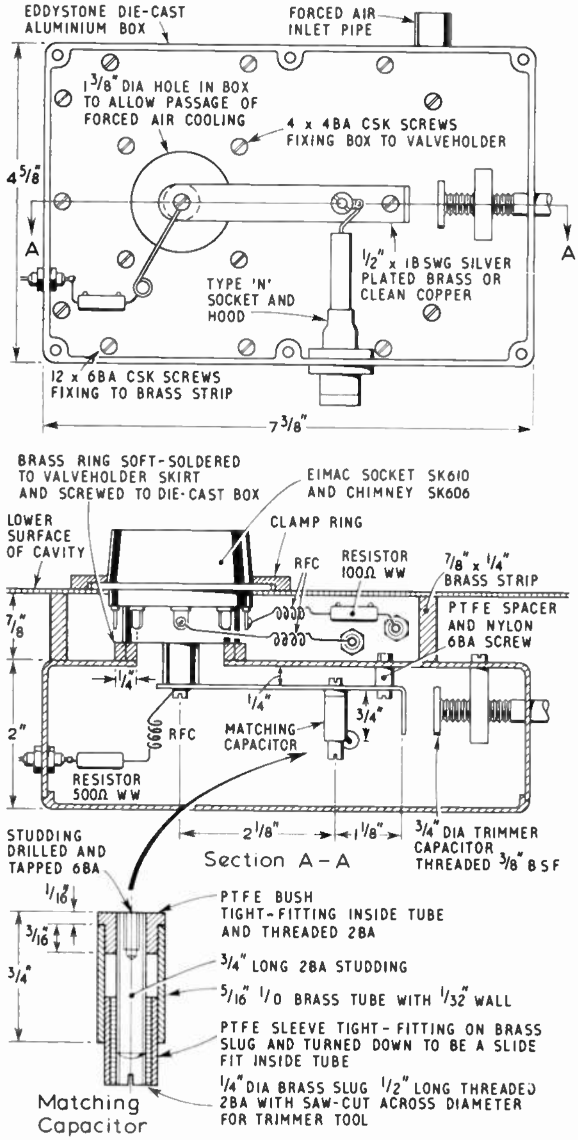


Fig. 7.47. Construction and layout of the grid circuit for the 430 Mc/s amplifier.

Fig. 7.49 shows a general arrangement of the anode circuit and gives the mechanical details of the various component parts.

The grid circuit may be similar to that previously described and built into a cast box, above this a plate is mounted and spaced to provide the necessary space for the supply leads.

A flange (L) is made to fit into the lower rim of the valve socket skirt, this is fixed to the cast box and completely isolates the input and output circuits, it also provides a suitable air duct for the cooling air which must be blown up from the underside of the valve socket and then through the anode cooler.

The valve socket is fitted to an extra top plate which is attached to the cast box by the normal clips.

The anode circuit outer (A) is attached to the top plate by four eyebolts at the bottom and then the inner tube assembly consisting of the anode line (C) is soldered or brazed to the inner top disc (D) which is bolted to the outer top disc (B) from which it is insulated by the p.t.f.e. washer (E).

The fixing bolts of the two top discs are insulated from the outer top disc by p.t.f.e. bushes.

The whole assembly is then attached to the outer tube by four lugs into which eyebolts at the top end of the tube locate.

Tuning of the cavity is provided by disc type capacitor consisting of (H) and (G). The adjustable element (G) should be provided with some tensioning device such as a spring locating into the screw thread. Since there is the full h.t. between these two plates one of them should be covered with suitable insulating material to prevent possible short circuit.

Output from the cavity is by the conventional series tuning loop. The loop spacing from the anode line should be

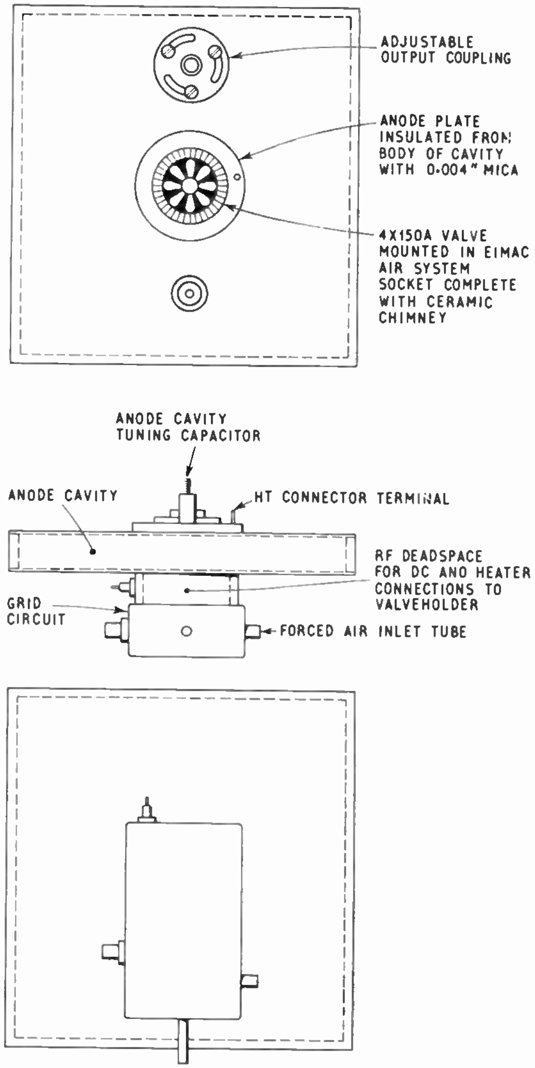


Fig. 7.48. Assembly of the complete 430 Mc/s amplifier.

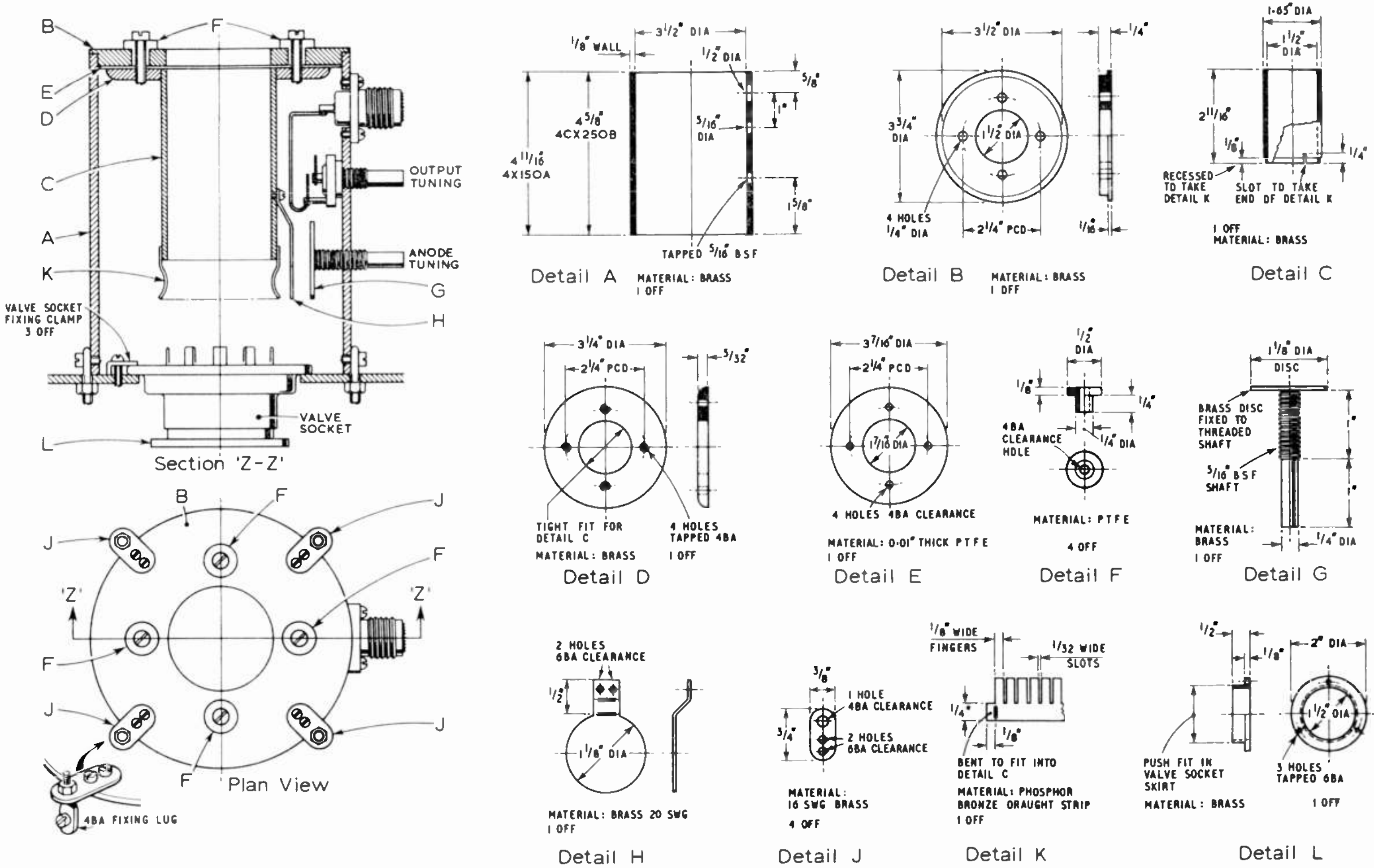


Fig. 7.49. Constructional details for the 70 cm. coaxial cavity

adjusted to give the maximum output, this does not need close coupling to the centre tube and it should be kept as far away as possible consistent with good efficiency.

TRIPLERS FOR 1296 MC/S

The choice of valves for the generation of any appreciable power at 1296 Mc/s is restricted almost entirely to the 2C39 in one or other of its many variants, although for lower power the DET24 is equally suitable. Semi-conductors such as Varactors and planar silicon power transistors that will give a few watts output at this frequency with good efficiency are available, but their high cost at the present time makes their use rare.

For the valve tripler, the circuit must be either of the cavity or coaxial type and both are described in this chapter.

CAVITY TYPE

The cavity is designed for use with either a TD1-100 or 2C39 disc seal triode, and can be driven by any 70cm transmitter capable of giving a few watts of output. From the circuit diagram of **Fig. 7.50** it will be seen that the valve operates in the grounded grid mode, the drive being applied directly to the cathode. The 23cm output is taken from a loop which couples into the cavity.

Construction

The cavity is constructed from 20 s.w.g. sheet brass, the general details of which, together with part of the assembly order, are shown in **Fig. 7.53**. Full details of the main body of the cavity are shown in **Fig. 7.52**. This is constructed by forming the four sides from a flat sheet to produce an open box. It should be noted that the corners are not soldered. This permits easy movement of the grid tray which is a sliding fit inside the open box. It also allows a good contact to be made between the grid tray and the box when the side fixing screws are tightened.

The position of the grid tray in the box is the coarse adjustment for the anode circuit. Fine tuning is achieved by means of a tuning paddle which is also illustrated in Fig. 7.52. This consists of a ⅝ in. by ¾ in. metal plate which is soldered into a slot cut into the end of a ¼ in. operating spindle.

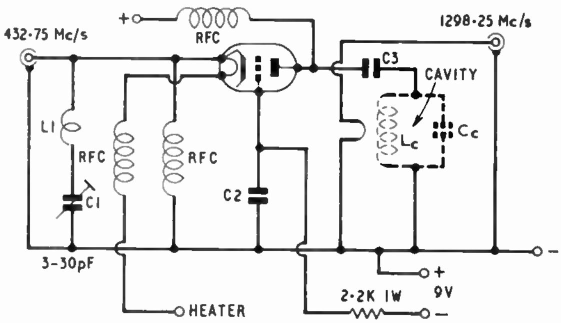
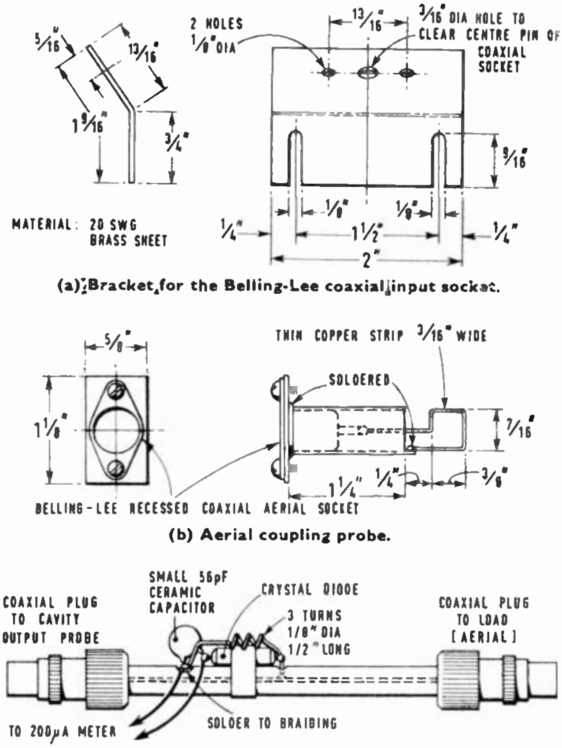


Fig. 7.50. Circuit of the tripler-cavity. C2 and C3 are formed by the anode and grid plates in the cavity; Cc and Lc represent the cavity; L1 is formed from thin copper strip, 3/16 in. wide and 2 11/16 in. long bent to a " U " shape; RFC, 2¼ in. of 20 s.w.g. enam. copper, wound on 3/16 in. diam. former.



(a) Bracket for the Belling-Lee coaxial input socket.

(b) Aerial coupling probe.

(c) Diode probe on a coaxial line.

Fig. 7.51. Details of sub-assemblies.

Output from the cavity is taken via the ½ in. inside diameter tube mounted directly opposite the fine tuning control. Into this tube slides the aerial coupling probe illustrated in **Fig. 7.51.(b)**. It is important to ensure that the probe unit is a reasonably tight fit into the tube on the side of the cavity, and to this end, if possible, the two parts should be constructed from telescopic tubing. Since the diameter of the outer tube is not critical to within 1/32 in. it is permissible to cut a lengthwise slot in it to allow it to be closed up slightly and thus ensure a tight fit round the probe.

The anode and grid plates are shown in **Fig. 7.54 (a)** and **(b)** respectively, and are quite straightforward. Careful attention must be paid to the contact fingers for the valve, and it is important to ensure that the final size is suitable for the particular sample of the valve being employed. The fingering is formed by making ⅛ in. spaced cuts with a fine saw part way across a short length of brass strip (phosphor bronze draught excluder is also excellent for the purpose). This strip is then formed into a circle, fitted to the inside of the hole provided and soldered into position. Small adjustments to the final size may be achieved by slightly bending the contact fingers.

Polythene sheet with a thickness of 0·008 in., suitably cut to shape, is used as insulation between the anode plate and the cavity body. This has been found satisfactory for anode voltages up to 600 volts. Between the grid plate and the grid tray, a similar sheet of polythene is used.

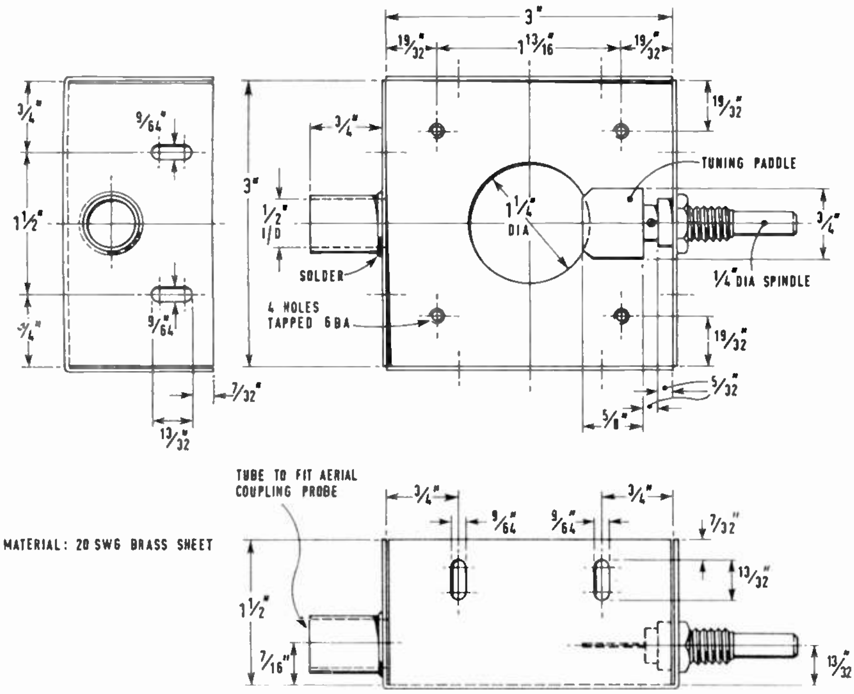


Fig. 7.52. Box cavity body.

on the anode plate, are filed off flush with the underside of the plate, otherwise the polythene insulation will be damaged, and short circuits occur.

Adjustments and Operating Conditions

Connect the 9 volt bias supply to the valve with a 0-50 mA meter in series with the negative lead. Apply a 6 volt source, capable of supplying 1 amp. to the heater of the valve. Do not attempt to apply h.t. at this stage, but connect the anode of the valve from the solder tag on the anode plate via a 100 mA meter to the cavity body.

If 70cm drive is applied to the cathode, grid current will flow, and some anode current will show on the 100 mA meter. The trimmer C1 and the inductance L1 should be adjusted to give maximum grid current. This should be in the region of 25 mA. The earthing point on the trimmer may need changing to a different position on the grid tray if insufficient grid current is obtained. Unfortunately there is a multitude of combinations of the capacity of C1 and the inductance of L1 which will give resonance at 70cm, but the correct ratio of these two is quite critical in order to achieve maximum drive to the tripler. Once the grid current is of the correct order, then the meter in the anode circuit should read between 30 and 60 mA.

The coaxial input socket is mounted on the bracket shown in Fig. 7.51 (a), and this is held on to one side wall of the cavity by means of two of the screws which hold the grid tray in position. The lead from the coaxial socket to the cathode of the valve is made from thin copper strip ⅜ in. wide. If copper is not available, brass may be used, so long as it is thin. It is important that the lead is not made too rigid as otherwise there may be a danger of damaging the valve.

The input inductance (see L1 in Fig. 7.53) consists of a length of brass or copper strip, 2 11/16 in. long by 3/16 in. wide formed into a U with arms of equal length. With the base of the cavity uppermost, this U-shaped inductance is inverted and soldered at one end to the coaxial socket, at the other end to the 3-30 pF Philips trimmer. The earthy side of this trimmer is soldered to a tag located under one of the two 6B.A. screws in the corner of the grid tray farthest away from the coaxial input socket.

The 2·2 K ohm grid resistor is soldered to a tag secured by a short 6B.A. screw to the centre of one side of the grid plate. It is particularly important that this screw, and its counterpart

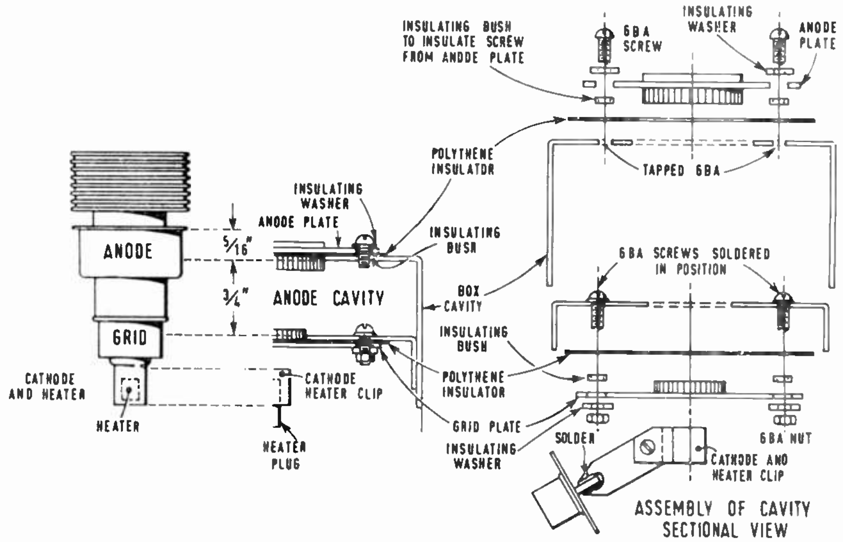
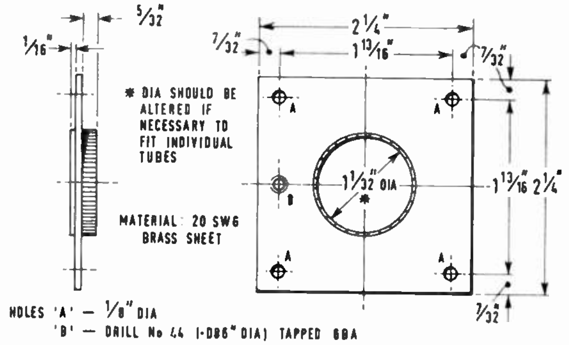
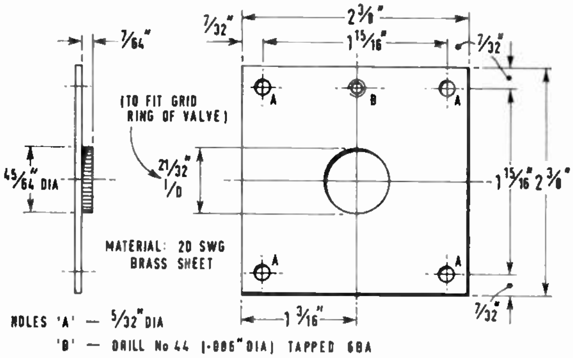


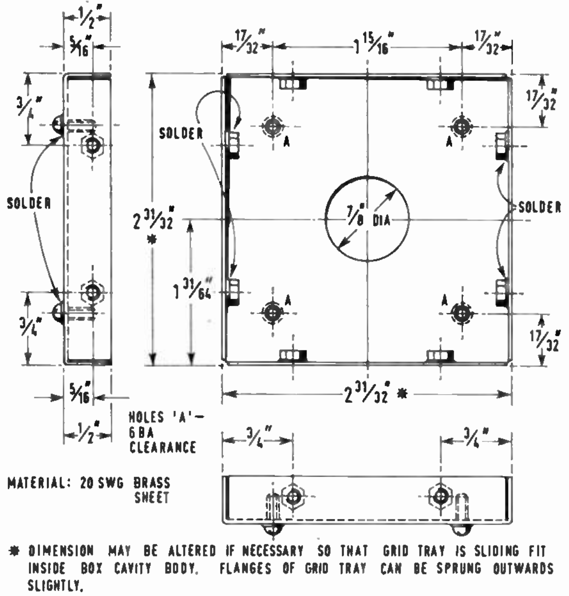
Fig. 7.53. Assembly of cavity. The position of the valve is very important: it should be placed for maximum output, but this will affect the tuning. Should the tuning paddle not correct the tuning, the grid tray will need to be moved slightly.



(a) Anode plate.



(b) Grid plate.



(c) Grid tray.

Fig. 7.54. Constructional details of anode and grid components.

When a suitable level of grid current has been achieved, attention can be turned to resonating the anode cavity. For this it is necessary to have some means of indicating 23cm output. One possibility is to use a 6 volt 0·04 amp. flash lamp bulb soldered directly to a coaxial plug. Alternatively, and in many respects better, a diode probe on a coaxial line, as illustrated in Fig. 7.51 (c) can be used.

First, set the tuning paddle to an angle of 45° to the plane of the anode plate. Loosen slightly the four grid tray fixing screws and then very carefully move the position of the grid tray until resonance is found. Resonance will be denoted by a small dip in anode current. If the diode probe on the coaxial line is in circuit a reading will be obtained on the diode meter. Lock the grid tray into position, and check that the tuning paddle will resonate the cavity more critically. The anode may now be disconnected from the chassis and taken to a supply of about 350 volts positive via an r.f. choke. The anode current may lie anywhere between 50-100 mA depending on the valve.

Acceptable levels of modulation may be achieved by modulating only the 70cm drive to the 23cm tripler. For c.w. operation, the 70cm stage can be keyed, and under key-up conditions, the anode current in the tripler should fall to zero.

Simple 23 cm Tripler

This tripler is based on the 2C39A triode and the use of a slab line anode circuit shown in Fig. 7.55. The complete unit is enclosed in a die cast box which acts as a tuned cavity.

The normal lid of the box should be replaced by a piece of copper or brass to which the screen and other components are soldered.

Construction

The anode circuit is enclosed in a standard die cast box

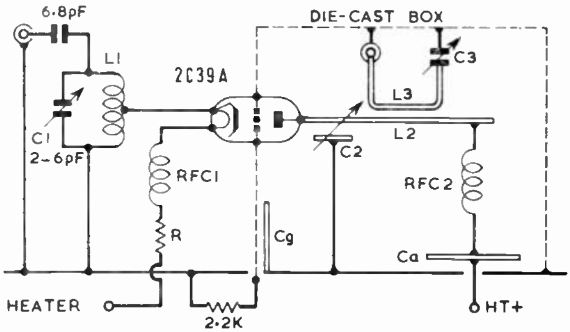


Fig. 7.55. Circuit diagram of simple 23 cm tripler

TABLE 7.13—List of Components

L1	2 turns ⅛ in. wide strip ¼ in. diam. length ⅜ in. cathode tap ⅜ turn from earth end.
L2	Anode slabline } See Fig. 7.56
	Output coupling loop } See Fig. 7.56
Cg	Grid plate to chassis by-pass capacitor } insulation 0·004/6 in.
Ca	H.T. to chassis by-pass capacitor } insulation 0·004/6 in.
RFC1	8 turns ¼ in. diam. 18 s.w.g.
RFC2	7 turns ¼ in. diam. 22 s.w.g.
R	Value chosen to drop heater voltage to 5·5 volts at the valve
C1	2-6pF concentric trimmer C004/EA Mullard
C2	Adjustable plate, see Fig. 7.56
C3	2-6pF concentric trimmer C004/EA Mullard

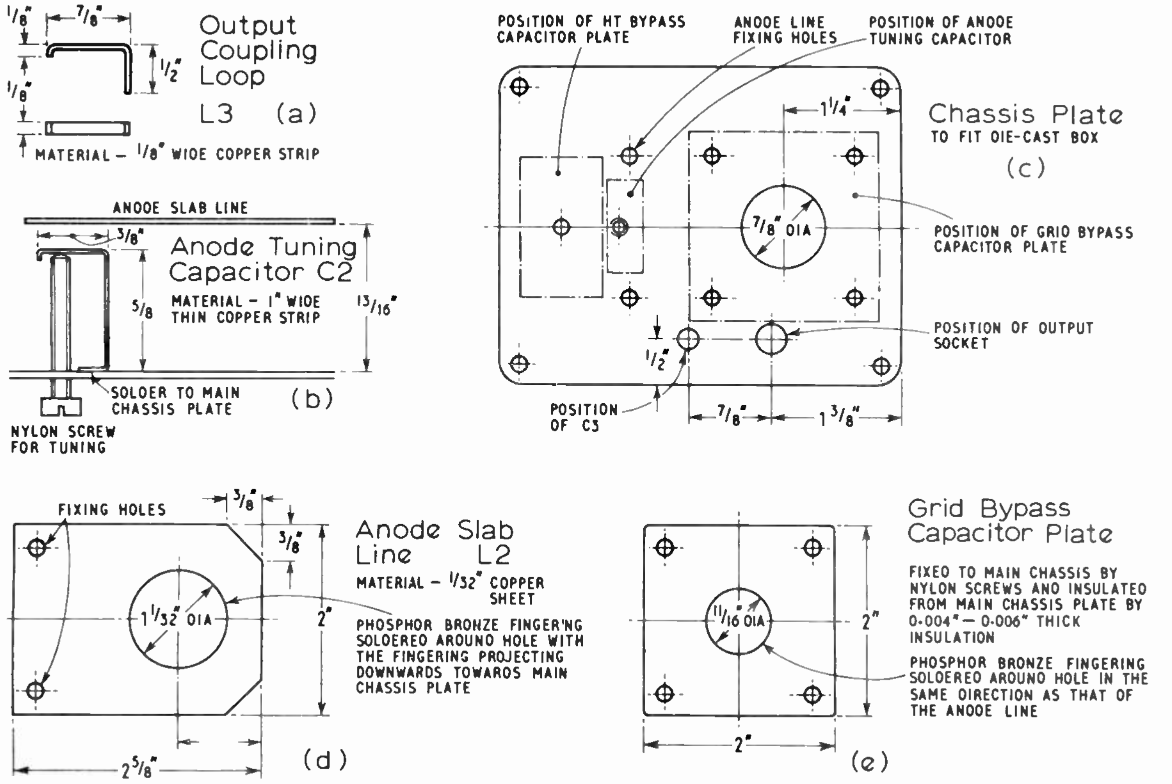


Fig. 7.56. Constructional details for the simple 23 cm. tripler.

measuring $4\frac{5}{8} \times 3\frac{3}{8} \times 2\frac{1}{4}$ in. The lid is replaced by a copper or brass plate $\frac{1}{16}$ in. thick and a V of copper or brass sheet soldered to the top, as shown in the illustrations.

With a new plate it is desirable to fit angle or square bars along the long sides of the box, one of these can be seen in the photograph.

The anode circuit is mounted $\frac{13}{16}$ in. from the chassis plate by means of insulated spacers, the h.t. feed to the valve is taken through an r.f. choke and by-pass capacitor, consisting of a plate $1\frac{1}{2}$ in. $\times$ $\frac{7}{8}$ in. insulated from the chassis plate by mica or other suitable material, between 0·004 and 0·006 inches thick.

The grid which is at r.f. ground potential is connected to a 2 $\times$ 2 in. plate **Fig. 7.56(e)** bolted to, and insulated from, the chassis plate using nylon screws and mica insulation, which should be 0·004 to 0·006 in. thick.

Tuning up

A drive power of 10 watts at 432 Mc/s is needed for full power operation.

The input circuit is tuned for maximum grid current which should be approximately 30-40mA.

Care should be taken to ensure that the heater voltage (measured at the valve terminals) does not exceed 5·5 volts.

The input circuit consists of a parallel tuned circuit L1, C1; the cathode connection to the valve is taken from a tap on L1 at three-quarters of a turn from the earthy end.

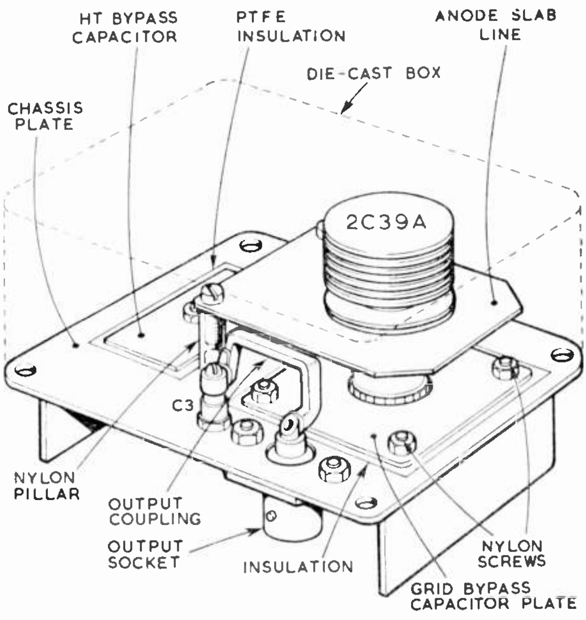


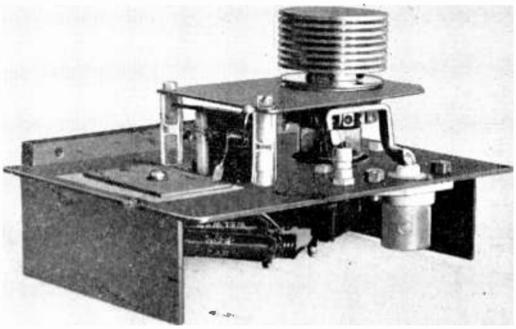
Fig. 7.57. Diagram showing position of major components

Contact to the anode and grid is made by phosphor bronze fingering $\frac{3}{32}$ in. long soldered to the hole in the anode slabline and the grid plate.

With the input circuit properly adjusted, apply h.t. and tune the anode circuit for maximum output. This is best done using a suitable output indicator.

Applied h.t. in the range 250-450 volts at up to 50 watts input may be used, with forced air cooling. At this power level cooling air may be passed through sides of the die cast box. Any holes for this purpose should be covered with fine wire gauze to avoid disturbing the r.f. field within the box (cavity).

Sufficient output can be obtained from this tripler for local contacts by simply earthing the h.t. lead to chassis.



General view of a completed 23cm. tripler.

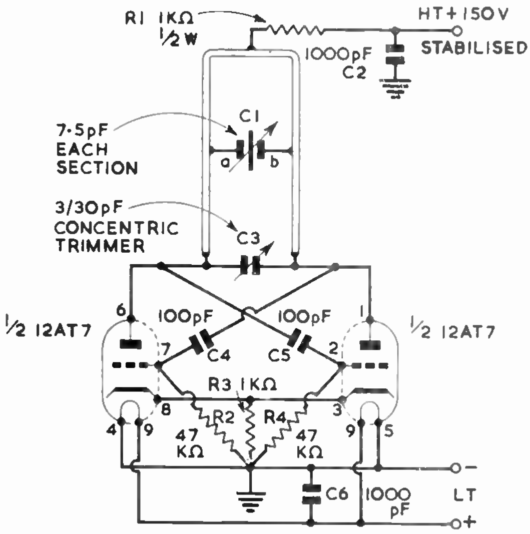


Fig. 7.58. The circuit diagram of the Kalitron oscillator.

setting being carried out by means of C3 a 3–30 pF concentric trimmer soldered across the line close to the connections to the valve anodes.

The output from the oscillator is taken by means of a loop of 10 s.w.g. wire 2 in. long and 1 in. wide coupled inductively to the low potential end of the line and situated ¼ in. from it. One end of the loop is soldered to the centre connection of the coaxial socket and the other to a tag under one of the screws securing the socket to the end plate of the box.

A V.F.O. FOR 144 MC/S

The v.f.o. to be described is designed to take the place of a crystal oscillator in a transmitter. The fundamental frequency may be varied between 72 and 73 Mc/s, thus enabling transmission to take place anywhere in the 144-146 Mc/s band. The high fundamental frequency of the oscillator ensures that no harmonics fall within the television or f.m. bands.

The circuit employed is the Kalitron (see Chapter 5—*V.H.F./U.H.F. Receivers*), the valve being a 12AT7 twin triode with capacity feedback from each anode to the grid of the opposite section as shown in **Fig. 7.58.** Details of the layout are given in **Fig. 7.59.** The sections are held together with PK self-tapping screws.

The tuned circuit consists of a hairpin shaped loop of 10 s.w.g. wire, preferably silver plated, measuring 5½ in. long with a spacing of 1 in. between the two straight sides. The last inch is bent down at right angles to the rest of the line and the ends are connected directly to the two anode tags of the valveholder which should have p.t.f.e. insulation. The opposite end of the line is supported by R1 from the feedthrough bypass capacitor C2 which forms the connection for the 150 volt stabilized anode supply. The split-stator tuning capacitor is connected midway along the line, band-

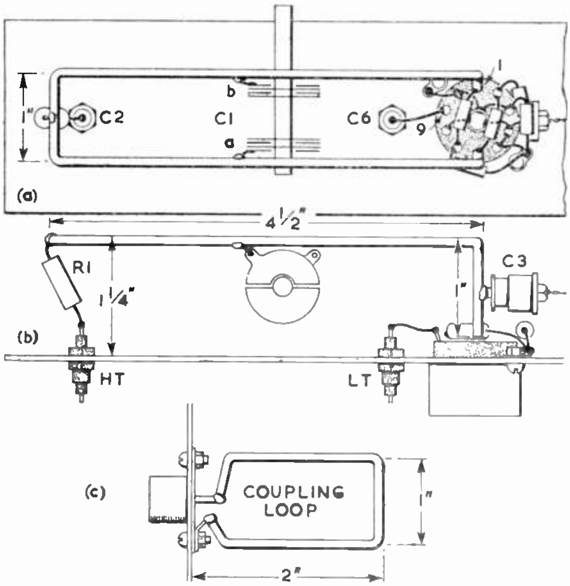


Fig. 7.59. Layout of components in the Kalitron oscillator.

The circuit diagram of a suitable buffer amplifier for use with this v.f.o. giving sufficient output at 72 Mc/s to drive a 5763 doubler stage to 2 mA grid current is shown in **Fig. 7.60.** The original amplifier employed Mullard type EF50 valves but equivalent minaiture valves such as the EF91 or 6AM6 may be substituted if desired. A frequency range of ± 100 kc/s on the 2m band may be obtained without having to retune the buffer amplifier.

The stability of the oscillator is such that after drifting approximately 25 kc/s lower in frequency during the first seven minutes after switching on, the frequency remains constant within half a kilocycle over a period of several hours.

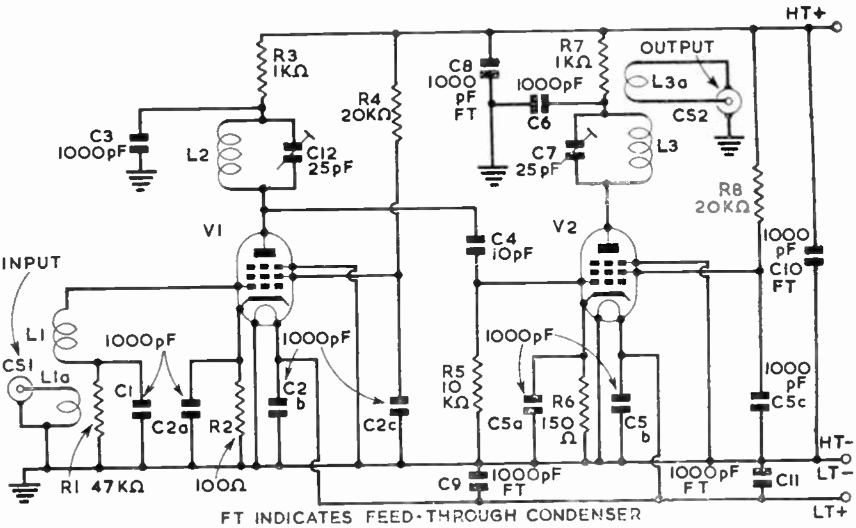


Fig. 7.60. Circuit diagram of a suitable buffer amplifier. The oscillator is connected to this unit by a short length of coaxial cable.

V.H.F. DUMMY LOAD

A simple dummy load is useful in carrying out non-radiating tests with transmitters, and also for the setting up of reflectometers, where a termination having a negligible reactive component is essential.

The physical form of a 25 watt dummy load is shown in the photograph and is that of a hollow tube with an inner rod down the centre; thus it resembles a short length of coaxial line, and its characteristic impedance is determined by the radii of the inner and outer conductors in a similar manner. One end of the line is closed with a disc of copper, the other is fitted with a suitable coaxial connector. The power applied to the load is dissipated as heat in the outer conductor of the line, and is radiated to the surrounding air more readily than if the inner conductor were the dissipative element. This is important because the resistive component of the load is dependent upon its temperature, decreasing as the temperature is raised. Where it is essential to have a known accurate termination it is desirable therefore that the input to the load should not be sufficient to raise its temperature appreciably.

A 25 watt dummy load for 145 Mc/s.

The dimensions of the coaxial line are determined by those of the Morganite type 702 tubular carbon resistor which forms the outer conductor. The resistor used had an overall length of 150 mm. with inner and outer diameters of 15 mm. and 25 mm. respectively. The outer cylindrical surface is plated for a distance of 1 in. from each end to assist in making electrical contacts to the resistor. The diameter of the inner conductor is calculated from the formula $Z_0 = 138 \log_{10} D/d$ (where D = inner diameter of outer conductor, and d = overall diameter of inner conductor) and in this instance was found to be 4·67 mm. (approximately $\frac{3}{16}$ in.) for a 70 ohm load. The length of the inner conductor is made sufficient for it to project through a suitably sized hole in the centre of the short circuiting disc when assembled, so that it can be soldered up as the final assembly operation.

Connection to the outer conductor is made by means of a 1 in. diameter capacitor clip which in turn is screwed to a 2 in. square by $\frac{1}{8}$ in. thick brass plate referred to as the mounting plate. On the other side of the mounting plate is the Amphenol socket used to make the external circuit connections. The inner of the Amphenol connector is soldered to one end of the inner conductor of the line, which is first drilled and slotted for a distance of $\frac{1}{4}$ in. to facilitate this operation. It was found necessary to interpose a spacing plate between the mounting flange of the Amphenol socket and the mounting plate to ensure that the end of the tubular resistor seated squarely without fouling the back of the socket. This may not be necessary with Amphenol sockets of American manufacture.

The short circuiting disc is soldered to the free end of the resistor using the mounting ears which are bent over so that they lie over the cylindrical plated area of the resistor. A very hot iron is required for this operation to ensure that the solder flows freely. It is not desirable to go over it more than once because the plating on the carbon will tend to lift off.

Exact details of the various components are not given as the individual constructor may wish to vary the design somewhat to suit his own requirements.

R.F. Output Indicators

For various reasons, such as circuit out of balance both inside and out of the amplifier, there is often a significantly different tuning for the minimum dip of the anode current and the maximum r.f. output.

It is therefore desirable to provide a reliable means of indicating direct r.f. output; this can most simply be done by using a simple semiconductor diode voltmeter such as that shown in **Fig. 7.61**, connected across the output socket. There is no need for the meter to be calibrated, all that is necessary is to to tune for maximum indication.

The actual values of R_1 and R_2 are not important but their total resistance should be high compared with the output line impedance, a value of approximately 10K (total) is suitable with the ratio of R_1 to R_2 between 5 to 1 and 10 to 1, the precise values will largely depend on the full scale current of the meter.

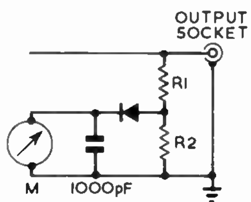


Fig. 7.61. Simple semi-conductor r.f. voltmeter.

Alternatively if a reflectometer is available, it should be kept permanently in circuit and when set to read *forward power* it will also provide the necessary r.f. indication for the correct transmitter tuning.

Variable Frequency Crystal Oscillator (VXO)

With the increased activity on the 70, 144 and 432 Mc/s bands, particularly under good propagation conditions, the need for frequency change becomes almost essential. It would be possible to follow the h.f. band practice of using a single frequency for contact between stations and operating a v.f.o. to cover the frequency range of the band, but considerable difficulty could be experienced in the construction of a suitably stable oscillator. A reliable alternative which is becoming more common, is to use the v.x.o. method of providing a stable but adjustable frequency source.

In this method, the crystal frequency is *pulled* from its normal frequency by introducing a series inductance, the value of which will depend on the crystal and the range of frequencies over which it is required to tune. With 6 or 8 Mc/s crystals a frequency range of about 100 and 200 kc/s at operating frequencies of 70 and 144 Mc/s respectively can be obtained. A typical circuit is shown in **Fig. 7.62.**

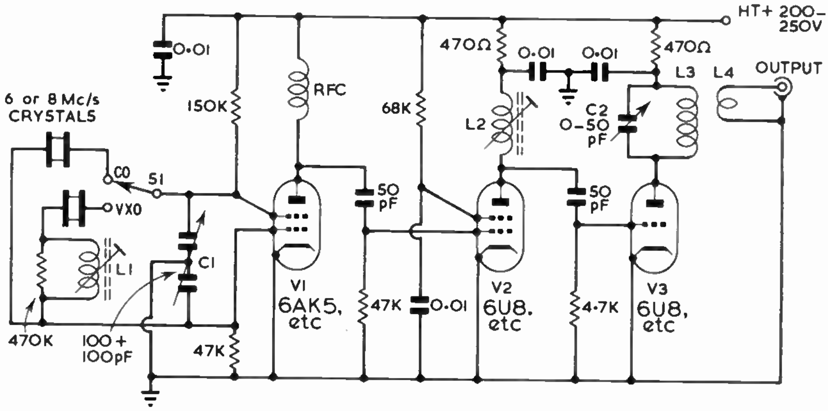


Fig. 7.62. Circuit diagram of a variable-frequency crystal oscillator.

TABLE 7.14—List of Components

C1	100pF + 100pF split-stator
C2	0–50pF
L1	16–24μH
L2	24–35μH
V1	6AK5 or similar
V2V3	6U8 or similar
L3	21T 24 s.w.g. 0·5 in. diameter 32 t.p.i.
L4	3T 24 s.w.g. 0·5 in. diameter 32 t.p.i. spaced 1T from L3

In the circuit, L1 is adjusted to give the required frequency change without introducing instability and L2 should be tuned to the centre of the frequency range at the fundamental frequency. V3 operates as a multiplier and the tuned circuit C2, L3 is set to the third harmonic of the crystal frequency, such as 18 or 24 Mc/s depending on the crystal in use.

COMPONENT TOLERANCES

When testing any of the transmitters made in accordance with the instructions given in this chapter, the current readings obtained may differ from those quoted in the various tables, which should be regarded as average values. If the results obtained differ widely from those given, some experimentation in the value of components, particularly resistors, is advised. This does not indicate a fault in the design of the equipment but may be necessary in view of the manufacturing tolerances of the components used. It must be borne in mind that the transmitters are " one off " prototypes, and component values were selected for them which gave satisfactory results in the equipment described; it is thus possible that, in some instances, all components having favourable tolerances have been used. In general resistors of ± 20 per cent tolerance were incorporated.

CHAPTER 8

KEYING AND BREAK-IN

To impress intelligence on a carrier wave, it must undergo a process of modulation, and the simplest way of accomplishing this is on/off keying of a transmitter. This necessitates encoding the information in the form of specific trains of pulses of different lengths, to represent each character, the usual system being the Morse code. The only method of keying the transmitter to be discussed in this chapter consists of changing the amplitude of the carrier; an alternative method, known as frequency shift keying, is to vary the frequency, but this offers very little advantage for normal Morse communication purposes, and is usually applied solely to operating teleprinters. This subject is considered in Chapter 11.

The bandwidth required by a properly keyed signal is small, and should be directly related to the speed of sending involved. The optimum condition is not always easily attained, however, and the result of a maladjusted keying system is usually the radiation of clicks and other useless emissions over the whole of one or more frequency bands. It is also possible to obtain the opposite effect, however, through an attempt to avoid harmful keying transients; the result of this condition is to produce a very " soft " characteristic which will prevent the reading of Morse signals at high speeds.

The Baud and Keying Weight

The rate of transmission of telegraph signals is measured in *bauds*, one baud being equivalent to the transmission of intelligence at the rate of 1 Morse Code dot per second. More commonly, code speed is measured in words per minute or w.p.m. The relation between w.p.m. and transmission speed in bauds is given by $n \text{ (w.p.m.)} = 52n/60$ baud. In establishing the equality, the assumption is made that (a) the average duration of signals per Morse character is equivalent to 9 dots, (b) the average English word contains 5 characters or letters and (c) each space between words is equivalent to 7 dots. It also follows that the duration of a dot, t, is given by $t = 1154/n$ milliseconds.

The standard Morse code (International Telegraph Code No. 1) dictates that a space between characters of a letter shall also be one baud, that the length of a dash and the interval between letters shall be three bauds, and a space between words seven bauds. **Fig. 8.1**, showing the length of a dot or baud for a range of transmission speeds, should be used for reference when a keying waveform is being tailored, because it is important to provide the correct amount of softening in relation to the upper speed which is to be used: the rise or decay time should not exceed about one sixth of a unit length at the highest speed, so that at 24 words per minute, or 20 bauds per second, such times should not exceed 8 milliseconds.

If a signal has its dots or dashes too long with respect to the spaces between them, it is said to be *weighted*. This term is used particularly in connection with keying relays, as a slow component would clip the dots of a transmission unless some weight is added to the switching pulses to compensate for it.

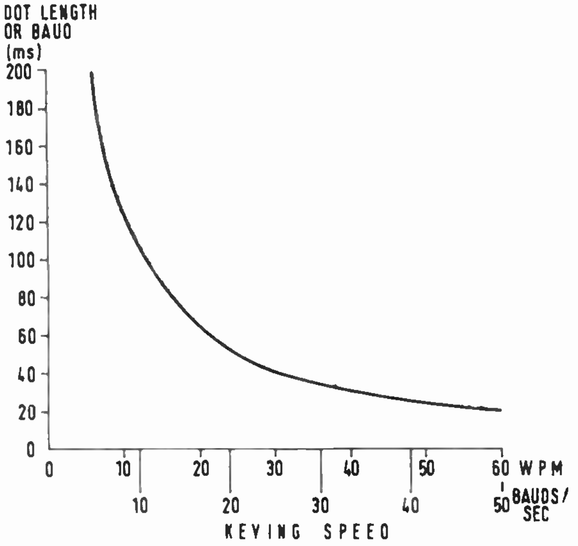


Fig. 8.1. Graph showing dot lengths for a range of transmission speeds. Once the design figure of the maximum speed of sending for a transmitter has been decided, this graph should be referred to for help in choosing keying relays and key-click filter circuits.

KEYING CHARACTERISTICS

There are two main components which affect keying characteristics: envelope shape and frequency stability. Any keying trouble such as key clicks, ripple, chirp and spacer-waves which are being discussed here, can be attributed to a fault condition in either or both of these components.

Envelope Shape

The envelope of a keyed signal is the outline of the pattern that the signal would display on an oscilloscope. It can be observed by feeding the keyed r.f. signal on to the Y plates or vertical amplifier of a slow-scan oscilloscope, and setting the timebase in synchronism with the keying speed. The transmitter is best keyed by an automatic device producing a regular sequence of dots or dashes so as to obtain a steady display.

In general, if no precautions are taken, the pattern will be square and sharp, or " hard," and will radiate clicks over a wide range of frequencies. The rise and decay times of the carrier must therefore be lengthened until the clicks are no longer objectionable, but without impairing the intelligibility at high speed. This requires rise and decay times of 5 to 20 milliseconds, and the methods of achieving this will be described in the sections on different keying systems.

Key filter circuits rely for their operation on the time constant of a CR (or LCR) circuit. The time constant ($t = CR$) of such a circuit is the time taken for the voltage across the capacitor to fall to a fraction of its full value given by $\frac{1}{e}$ (about $\frac{1}{2 \cdot 7}$), not to zero. Theoretically the voltage would never reach zero, but for all practical purposes, it may be regarded as such after a period of $3CR$ milliseconds (C in microfarads, R in kilohms). Thus if a *decay-time* of 6 mS is required, the circuit time constant should be about 2 mS. The situation is not always so simple; for example, if the p.a. has a grid-block keying system providing a blocking bias of −300 volts where 150 volts would be sufficient to reach cut-off, the p.a. will have ceased to conduct well before a time $3CR$. This is a good reason for designing key filters experimentally.

The characteristics of the p.a. power supply may contribute to the envelope shape, as the voltage from a power unit with poor regulation will drop quickly each time the key is closed, and rise when it is released. This can lead to the shape in **Fig. 8.2(d)**. This effect is not necessarily undesirable, but it can be lessened by use of a choke-input filter in the p.a. supply.

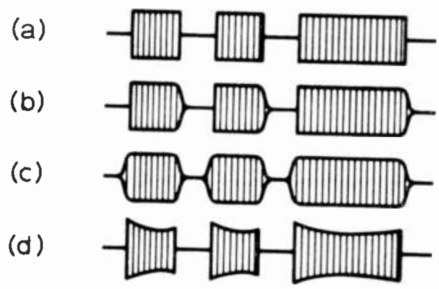


Fig. 8.2. Keying envelope characteristics. (a) Click at make and break; (b) Click at make, with click at break suppressed; (c) Ideal envelope with no key clicks; (d) Effect on keying envelope of poor power supply regulation.

Ripple can be caused by r.f. feedback, but poor power unit filters are generally responsible. If ripple is present on the p.a. supply, the carrier will be amplitude modulated at the ripple frequency, whereas on the oscillator h.t. line it would probably cause both frequency and amplitude modulation. Clipping a large electrolytic capacitor on to various points in the transmitter will invariably indicate the necessary treatment.

R.F. Clicks

Although clicks caused by a hard keying envelope are radiated with the signal, local interference may be caused by r.f. clicks from sparks at the key contacts, particularly if an appreciable current is keyed in an inductive circuit. This interference is usually removed by connecting a capacitor (typically 0·005μF) directly across the key or keying relay contacts (see **Fig. 8.3**), and in severe cases by also inserting an r.f. choke (1 or 2 mH) in the live keying lead. The effectiveness of such treatment is judged by listening to the station receiver tuned to a frequency well removed from the transmitter frequency.

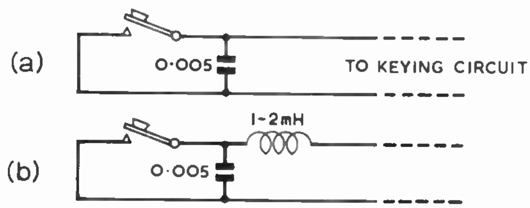


Fig. 8.3. Circuits for suppression of interference caused by arcing at the key contacts. The r.f. choke will be necessary only in severe cases. The 0·005 µF capacitor must be mounted directly across the key contacts. If an r.f. choke is used, it must be able to withstand the current flowing in the keyed circuit: in the case of cathode keying of a high power stage, this may be considerable, but then a high current choke is desirable so that a minimum of resistance is added to the cathode circuit.

Some semi-automatic "bug" keys and keying relays tend to give more trouble with clicks than hand keys because of contact bounce or mis-shapen contacts or both: with bug keys contact bounce can often be prevented by adjusting the position of the moving dot contact so that only about one-quarter of the contact surfaces are in use, and by mounting a small block of foam rubber inside the dot contact U-spring (see **Fig. 8.4**).

Worn, pitted or dirty contacts are obviously undesirable, but great care should be exercised in treating them. Only a burnishing file should be used—anything rougher would leave a burred surface which would only cause further trouble.

In some transmitter designs, the power amplifier is clamp-tube controlled and an earlier stage keyed. If the p.a. is operated in class C, it is capable of sharpening a softened signal from the buffer stage, thereby reintroducing clicks on the radiated signal. This is one reason for using sequential keying of more than one stage. The p.a. may also be triggered into low-frequency parasitic oscillations, again causing clicks. This is usually due to poor choice of r.f. chokes in the p.a. grid and anode circuits.

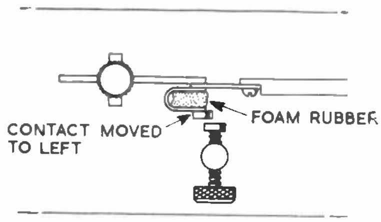


Fig. 8.4. Method of reducing contact bounce from "bug" key dot contacts. Note the displacement of the two contacts.

Chirp

Chirp is a form of frequency instability occurring each time the transmitter is keyed, and is recognised by a change in beat frequency at the beginning and end of each character when the signal is monitored on a receiver. A signal with chirp is undesirable, as it is less pleasant to copy, and is not suitable for reception by narrow-bandwidth receivers. It is usually more prevalent in transmitters controlled by a v.f.o., and there are three principal causes, which are:

(a) D.C. Instability

D.c. instability occurs when a common power supply is used for the oscillator and the p.a. (or any circuit through which the current changes in sympathy with keying). No oscillator exhibits absolute stability under varying h.t. conditions, and therefore a voltage regulator should be incorporated in the oscillator h.t. line to reduce the chirp or perhaps remove it completely. A separate oscillator power supply might be needed to cure a difficult case, but improving the regulation of the common power unit and redesigning the oscillator to be less dependent on supply voltage variations should generally suffice.

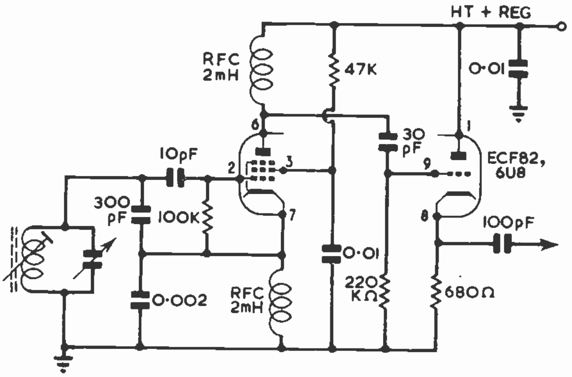


Fig. 8.5. V.f.o. and cathode follower designed to be stable against h.t. voltage and load fluctuations. The choice of components for the tuned circuits obviously depends on the operating frequency, and slight temperature compensation will generally be needed.

(b) **Pulling**

Pulling refers to the effect on the oscillator frequency of one or more subsequent stages whose operating conditions change during the keying cycle. It can be expected if the stage following the oscillator draws grid current (as would a class C buffer or doubler), or if the early stages of the transmitter are tightly coupled. If the oscillator is on the same frequency as the p.a., i.e. where no frequency multiplying or mixing is used, the likelihood of pulling is increased. Yet another cause is a fluctuating load on the oscillator owing to changes in the input impedance of a later stage.

Pulling can invariably be treated by improving the isolation of the oscillator, and a cathode follower is highly recommended for this. The oscillator should be loosely coupled to it through a 30 pF capacitor, and by careful design it should be possible to produce a unit whose output can be short-circuited without shifting the frequency by more than a few cycles. Fig. 8.5 shows a circuit combining a stable oscillator with a cathode follower.

It may be simpler to replace the frequency-determining components in the oscillator with values to halve the original operating frequency and to use a subsequent stage as a doubler. This may, however, reduce the drive to the p.a. to such an extent that a further buffer stage would be necessary.

(c) **R.F. Feedback**

R.f. feedback is a stray signal leaking back from a high-level stage to a previous stage, particularly a variable frequency oscillator. It may have an appreciable effect on the frequency of oscillation, depending on its strength and phase. The presence of feedback may be verified by noting the pulling which occurs on tuning the output circuits of the keyed stage through resonance. The feedback path may be either internal, by virtue of valve capacitances, or external, because of poor constructional layout.

Internal feedback may be treated by the methods outlined for pulling; isolation of the oscillator is of great importance. External feedback is only discovered after a transmitter has been built, and the commonest cause is the p.a. valve and circuitry being close to the oscillator section, owing to in-line layout not being adopted. Here also, pulling is most likely to occur when the p.a. runs on the same frequency as the oscillator, and the cure is either to resort to doubling, as before, or to screen both p.a. and oscillator. Sometimes it is sufficient to mount a metal plate between the two circuits.

It is recommended that any long leads carrying r.f. currents or voltages near the oscillator be screened. In addition, the h.t. line must be bypassed to r.f. by means of series resistance and shunt capacitance.

Spacer-Waves or Backwaves

The spacer wave is the small signal often radiated during key-up conditions. It is common for the spacer wave to be audible in the station receiver, but this does not mean that it is radiated far. If an appreciable spacer wave is radiated, it would make a signal difficult to copy.

Causes are (i) the keyed transmitter stage not being completely cut off, and (ii) leakage of r.f. through the p.a. valve capacitance.

The first is often due to insufficient negative bias being applied in grid-block or screen-grid keying. If drive is applied to the p.a. continuously, the bias required for cut-off during key-up conditions will be the static value obtained from data *plus* the peak value of the drive voltage, which is considerable. In a cathode-keying system, a leaky capacitor in the keying filter may cause a spacer wave.

The second cause obviously worsens as the grid-to-anode capacity of the p.a. valve increases: it is in general more severe with larger valves, and considerably worse with triodes than pentodes. It may be cured either by neutralizing the stage in question or by keying more than one stage.

KEYING METHODS

There are many possible methods of keying, and the choice is largely one of practical convenience, personal preference and, particularly in break-in systems, suitability to the station as a whole. The important requirements of the system adopted are avoidance of the troubles outlined in the previous section, and the operation of all valves in the correct regions of their characteristics.

Almost any stage in the transmitter may be keyed, but there are good reasons for keying the final power amplifier rather than an earlier stage. For instance, if the oscillator is keyed, the requirements of a short time-constant to reduce chirp and a long time-constant to eliminate clicks conflict. If any stage before the p.a. is keyed, with softening, the p.a. may harden the keying, as described earlier in connection with key clicks. In some cases it is useful to key more than one stage sequentially, and this is covered in the section on break-in.

When keying the p.a. of a transmitter designed for anode modulated telephony it is necessary to short-circuit or remove the modulation transformer. If this is not done, there is the risk of damaging this component with the very high voltages which may appear across its windings when the anode current of the keyed stage is switched. There is also a possibility of generating clicks owing to the transformer "ringing."

If, in an anode modulation system, the modulator output valves are fed by the same h.t. supply as the p.a., switching from one mode to the other may be achieved with a single pole change-over switch (see **Fig. 8.6**).

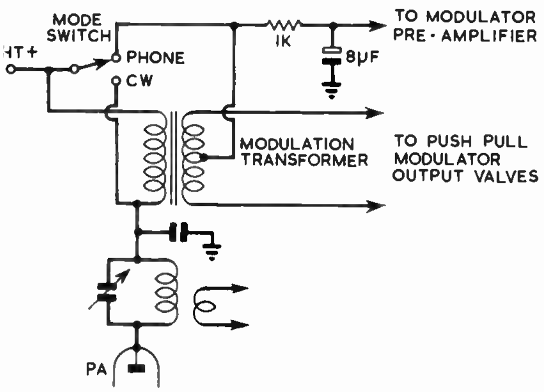


Fig. 8.6. Circuit using a single switch to protect the modulation transformer of an a.m./c.w. transmitter and to control the modulator h.t. supply.

Power Transformer Keying

This is a method rarely used today, owing to many disadvantages. Use of a keying relay is imperative, and it is suited only to very slow keying speeds.

H.T. Supply Keying

A neglected but very reliable method, corresponding to anode modulation in a telephony transmitter (or anode-and-screen modulation if a tetrode or pentode is used), is to key the h.t. supply to the p.a. stage. In the circuit shown in **Fig. 8.7** a filter (L, C, R) is incorporated to produce the desired shape of keying envelope. For reasons of safety, this method usually requires the use of a keying relay.

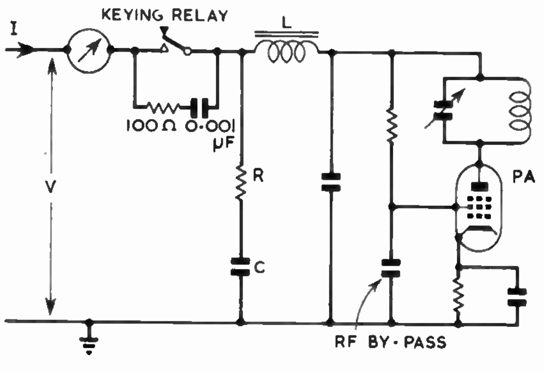


Fig. 8.7. H.t. supply keying. See text for values of L, C, and R. An r.f. filter is connected across the keying relay contacts to suppress local clicks.

Cathode Keying

One of the most popular systems of keying is to insert the key in the cathode lead of one or more stages, often the p.a. alone. It has, however, three disadvantages.

(i) Unless precautions are taken, the cathode potential tends to float when the key is open and may rise to such a value that there is a serious danger of breakdown of the heater/cathode insulation of the valve. To avoid this excessive potential difference a resistor not exceeding 0·25 Megohm should be connected between the cathode and the heater. If drive is applied to the valve when the key is open, the cathode-to-heater voltage will rise to the sum of the cut-off bias and the peak positive drive voltage, and it is therefore desirable to avoid any unnecessary rise by ensuring that the driver stage has good output voltage regulation. If directly heated valves are used, a separate heater winding will be needed for the keyed valve, and unless a limiting resistance equivalent to R1 in **Fig. 8.8** is included, there may be quite high voltages developed across the key.

(ii) Cathode keying is contrary to the practice, particularly in high level amplifiers, of avoiding any impedance in the cathode lead, as this is an accepted cause of parasitic oscillations. It is widely recommended, especially for v.h.f. transmitters, that a short and substantial copper plate be soldered between the cathode connection and chassis. If cathode keying must be used in any stage handling appreciable r.f. currents, the keying lead must be bypassed to r.f. at the cathode pin. About four high-grade capacitors, typically 0·01 μF each, with low internal inductance, should be used in parallel.

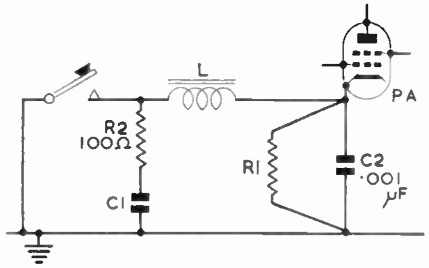


Fig. 8.8. Cathode keying. The rise time is determined by L, a low-resistance choke of 1 to 5 Henrys. The decay time is determined by C1 which can have a value of 0·5 to 2·0μF. R1 is a safety resistor, and should not be more than 0·25 M ohms.

(iii) As the cathode circuit of a p.a. valve carries a high current, any key-click filter must include a series inductor, which will add its internal resistance to the cathode circuit thus introducing an unwanted additional bias. The operating conditions should be adjusted to allow for this.

The optimum component values of the click filter circuit depend partly on the magnitude of the cathode current and are best found by experiment. When the key is depressed, capacitor C1 is short-circuited through the key (the resistance R2 being included to limit the discharge current to a reasonable value) and the capacitor therefore plays no part in determining the rise-time of the carrier. Thus the first step is to choose a suitable inductance value to give an acceptable " make." The value of C1 may then be altered to give the desired decay time, or " break " characteristic.

Grid-Block Keying

This system overcomes all three of the problems associated with cathode keying and is a very satisfactory method. It does, however, require a negative supply of up to −250 volts.

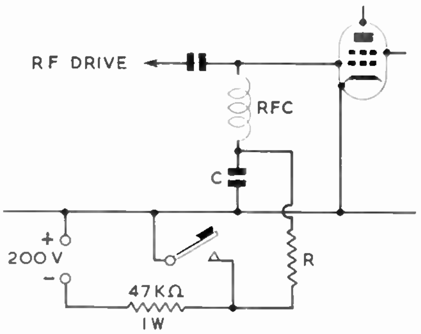


Fig. 8.9. Grid block keying. The resistor R has the normal bias value, and the keying envelope is determined mainly by the time constant of R and C. The product of R (in k ohms) and C (in microfarads) should be about 5. The bias supply should be capable of supplying 4 mA in the key-down condition and should have an open-circuit output of about 200 volts.

If a sufficiently large negative bias is applied to the control grid of a p.a. or other stage, no cathode current will flow even in the presence of drive voltage. For most tetrode or pentode power amplifiers a bias of about 200 volts will be found adequate: for oscillators and other low-level stages, 50 volts or less is sufficient. As there is usually little or no grid current flowing in the stages to be keyed, this method lends itself to key-click filters using R and C components only. A suitable circuit is shown in **Fig. 8.9**. The capacitor C is not only part of the softening circuit, but also serves to decouple any r.f. remaining at the earthy end of the r.f. choke. It should therefore be mounted near the keyed valve, particularly if it is an oscillator.

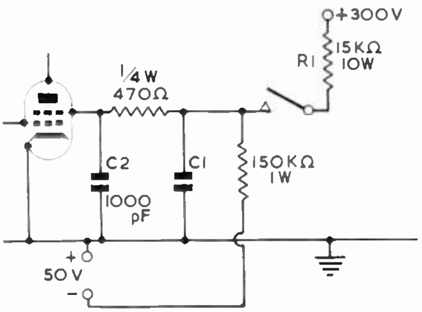


Fig. 8.10. Screen-grid keying. The values shown are suitable for a valve of the 807 class but may vary widely. R1 will be somewhat lower than would be required if it were a simple series dropper resistor. The capacitor C1 is chosen to give suitable hardness to the keying: a typical value is 0·25 µF. C2 is the r.f. bypass capacitor.

In the circuit shown the voltage across the key contacts will rise to 200 volts with the key up, but the impedance is high enough to prevent dangerous currents from flowing through the operator, and a hand key may be safely used. The voltage must not be allowed to exceed the specified maximum d.c. negative control-grid potential, or there is a danger of flashover, particularly in high-slope valves.

Grid-block keying is easily applied to several stages simultaneously, each of them having individual hardness characteristics. By adding a circuit to invert the switching of the blocking bias, the receiver muting may be controlled by a similar system operating on the a.g.c. line. This is discussed further in the section devoted to break-in.

Screen Grid Keying

Screen grid keying permits a p.a. of reasonable power to be keyed with contacts and filter components of low power rating. In its simplest form the method consists of breaking the positive supply line to the screen, but most pentodes and

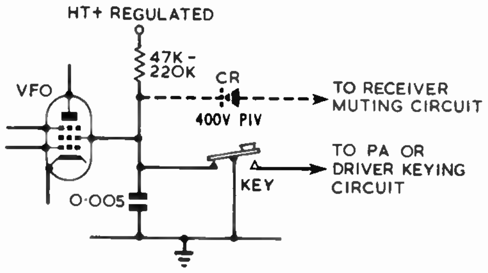


Fig. 8.11. Simple circuit for obtaining differential keying of both oscillator and driver or p.a. using back and front contacts of the key. For full break-in operation, the receiver also may be controlled by the back contacts in the manner shown later in Fig. 8.33, but diode CR must be included to prevent the receiver muting circuit from reducing the v.f.o. screen voltage. This method cannot be used with " bug " keys or many types of el-bug unless a fast keying relay is used to carry out the switching functions.

tetrodes do not completely cease to conduct when the screen is isolated and it is usually necessary to apply a negative bias of the order of 20 to 50 volts in the key-up condition to eliminate the spacer-wave. The bias may be derived from a battery or a separate bias unit.

A typical screen grid keying circuit suitable for valves similar to the 807 is shown in **Fig. 8.10**. It may be worth noting that this arrangement causes the " make " to be harder than the " break," although a measure of such disparity is often favoured.

In a differential keying system, involving the transmitter oscillator, screen grid keying of the oscillator is recommended, as a negative supply is unnecessary. This is particularly suited to control by the back contacts of the key, as shown in **Fig. 8.11**.

Suppressor Grid Keying

The output from a pentode amplifier may be reduced to zero by the application of a sufficiently high negative voltage to the suppressor grid, but this is seldom used as a method of keying. When the suppressor is biased to cut-off the anode current, heavy screen-grid current will flow and the screen dissipation may be exceeded. Moreover, the suppressor potential is a very insensitive form of control in most pentodes, and the keying voltage required is therefore large.

The Application of Keying Methods

In a transmitter not including break-in circuits, it is usual to apply keying in one of two ways. Either the master oscillator and other low-level stages function continuously and the p.a. is controlled, or the oscillator and power amplifier are left on while an intermediate stage is keyed. The latter system has the advantage that low-level stages are

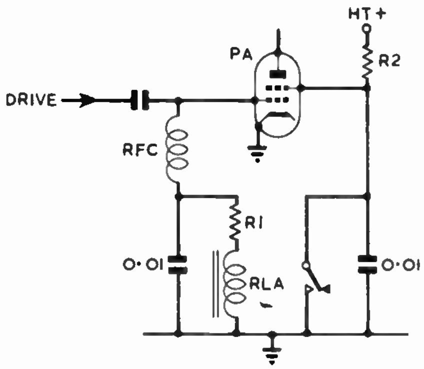


Fig. 8.12. Method of clamp-controlling the p.a. with a relay, which must be a high-speed type as it operates at keying speeds. Its resistance is unimportant provided that it will function solidly on the normal grid current of the p.a. The value of R1 must be such that when added to the relay coil resistance, the rated grid leak resistance is obtained. The value of R2 will depend very much on the p.a. valve used and the h.t. supply voltage: its power rating must allow for the dissipation during long periods in the key-up condition, although this can be reduced by introducing a resistor to the circuit from screen to earth. This need only be sufficiently low to hold the p.a. anode and screen currents to safe values.

easier to key properly, but most class C p.a. stages would need to be protected in some way under key-up conditions.

There are several ways of doing this, which are:

(a) A clamp tube may be added to the screen grid circuit of the p.a.

(b) An equivalent system using a relay operated by the p.a. grid current may be used, and a suitable circuit to do this is shown in Fig. 8.12;

(c) The p.a. can be operated in a linear mode, such as class B or AB, where its operating conditions are such that it passes safe currents, within the valve dissipation limits, when drive is absent. Although the p.a. efficiency would be lowered, harmonic output would be much reduced.

There is little advantage in keying the master oscillator of a transmitter, unless break-in operation is envisaged. Indeed, satisfactory oscillator keying is not easily achieved, and the system should be avoided wherever possible.

Keying Relays

In many systems, the voltages in the part of the circuit where the key is connected are low enough to preclude any possibility of shock to the operator, but if the voltages are higher than about 150 volts, or preferably 75 volts, a keying relay is recommended. Such a relay should be of robust construction, rapid in operation, and its contacts and insulation must be adequate for the current and voltages in the keyed circuit. It should also be reasonably silent; if necessary a noisy relay can be quietened by mounting it on a sheet of rubber or a pad of foamed plastic.

Keying relays are particularly useful where several circuits are keyed simultaneously, but it is unfortunate that robust relays with high current and voltage ratings are usually neither fast nor quiet. Many relays take 10 or more milliseconds to act, and such devices would impose severe restrictions on keying speed. A recent development is the dry reed relay, which is fast, with operate and release times of 1 or 2 mS, is quiet and compact, and can key currents of up to 1 amp. Mercury-wetted relays, with their freedom from contact bounce, may be found suitable for some applications.

When a keying relay is operated from a mains supply unit, there is a possibility that it will hum. This can usually be cured by using rectified a.c. and a generous filter capacitance.

Should a relay be selected which proves to be too slow in operation, it is possible to speed up its action on " make " by driving a large pulse of current through its coil at the beginning of each operation. The circuit of Fig. 8.13 shows a method of achieving this.

Keyer Valves

In a class C amplifier every electrode, except the suppressor grid in a pentode, normally passes a current. This means that the keying circuit must be designed to deal with power and not merely voltage, and efforts must be made to prevent local interference being caused through arcing at the contacts or radiation from the key leads. A further problem is that in most circuits it is difficult to obtain the desired degrees of softness at both make and break with the same component values, and a compromise is necessary.

It is occasionally worthwhile to overcome these drawbacks by introducing an auxiliary valve used as a variable resistance in series with the h.t. supply to some electrode of a convenient valve in the transmitter, the key being used to control the grid bias on the auxiliary valve.

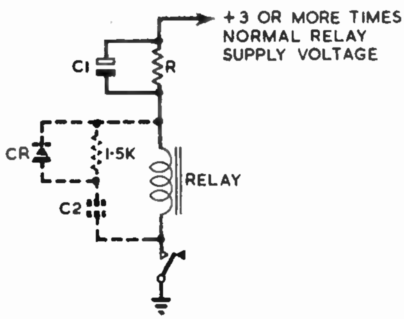


Fig. 8.13. Circuit for speeding up the " make " action of a relay. The resistor R should be such that the steady-state relay current is about its rated figure, and C1 should be a few microfarads. If the relay drop-out action is to be delayed, the diode CR and the 1·5 K ohms resistor prevent the capacitor C2 from affecting the rise in coil current on " make." The value of C2 depends on the required delay, and the relay characteristics, but would again be a few microfarads. Diode CR could be a silicon type rated at 150 p.i.v. or more, such as the OA202.

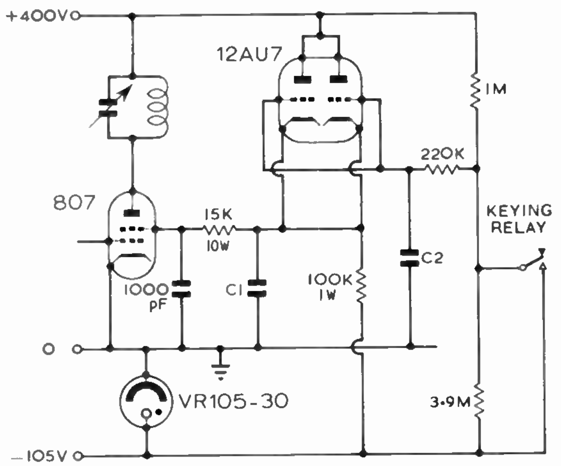


Fig. 8.14. Valve keying of a screen grid supply. This circuit is suitable for inputs of 25-35 watts using a p.a. valve of the 807 class. The capacitors C1 and C2 determine the hardness at "break" and "make" respectively: suitable values for normal keying are C1 = 0·1 μF and C2 = 0·005 μF. A separate heater supply is required for the 12AU7 keyer valve, and the heaters should be strapped to the cathode through a 220K ohms resistor.

Fig. 8.14 shows a circuit where a 12AU7 (with the two sections connected in parallel) is used to key the screen supply to an 807. This method uses a contact on the keying relay which is closed when the relay is unenergised. The steady current flowing through the contacts is only 0·5 mA in the key-up condition, although the open-circuit voltage, i.e., with the key depressed, is about 400 volts. If desired, the current can be reduced by using a higher resistance values in the keyed circuit, but generally there is little advantage in so doing.

If the transmitter is controlled by an electronic key, it is often possible to dispense with relays in this arrangement as suitable voltages may be derived directly from the keyer circuit. With a 12AU7 keyer valve as in Fig. 8.14, the two grids should be at +300 volts to earth with the key down, and −100 volts with the key up.

In a transmitter using clamp-tube efficiency modulation for telephony, it is possible to use the clamp tube itself as a keyer valve, as shown in Fig. 8.15.

Many keyer tube systems have been devised and applied to power amplifier stages. In any circuit of this type, key click suppression or softening is derived from grid-block keying principles.

Change-over Switches and Relays

Switching the station between transmitting and receiving states usually involves several functions, and some of these may need to take place sequentially. Post Office key switches can be used, but relays are more satisfactory.

A multiple-section switch mounted on the front panel of the transmitter or receiver may necessitate long runs of cable carrying r.f.; it is advantageous to have a separate aerial relay, specially designed for the purpose and mounted at a convenient point on the chassis. When other units are involved (linear amplifiers and other accessories) it is more satisfactory to run a single line between them which will control their individual relay switching circuits. This may mean that equipment built at a later date can be accommodated more easily. It is also simpler to adjust the timing of relay circuits (by the use of capacitors, for instance) than multisection switches, and this is important for sequential switching.

Aerial switching can be avoided completely by using separate arrays for transmitter and receiver, but this is not recommended. Few situations allow the erection of two high-gain aerials, and so the communication efficiency of a station using this system is likely to be affected. If separate arrays are used, they should be as far removed from one another as possible, or large voltages may appear on the receiver aerial and resonance effects may drastically reduce the transmitting aerial efficiency. Separate aerials are necessary for full duplex operation, but where different bands are used, these may already be available.

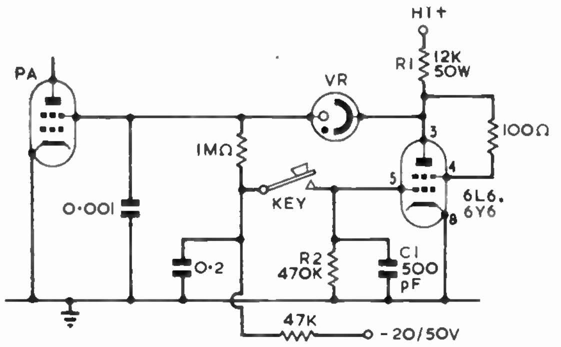


Fig. 8.15. Circuit for using the clamp-tube as a keyer valve. The negative supply may be derived from the p.a. grid circuit provided that drive is applied to the p.a. continuously. This circuit also is designed for valves similar to the 807. For other types, it will be necessary to modify the value of the screen dropper R1. The keying characteristics are determined by C1 and R2.

Transistors for High Voltage Switching

Transistors capable of handling collector voltages up to 800 volts may be employed in keying circuits for valve transmitters. The expense is not usually justified, however, and their usefulness is doubtful for normal keying. They are of greater value if used in conjunction with transistorised electronic keyers, whenever it is found desirable to use solid-state switching throughout.

A common problem is "suck-out" of incoming signals by the transmitter p.a. and noise generated by the p.a. reaching the receiver. Both of these effects may be observed by swinging the output tuned circuit of the transmitter through resonance, and listening for fluctuations in signal strength and noise. The noise problem can be aggravated if the p.a. is not neutralised properly, it is also wise to ensure that it is completely cut off during receiving periods, and by improving screening around the p.a. circuit.

Aerial relays are found in many forms. Heavy duty devices with one change-over contact and one make contact are useful (see Fig. 8.16), but where better isolation and screening are required, co-axial relays are preferable.

Electronic aerial relays (TR switches) are satisfactory, and these are described in the section on break-in, together with other methods of high-speed aerial switching.

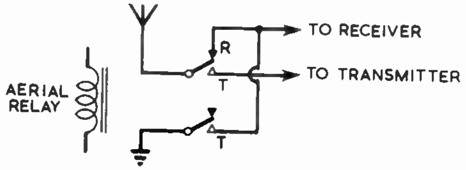


Fig. 8.16. Circuit showing the use of the usual arrangement of contacts on aerial relays.

Control of Linear Amplifiers

Where the transmitting equipment includes a high power linear amplifier, it is desirable to bias such stages to cut-off during listening periods, either because of noise generated in the valves affecting receiver performance, or because the continuous anode dissipation rating is being exceeded. The problem of noise is more likely to occur in a system where an electronic TR switch is used, and the final amplifier is connected directly to the aerial at all times.

A circuit which will switch the grid of a linear amplifier from operating bias to cut-off, regardless of the class of operation, is shown in Fig. 8.17(a). It utilises a change-over type dry reed relay, which switches the grid of the linear amplifier between TRANSMIT and RECEIVE states at a very high speed. The normal, or unenergised, state is that corresponding to the RECEIVE condition, where 200 to 300 volts negative is applied to the valve control grid.

If any method of linear amplifier switching is used, the timing considerations mentioned in connection with aerial switching are relevant. If the linear amplifier is switched on late—that is, after the exciter output has started building up—the envelope of the radiated r.f. signal will rise sharply, corresponding to a click on c.w., or a missed consonant on s.s.b. Similarly, there will be a sharp decay if the linear amplifier is turned off before the softened exciter signal has fallen almost to zero. Hence it is necessary for the circuit controlling the linear amplifier grid to switch on instantaneously, and off after a delay of 20 mS or more. The circuits of Fig. 8.17 switch to TRANSMIT in about 200 μS, and back to RECEIVE in about 20 mS, this delay being determined by the value of C1.

Fig. 8.17(b) shows the principles outlined above adapted to the control of a linear amplifier using two 6146 valves in class AB1.

KEYING MONITORS

In the interests of good sending, it is desirable that an operator listen to his own Morse transmission. This is particularly important when automatic keys are used. Several monitoring methods have received favour, especially the use of the station receiver, and the keying of an audio oscillator simultaneously with the transmitter by means of either d.c. circuits or the r.f. voltage appearing at the transmitter output. Keyed oscillators have the disadvantage that signal characteristics such as chirp, drift, ripple and key-clicks cannot be detected, and that a note of constant pitch causes operator fatigue over long periods. Furthermore, r.f.-controlled devices tend to produce a " weeping " note which may fade out if the transmitter power is reduced, or its operating frequency raised. This can sometimes be overcome by using Zener diode voltage regulation.

On the other hand, keyed audio oscillators are indispensable when duplex operation is used on two frequencies, for obvious reasons. They are also more convenient where c.w. signals are generated in an s.s.b. transmitter by a tone oscillator, as this can be used directly as a monitoring system if the oscillator is coupled to the receiver output stage. Where a separate oscillator is used, the problem of operator fatigue could be overcome by giving the oscillator a rough note by increasing its harmonic output with a non-linear device such as a diode, or by providing a frequency control for periodic adjustment.

Oscillators controlled by d.c. circuits are easily devised, a simple R-C phase-shift circuit with a keyed output stage being sufficient. This output stage could generally be

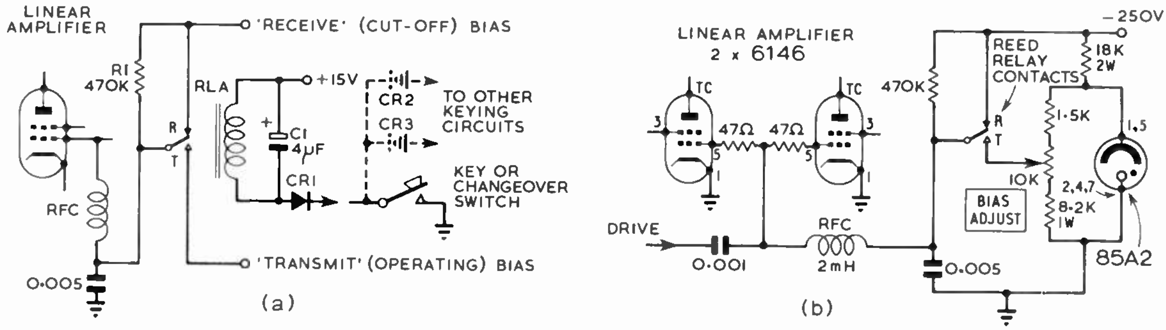


Fig. 8.17. (a) Dry reed switch circuit for linear amplifier control. Diodes CR1, CR2 and CR3 (rated at 400 p.i.v. or more) ensure that different keying circuits do not interact: if other circuits have negative polarity—as in grid-block keying—the diodes and C1 and the 15 volt supply must be reversed in polarity. R1 prevents the linear amplifier grid circuit from floating during armature travel. (b) The same principle adapted for control of a pair of 6146 tetrodes in class AB1. The relay coil circuit is as in (a), and may be switched by the change-over relay or, for break-in operation, by the key itself.

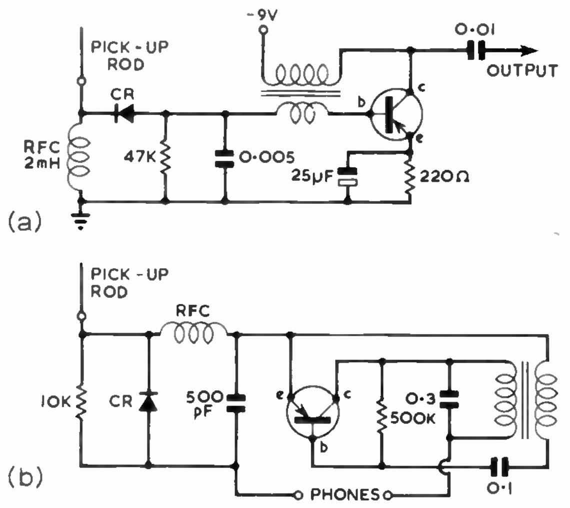


Fig. 8.18. Circuits of transistorised r.f.-operated tone oscillators for transmission monitoring. In (a) the base circuit has applied to it a negative voltage derived from the r.f. signal from the pick-up rod. A battery supplies the power for the collector circuit. In (b) the r.f. provides both facilities. In either circuit diode CR can be almost any germanium type, the transistor a medium-gain audio type and the transformer a Repanco TT9, or similar. If a co-axial feeder is used, a capacitive tap should be made to the centre conductor instead of using a pick-up rod.

connected to the same keying point as the keyed r.f. stage in the transmitter. For a rough note, the oscillator could be a multivibrator.

Fig. 8.18(a) shows an r.f.-controlled transistor oscillator. where a battery provides the power for the circuit, and the rectified input signal is used to switch the base circuit. A circuit in which the sampled r.f. provides power for the transistor is shown in Fig. 8.18(b).

Obviously if a tone oscillator is used for monitoring, the station receiver must be completely silenced during transmission periods. This can be done in several ways, and the choice of a method often depends on the voltages available via spare contacts on the change-over relay. H.t. circuits can be broken, negative blocking bias can be applied to i.f. or a.f. stage grids or positive bias can be applied to cathode circuits. The circuit of Fig. 8.19 is often convenient. VR1 could be replaced by a fixed resistance of, say, 47 K ohms, and it should remain a simple matter to listen to actual radiated signals occasionally.

The method of using the station receiver for monitoring is more popular, but there are three important conditions which must be satisfied.

(a) The signal appearing at the receiver must be as small as possible to prevent overload or even damage. This means, of course, that the receiver must be disconnected from the transmitting aerial, and it may be necessary to short the input terminal to earth. These operations may be carried out by the change-over relay or switch, or by another relay specially designed for aerial switching.

(b) The receiver gain must be further suitably reduced except when an exceptionally good a.g.c. system is available *in the c.w. mode*. Reference should be made to the section devoted to " Muting the Receiver " for details. The method shown in Fig. 8.19 may be used, although the same result may be obtained more conveniently (when grid block keying is used in the transmitter) with the circuits shown in Fig. 8.9 where the contacts on the changeover relay replace the keying relay contacts.

(c) The sequence of switching the transmitter and receiver must be correct. If the transmitter oscillator and driver stages come on just before the receiver is muted, a burst of r.f. will reach the receiver. It is also undesirable for the receiver to be resensitised before the transmitter is switched off. This can be overcome by slightly setting the appropriate contacts on the change-over relay.

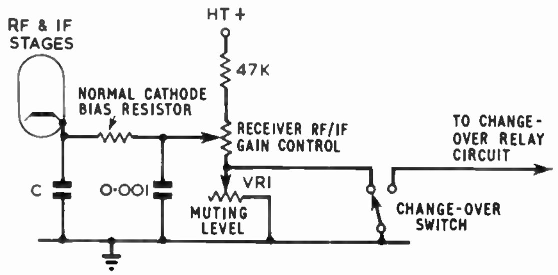


Fig. 8.19. Method of muting the station receiver during transmission periods. VR1 should be 25-50 K ohms. Needing only occasional adjustment, it is usually mounted at the back of (or inside) the receiver cabinet.

MORSE KEYS

The various devices for sending c.w. can be classified as straight keys, sideswipers, semi-automatic (" bug ") keys, electronic semi-automatic and automatic keys (" el-bugs ").

Side-swipers are instruments built with a horizontal arm having a contact on either side of it, and pivoted in a vertical plane. Operation involves the use of each contact alternately as the keying contact, so that dots and dashes are sent on one side and then on the other. Side-swipers are not popular, and tend to produce incorrectly spaced characters.

Semi-automatic or " bug " keys are usually purchased complete, as their construction involves considerable accurate machine work. They are similar in function to side-swiper keys, except that whereas one of the contacts remains fixed and is used to provide dashes, the other initiates oscillations in a spring-loaded rod so that a sequence of dots is produced. With bug keys it is possible to send at much increased speeds, but no such key, electronic or otherwise, will improve an operator's sending.

Electronic Keys

Electronic keys vary in complexity from those solely providing dot sequences like a bug key, to vast machines with typewriter keyboards, using pulse and logic circuits. The first category is represented by Fig. 8.20; the second is outside the scope of this chapter. Between these categories are circuits which produce a series of dots or dashes when the moving arm or paddle is held against one or other of two fixed contacts. In these keyers, once a dot or dash has been started, it will be completed irrespective of what happens to the paddle, and a space of the correct length will be left before the next dot or dash can commence. The timing of the hand movements is thus very much less critical than when using a "straight" key; the result will be either correct Morse or complete gibberish.

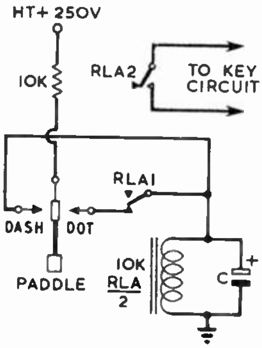


Fig. 8.20. Circuit of a device with the same properties as a "bug" key, producing a sequence of dots electronically, at a speed determined by C (which can be made switchable). Contacts RLA1 are normally closed, contacts RLA2 are normally open.

A simple form of keyer is shown in **Fig. 8.21.** Each half of the double-triode valve is biased beyond cut-off by the potential dividers R7, R8 and R9, R10. When the paddle is moved on to either contact, a positive voltage is applied to the two grids. Both triodes then conduct and both relays operate. Relay A has a "break" contact (RLA1) which isolates the grids from the paddle, while relay B has a "make" contact (RLB1) which keys the transmitter. The grids are held positive by the charge on C1, but this charge leaks away through R4 and the speed control R6. The potentiometers R7 and R9 are adjusted so that the cathode of V2 is held more positive than that of V1; therefore as the grid voltage falls, relay B releases before relay A. The transmitter is switched off when relay B releases, but the cycle of operations cannot repeat until relay A has released, thus giving the required space between the letter elements.

When the paddle closes on the dash contact, a higher positive voltage is applied than when it closes on the dot contact. The time of discharge of the capacitor C1 through R4 and R6 is therefore longer, although the space period remains unchanged.

Some care may be required in selecting suitable relays. The design of relay B is not critical, except that it must operate quickly and firmly on about 5 mA; relay A must also operate solidly on about 5 mA but it should not open too quickly, because C1 may then not be fully charged before the circuit is broken by the A relay contact. Post Office Type 600 relays with 3000 to 6000 ohm coils are recommended although they tend to be noisy. High speed relays could be used, but it may be necessary to shunt the coils of relay A with a capacitor to retard its operation.

To adjust the keyer, first set R7 and R9 so that V1 and V2 are both just beyond cut-off. Hold the paddle in the dash position, adjust the speed control R6 for a convenient rate of sending and then set R9 for the correct on/off ratio for dashes (i.e., three units on, one unit off). Now hold the paddle in the dot position and adjust R5 for the correct dot ratio (i.e. equal times on and off). This may be done aurally, or with an oscilloscope, or by observing the anode current of a keyed stage in the transmitter; provided that the current is zero in the key-up condition, a string of dots should result in an average anode current equal to half the steady carrier value, while a series of dashes should give three-quarters.

A very comprehensive electronic key is shown in **Fig. 8.22.** It consists basically of a variable-frequency multivibrator whose output is switchable to various stores and bistable circuits (flip-flops). It produces self-completing dots and dashes of correct length, at sending speeds from four to 40 words per minute. The output waveforms of the circuit are shown in **Fig. 8.23.**

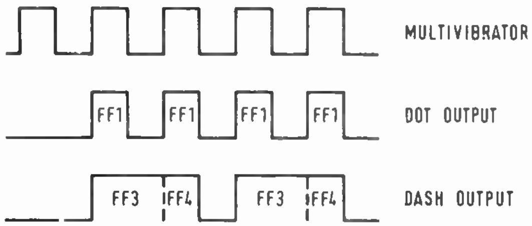


Fig. 8.23. The output of the various timing circuits of the key. FF1 refers to the dot-timer flip-flop, FF3 to the dash-timer and FF4 to the dash lengthener.

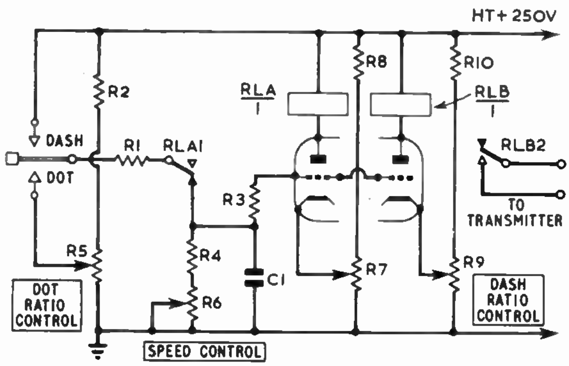


Fig. 8.21. The OZ7BO electronic keyer. The relays A and B may be Post Office Type 600 with 3000 to 6000 ohm coils. Relay RLA has one break contact, and relay RLB (the keying relay) has one or more make contacts. The valve may be a 6SN7 or 12AU7.

When the dot contact on the key is closed, the dot-storage bistable circuit (flip-flop 2) is switched on and the gate between the multivibrator and dot timer (flip-flop 1) is opened. The key then sends dots for as long as the key contact remains closed. When the contact is opened the last dot completes itself and resets the dot-storage bistable.

Closing the dash contact causes the corresponding dash gate to open, and the key sends dashes. These are formed by one pulse from flip-flop 3 whose length is equal to the length of the multivibrator period (two dot lengths), followed by another from flip-flop 4 equal to a single dot, thus making three dot lengths in all.

It is interesting to note that the dot-storage bistable can be switched on during a dash, and the dot is sent immediately after the dash (letter N). Also the key may be pressed to the dash side after storing a dot, and the letter A is sent. Two

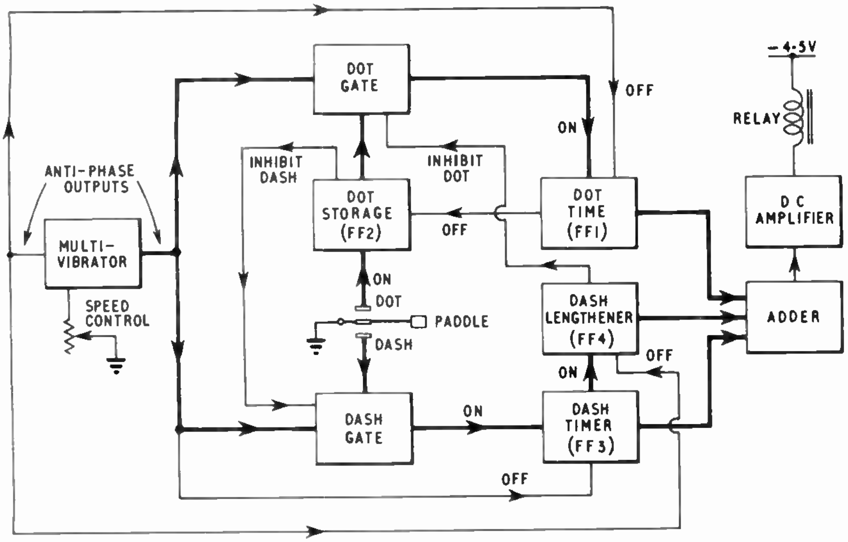
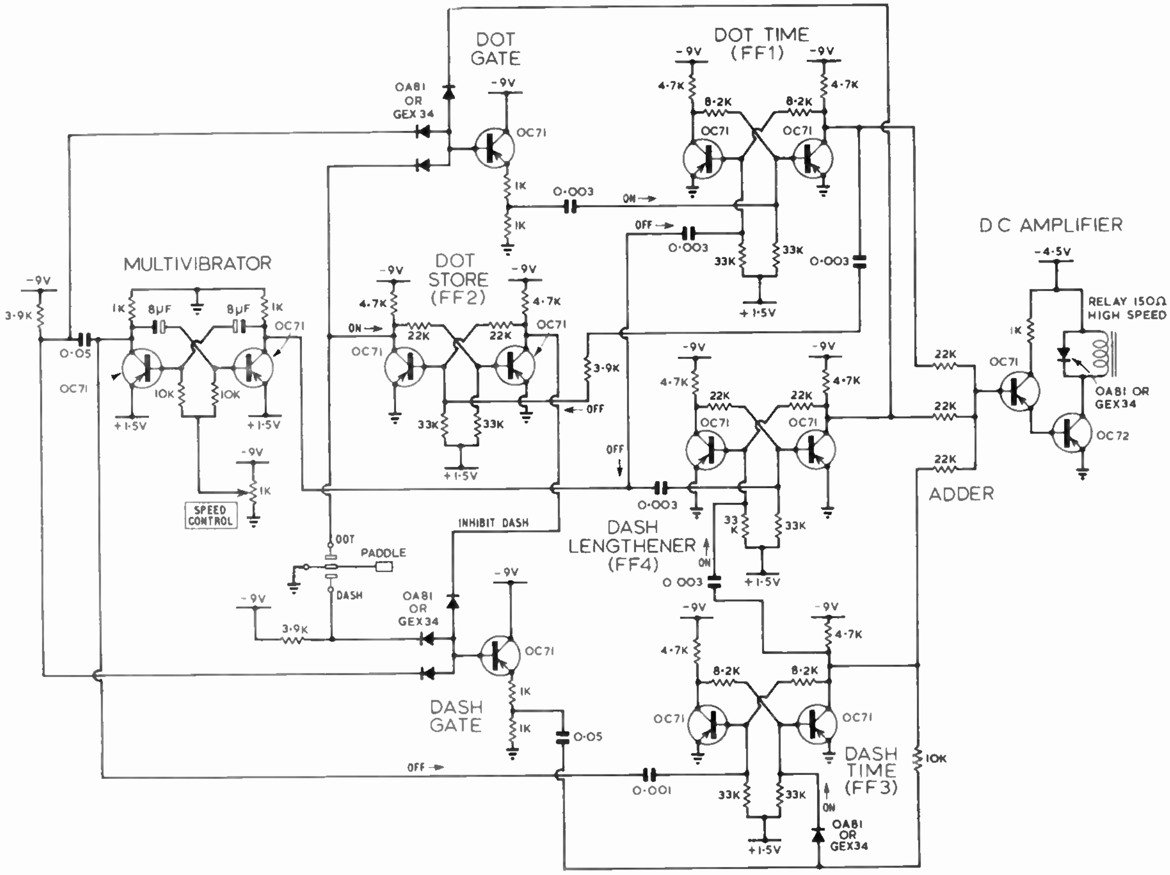


Fig. 8.22. Block diagram and circuit diagram of the G3PEW electronic keyer. The characters are self-completing and accurately timed by locking to a multivibrator



" inhibit lines " prevent dots and dashes being sent at the same time.

The timing of this key is extremely accurate, as the beginning and end of each character is locked to the multivibrator. The operating speed of the circuit is controlled by variation of the multivibrator frequency.

Tone Keying

Many applications exist for devices which enable keyed audio tones to control a transmitter operating in the c.w. mode. For instance, tape recordings can be used for slow Morse transmissions, automatic CQ machines and transmission at very high speeds.

All that is required is an a.f. amplifier to bring the output of the tape-recorder or receiver up to a suitable level, a simple rectifier, and a valve or transistor switch operating a keying relay. **Fig. 8.24(a)** and **(b)** show simple circuits to achieve this, which will key up to 30 words per minute with the specified relay, and a 3 kc/s input signal. The keying speed is limited by the time constant of the smoothing circuit in the rectifier section (C1 and R1), and the operating time of the relay.

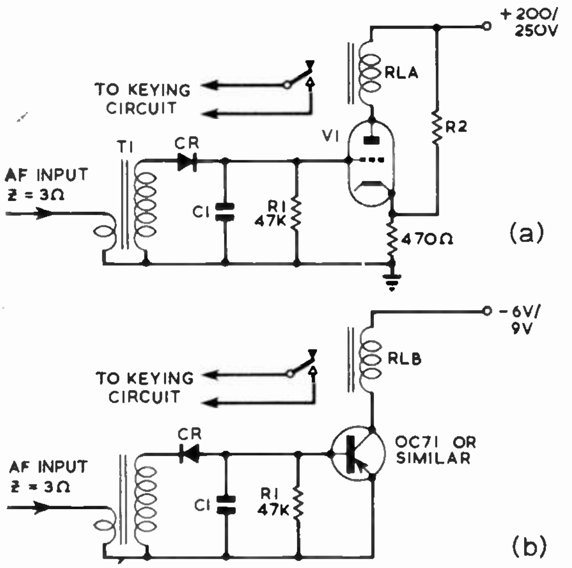


Fig. 8.24. (a) Valve circuit for keying a transmitter by means of a tape-recorder or receiver. Almost any triode may be used (e.g. EC90), and R2 should be adjusted so that under no signal conditions, the relay is on the point of tripping in. C1 and R1 form the rectifier smoothing circuit, and also limit the highest keying speed which may be used. C1 should therefore be as small as possible, and may be about 0·1 μF. Diode CR may be almost any semi-conductor type. T1 is a standard output transformer, and the relay should be a high-impedance high-speed type. (b) A similar circuit using a transistor as a switch. Relay RLB would need to be of lower impedance and higher sensitivity, but otherwise the considerations are the same as for the valve circuit.

The block diagram of a more complicated unit which has been devised for higher keying speeds is shown in **Fig. 8.25.** The principal advantage of this circuit is that by using a square-wave signal from the limiter, followed by full-wave rectification, the smoothing capacitance C required is very small and consequently the keying speed may be high (above 100 words per minute if a suitable relay is used). A dry reed switch would be capable of operation up to almost 250 w.p.m., but for most purposes a standard high-speed type would suffice.

BREAK-IN

Before describing full break-in systems a simple system which is often incorrectly referred to as break-in will be mentioned. This refers to methods of switching the change-over relay to transmit by means of the key, but having it return automatically to the receive state after a fixed delay of about one second. This method has been adopted in cases where oscillator stability had not been found high enough for real break-in operation, and where the energy used in throwing a toggle switch was considered wasted. This control system does not, of course, permit continuous monitoring of the transmitter channel. The circuit in **Fig. 8.26** shows one method of acquiring this rather dubious facility.

Full-Break-In

The essential feature of a break-in system is that the operator is able to receive incoming signals in the spaces between his own transmitted Morse characters so that duplex operation becomes possible. Much time is thereby saved, particularly in multiple contacts or when interference is present, and the advantages in contests or when hunting DX are obvious. Transmissions are interspersed with the sign " BK " and the call-signs are given at frequent intervals to conform with the conditions of the licence.

When using break-in, the normal change-over and keying functions are controlled by the key, and they must take place in the right sequence. The station should return to the receiving condition at the sensitivity level required by the operator between each dot and dash of the transmitted message.

It is not easy to install a good break-in system, one of the problems being that of keying the oscillator stage. This can be avoided in two ways: either the oscillator is screened so well that it is inaudible in the station receiver, or a mixer-type v.f.o. with a keyed mixer is used. With either of these methods, the master oscillator can be made to run continuously. However, as it is difficult to screen a v.f.o. to the extent required, and as mixer-v.f.o.'s are not common except in single-sideband transmitters, it is more usual to find break-in circuits involving oscillator keying.

In the section on change-over methods it was pointed out that the transmitter oscillator and buffer stages should be switched on after the receiver is desensitised, and off before it is turned back on. Further timing considerations are important in designing a break-in circuit, and the desired switching sequences are generalised in **Fig. 8.27.** This shows that the receiver must be disconnected from the transmitting aerial (or the aerial must be switched from receiver to transmitter) just before radiation commences, and it must remain so until after the r.f. output has fallen to zero. The oscillator and receiver muting must be timed relative to the keyed stage in the same way. In fact aerial, oscillator and receiver may be switched together, and it is then only necessary to provide a delay before they return to the receive state, as the softening of keying automatically ensures that they are in the transmit condition before the r.f. output rises.

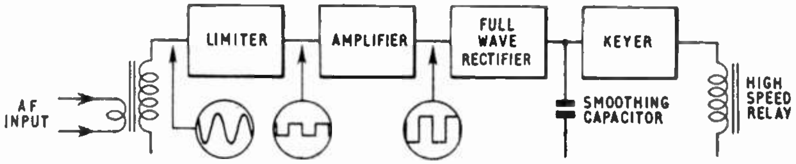


Fig. 8.25. Block diagram of a high speed tone keyer.

The delay time at the end of each Morse character will naturally depend on the degree of softening used, but from 10 to 20 milliseconds is adequate for most situations. Three methods of obtaining this delay are in common use: utilisation of the back contacts of a hand key, suitable timing of relay circuits, and application of bistable pulse circuits such as the Schmitt trigger unit. The first system precludes the use of bug-keys and many el-bugs, the second demands one or more keying relays (with their own inherent disadvantages) and the third tends to be more complicated.

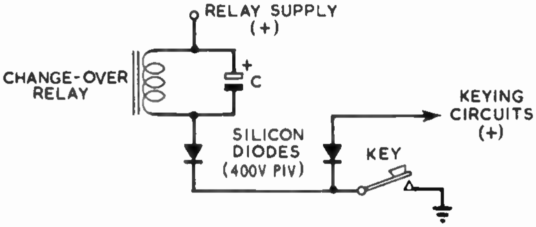


Fig. 8.26. Circuit for returning the station to the receive state a given time after the end of a transmission. The relay is selected to suit the supply voltage available, and capacitor C is given approximately by 20/R microfarads, where R is the relay coil resistance.

Fast Aerial Switching

A simple expedient often used in a break-in system is to erect a separate aerial for the receiver, but the disadvantages of this method have been mentioned elsewhere. A single aerial may be switched by an accurately timed keying relay, but fast high-current aerial relays are scarce.

Alternatively, a transmit/receive switch of the contactless type may be installed. T/R switches may be divided into two classes: (a) those which require the application of some control voltage or current synchronised with the key, and (b) those which act as limiters and are entirely automatic. With either type, the transmitter is connected permanently to the aerial, the receiver being connected to it through the T/R switch. In the key-down condition the T/R switch must not absorb any appreciable fraction of the transmitted output power, nor must it permit any excessive amount of power to reach the receiver. The switch is usually arranged so that it presents a high impedance in the key-down condition, and is connected to the aerial at a point of low impedance, and thus of low voltage.

The T/R switches in class (a) have the disadvantage that they require a high negative blocking bias voltage although they reduce the problems of harmonic generation and TVI inherent in class (b). Both types suffer from p.a. suck-out and noise generated by the p.a.

A switch in either class usually consists of a sharp cut-off valve, often in cathode follower or grounded-grid configuration, so arranged that it will withstand the full transmitter output without being damaged. This requirement precludes the provision of any useful amount of impedance step-up in its grid circuit, and the noise figure of the receiver may be worsened. The best form of circuit is probably the grounded-grid arrangement using a small high-slope triode (see Fig. 8.28). This may be muted either by applying a large negative bias to its grid or by breaking its d.c. cathode return lead, which ever is the more easily adaptable to the muting system used for the main receiver.

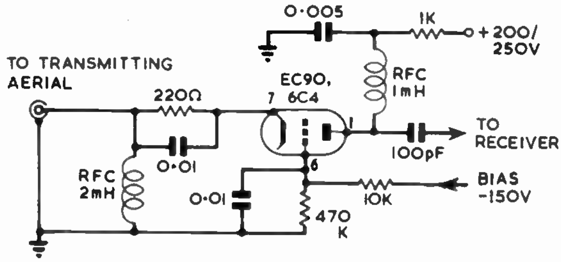


Fig. 8.28. D.c.-controlled T/R switch using an EC90 (6C4). The s.w.r. on the transmitting aerial feeder must be as low as possible, and the output power limited to about 75 watts. If it would be an advantage to use a circuit with appreciable gain, the r.f. choke in the anode circuit could be replaced by tuned circuits for the bands in use.

A carrier power of 100 watts in a 75 ohm circuit gives a peak voltage of about 125 volts (provided there are no standing waves); at least this amount of bias must be applied to prevent conduction in the T/R switch valve, and the peak grid/cathode voltage with the carrier applied will therefore be 250 to 300 volts. This is beyond the rating of most high-slope valves, but the majority of samples will withstand it without premature failure.

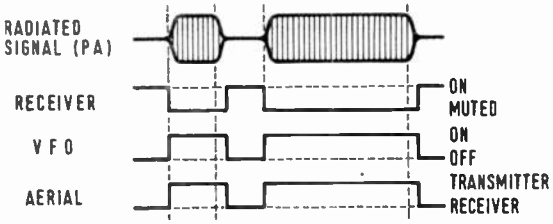


Fig. 8.27. Diagram showing the required switching sequence for a typical break-in system. Note the delay in returning the v.f.o., aerial and receiver to the receive state at the end of a character.

The heater/cathode insulation is subject to voltages of the same order, and it may be necessary to insert r.f. chokes in the heater supply leads, each heater pin being bypassed to the cathode with a 1000 pF capacitor.

Other forms of amplifier may be preferred but it should be remembered that pentodes generate more valve noise than triodes, while triodes in circuits other than the grounded-grid circuit suffer from excessive breakthrough owing to the anode—grid or grid—cathode capacitance.

The circuit in Fig. 8.28 is suitable for power up to about 75 watts. For powers up to 200 watts in a low impedance line

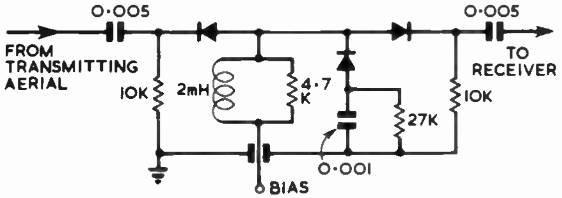


Fig. 8.29. Diode T/R switch due to VE2AUB. The receiving bias is +30 volts, and the transmitting bias −350 volts, although the switch offers some protection without this. It may be convenient to reverse diode and bias polarities to suit switching systems already in other equipments.

with a reasonable s.w.r., the semiconductor circuit of Fig. 8.29 is recommended. The silicon diodes may be any type with a very low capacitance at reverse voltages about 30 volts, and with a peak inverse rating of 600 volts or more. The insertion loss is quoted as less than one decibel in the receive condition and up to 80db in the transmitting state. The receiving bias should be about +30 volts, and the transmitting bias about −350 volts.

Fig. 8.30 shows a circuit which is switched off by the voltage generated when the r.f. signal causes current to flow through the grid bias resistor; it possesses useful gain and has a low output impedance. The design of the tuned circuit L1, C1 will depend on the bands to be covered.

Aperiodic T/R switches with gain may be beneficial to receivers of poor sensitivity, but they may lead to trouble with cross-modulation or overload as the pre-selectivity gain is increased.

All T/R switches involving tuned circuits, however, need adjustment when large frequency changes are made. The circuits working on the self-limiting principle should be well screened to reduce radiation of harmonics and should be

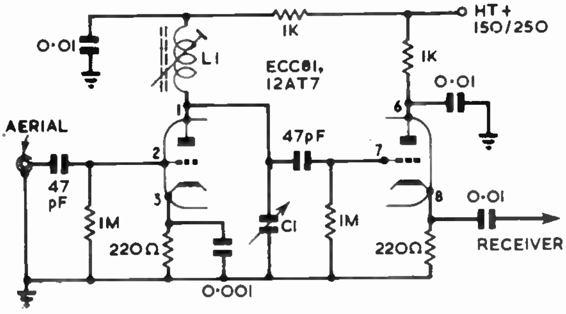


Fig. 8.30. A further self-limiting T/R switch, with fair gain and low output impedance. The design of the tuned circuit L1, C1 will depend on the bands to be covered.

connected to the aerial *between* the transmitter and low-pass filter or aerial tuning unit. Circuits using diode limiters may cause strong harmonics of a local broadcasting station to be heard in the receiver: rejection circuits tuned to its frequency would overcome this defect.

It is recommended that T/R switches be completely screened and mounted on the back of the receiver or transmitter cabinet. The coaxial lead to the receiver should be as short as possible to avoid resonance effects and to minimise pick-up of the transmitted signal.

Dry Reed T/R Switches

Electronic T/R switches can be made to act exceedingly fast, operating times of 50 microseconds being readily attained. A system which is rather slower (yet 10 or 20

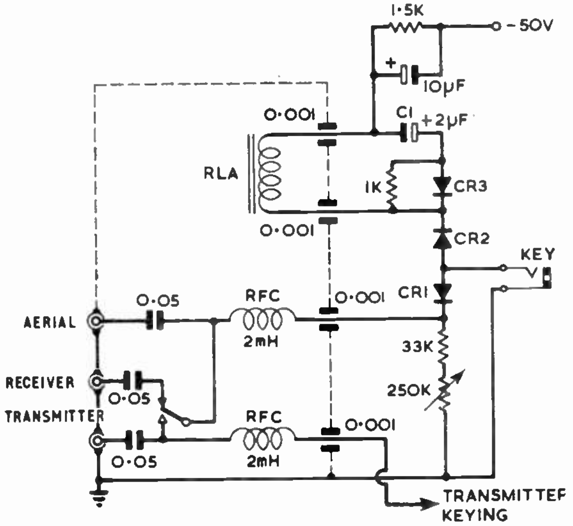


Fig. 8.31. A dry reed switch used in a modified version of a T/R switch devised by VE3AU. Here sequential keying of the aerial and other circuits is obtained by (a) the use of the relay contacts to switch both d.c. and r.f. and (b) the release delay for the relay due to C1. Diodes CR1 and CR2 isolate the relay and transmitter switching circuits, and CR3 together with the electrolytic capacitors provides the fast attack and slow release characteristics of the relay. As mentioned in connection with the diode T/R switch of Fig. 8.29, it is simple to adapt the circuit for keying systems of the opposite polarity. The circuit shown is for negative keying lines: positive voltages may be controlled by reversing all three diodes, both electrolytic capacitors and the 50 volt supply.

times faster than normal keying relays), but which avoids most of the problems peculiar to electronic devices, is one using dry-reed switches.

Fig. 8.31 shows a circuit which ensures that the aerial is switched at the right times relative to the transmitter. Unlike previous circuits, it switches the aerial to and from the receiver and transmitter, and in times of less than one millisecond. Dry-reed relays will *switch* only quite low currents, but if the switching operations are completed before the r.f. current appears, they can pass much higher currents through the closed contacts. The circuit shown will handle 700 watts into a well-matched 50 ohm line.

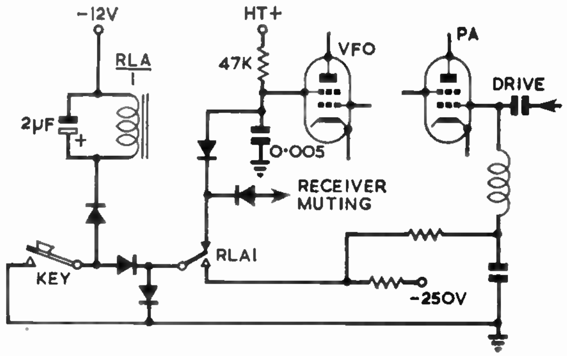


Fig. 8.32. A complete break-in keying system with sequential keying. When the key is closed, it operates relay RLA. The relay contacts first switch on the v.f.o. and mute the receiver, then switch on the p.a. by shorting out the blocking bias. When the key is released, it immediately initiates the blocking of the p.a., and after a delay determined by the 2 μF capacitor, the relay returns the v.f.o. and receiver to the " receive " state. The circuit is suitable only for a limited number of keying and muting methods, although careful rearrangement of the diodes gives a certain amount of flexibility.

Transmitter Keying Requirements

It has already been mentioned that it is necessary to key the master oscillator unless it is exceedingly well screened or unless it works on mixing principles, but it is difficult to achieve perfect keying characteristics from the v.f.o. alone. It is usual to find the v.f.o. and the p.a. or driver keyed differentially. This method requires the v.f.o. to be keyed as " hard " as possible, with softening applied to the later stage. Furthermore, as shown by Fig. 8.28, the v.f.o. must be held on for a time while the keyed amplifier is being exponentially turned off: it must then be switched off sharply, to prevent radiation reaching the receiver.

A reliable method is to use a keying relay with at least two pairs of contacts, one pair switching the v.f.o. and another the keyed amplifier stage. The contact springs should be set so that the v.f.o. contacts close as soon as the armature starts to move, the amplifier contact closing later in its travel. On release the sequence is reversed so that the v.f.o. stays on for a short time after the amplifier contacts have opened. A difficulty is that the time interval between the operations of the two pairs of contacts is limited to a few milliseconds, and therefore the keying cannot be made very soft.

The circuit in Fig. 8.32 is more flexible in this respect, and uses a high speed relay. It includes several silicon diodes (400 p.i.v.) and is rather limited in the types of keying which can be applied to the various units. For instance, the p.a. must be grid-block keyed, but where high-speed relays are used, this is inadvisable because of their low current-handling capabilities.

Muting the Receiver

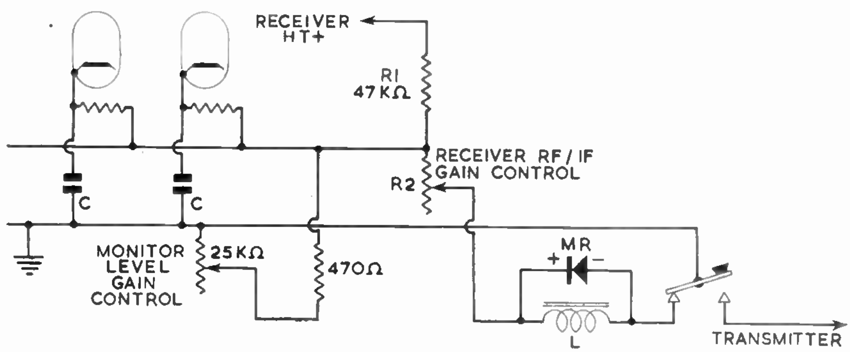
When the key is depressed, the receiver must lose sensitivity quickly enough to prevent a click being produced from the leading edge of the transmitted signal: when the key is raised the gain should remain low until radiation has ceased and then rise rapidly.

If the receiver is used for monitoring, the amount of desensitisation required will depend on the transmitter power, the type of aerial switching used, and the effectiveness of any screening. As it is generally insufficient to cut off the receiver r.f. stage alone, the i.f. stages are usually controlled with it. It is often convenient to do this by breaking the cathode return of the r.f. and i.f. stages and replacing the normal r.f./i.f. gain control with a MONITOR LEVEL control of higher resistance value. The normal control can be restored in parallel with this through the back contacts of the key or a pair of similar contacts on the keying relay, adjusted to open at the same time or a little earlier than the instant when the v.f.o. contacts close. Fig. 8.33 shows a circuit of this kind.

In order to obtain a sufficiently rapid fall in sensitivity where the key is depressed, it may be necessary to reduce the values of the r.f. and i.f. cathode decoupling capacitors C in the receiver to the minimum required for proper working, or in mild cases to reduce the value of the bleed resistor R1 which usually exists between the h.t. line and the cathode line, thus reducing the charging time of the cathode capacitors.

When the key is released, the gain will rise rapidly owing to the relatively low value of the normal gain control potentiometer R2 which forms the main discharge path for the cathode capacitors. If this process is not delayed by timing of relay circuits or by another method, the choke L of several Henrys inductance may be added, with a diode acting as a spark suppressor. The d.c. resistance of this choke, if too high, may result in a considerable reduction in the maximum gain of the receiver unless the bias resistors are suitably adjusted.

Fig. 8.33. Receiver muting by the method of connecting a monitor gain or " listen through " control in the cathode line of controlled r.f. and i.f. stages. The choke L should have an inductance of 2-20 Henrys depending on the rate of recovery required. CR is a silicon diode rated at 400 p.i.v. or more, and serves as a spark quencher.



Another method of receiver desensitisation, using the a.g.c. line, is convenient in conjunction with grid-block keying systems. The line must be isolated from the large capacitance (C2) determining the gain control time constants, as shown in Fig. 8.34.

If the receiver recovery is not delayed by external circuits, it is possible to achieve this by including a diode at a suitable point in the control line so that once the muting voltage disappears, the main discharge paths for the grid capacitors C are removed.

With either of these systems, trouble may arise from the beat frequency oscillator in the receiver. If the b.f.o. injection point is followed by a gain-controlled stage, the high b.f.o. voltage at the second detector will be modulated by the muting voltage, and for fast rates of muting this will produce a click or thump at the output. The remedy is either to inject the b.f.o. voltage at a later stage or to remove the gain control from the stage in question.

Another problem may arise from the use of a crystal filter placed at the beginning of the i.f. amplifier chain, e.g., in the anode circuit of the mixer. In such a receiver, the crystal filter is followed by at least one gain-controlled stage, and in the key-down condition the signal voltage at the crystal may be many times that which corresponds to a reasonable output when the receiver is being operated at maximum sensitivity. Due to the very small bandwidth (very high Q value) of the filter this voltage cannot die away quickly, even after the transmitter output has fallen to zero, and if the gain of the following stages of the receiver is allowed to increase rapidly to normal after the key is released, this signal will be heard as a loud click in the receiver immediately after the instant of break. A possible remedy is to avoid controlling the i.f. gain and to rely on controlling the r.f. gain, if by doing so it is possible to reduce the overall gain sufficiently. Otherwise the receiver gain must be held down until the crystal filter has recovered.

Complete Break-In Systems

A common requirement is the need to add break-in facilities to commercial equipment without any drastic modifications, and this usually means controlling voltages of both polarities. In this instance a keying relay must be resorted to, or its equivalent using the properties of diodes, neons, or keyer valves. The complexity of the circuit is closely related to the precision of control, and flexibility, required of the system: a simple circuit involving the back contact of a hand key is unsatisfactory by timing considerations, and would for instance prevent a guest operator from using his own " bug " key.

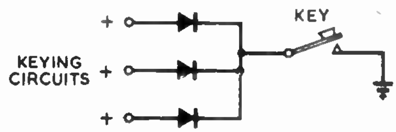


Fig. 8.35. Method of switching several voltages of the same polarity with a single switch contact.

In designing a break-in circuit there are several points to observe. As has been shown in several instances already, it is possible to switch any number of voltages of the *same* polarity with a single key or relay contact by isolating each control line with a small diode, suitably connected (see Fig. 8.35). It is helpful therefore to use keying methods which have identically polarized control lines from stages which need to be switched simultaneously.

T/R switches of the self-limiting type should be used where possible, as this reduces the number of circuits needing precise control. Otherwise the choice should be between diode and triode types requiring keyed bias voltages, and dry reed relay circuits.

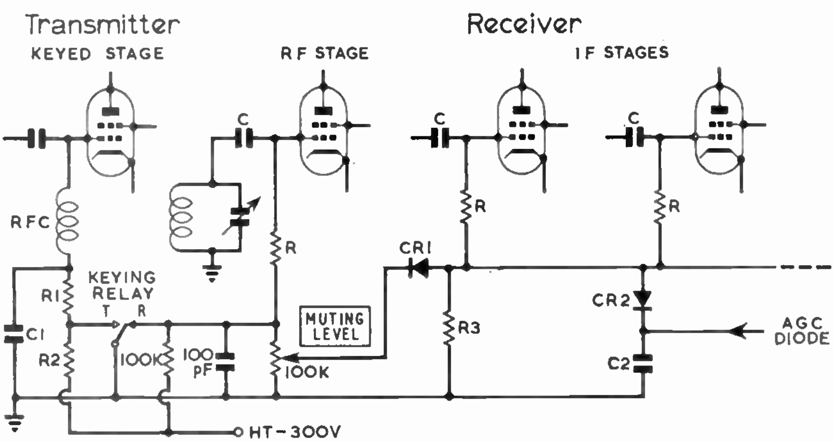


Fig. 8.34. Receiver muting system designed around a transmitter grid-block keying circuit. C1, R1 and R2 are the normal key filter components. In the " transmit " state, the voltage across the muting level potentiometer (about −120 volts) cuts off the receiver r.f. stage, and a fraction of it, determined by the monitor level control, reduces the gain of the i.f. stages. Diodes CR1 and CR2 isolate the desensitising voltage from the a.g.c. delay capacitor C2. R3 is part of the original a.g.c. circuit, but if the receiver is one where the a.g.c. line is earthed when the b.f.o. is on, CR2 must also isolate the shorting switch from the controlled stages. As the receiver is held in the muted state for a short time only after the key is released (due to the travelling time of the keying relay armature), the softening that can be applied to the p.a. keying is somewhat limited. Also the receiver controlled-stage grid components C and R should be as small as possible, so that the receiver is muted as soon as possible after the key is pressed. Typical values for these are 50 pF and 100K ohms respectively.

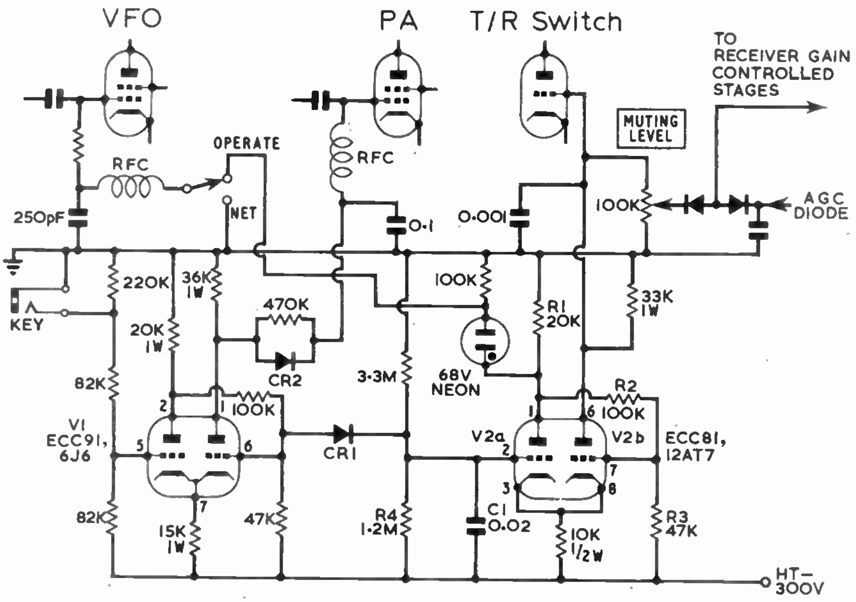


Fig. 8.36. The G3PDM break-in circuit. The diodes are silicon types (150 p.i.v. or more), R4 and R5 are given approximate values only, and resistors are all ½ watt rating unless otherwise stated.

If a switching voltage is available which is of the right amplitude and polarity for controlling a keyed stage, but which is of the wrong phase relative to the transmitted characters, it may be inverted by an auxiliary valve being switched from saturation to cut-off. This principle may be developed to the extent of using pulse techniques, and devices (such as the Schmitt trigger circuit) which produce two keying voltages in anti-phase. In this case, one is used to control the transmitter v.f.o. and p.a., and the other the receiver and T/R switch.

It is commonly believed that a v.f.o. can be held " on " while the transmitter output is falling by using the exponential discharge of capacitance through resistance. This is a fallacy, for such a method would result in chirp, clicks, or distressing noises from the receiver, or a combination of these.

Break-In with the Schmitt Trigger Circuit

A break-in circuit using the Schmitt trigger is shown in Fig. 8.36. This provides comprehensive differential keying on the blocked-grid principle.

The circuit uses two double valves in bi-stable (flip-flop) form. The first controls the p.a. alone, and the second the v.f.o., receiver and T/R switch. Negative bias is obtained from the valve anodes by connecting the positive side of their h.t. supply to earth, and the negative rail to the cathode circuits. When a triode section is cut off the bias across its anode load falls to zero (except with V2a, where a 68 volt neon isolates the v.f.o. from the small voltage across R1 due to the leakage current through R1, R2 and R3).

V2 is switched to the transmit state by a pulse from V1 passing through CR1. When the key is released V1 returns to the receive condition immediately, whereas the return of V2 is delayed until C1 has discharged through R4.

The immediate advantages of this break-in method are that (a) it operates quickly, an attack time of 10 microseconds being typical, (b) there are no keying relays to produce noise, r.f. clicks and contact bounce, and (c) the relative timing of circuits is easily adjusted by means of C1.

CHAPTER 9

MODULATION SYSTEMS

A RADIO-FREQUENCY carrier wave can be used for the transmission of speech by subjecting it to a process of modulation. The carrier wave itself is simply an alternating r.f. current of constant amplitude and constant frequency, and the modulation may take the form of a periodic variation of its amplitude, its frequency or its phase. In all cases the rate of variation is low compared with the frequency of the carrier wave.

If the carrier wave is represented by the equation

$$e = E_0 \sin (2\pi ft + \theta)$$

the variation of E_0 gives rise to amplitude modulation. Similarly, variation of f is known as frequency modulation and variation of θ as phase modulation.

Amplitude modulation is the most common modulation system used in amateur radio. Frequency modulation is of course widely used for high fidelity broadcasting and may be used by the amateur radio operator under certain conditions.

Principles of Amplitude Modulation

No matter what system is employed, the process of modulation produces additional frequencies above and below that of the carrier wave. Thus the modulated carrier wave consists of a band of frequencies as distinct from the single frequency of the carrier. The bandwidth depends upon the modulation system and the frequency of the modulating signals.

The bands of frequencies produced above and below the carrier wave frequency by complex modulating signals (i.e. those composed of many different frequencies as in speech or music) are known as the upper and lower *sidebands* respectively. In the case of amplitude modulation the highest sideband frequency is equal to the *sum* of the carrier frequency and the highest modulation frequency; similarly the lowest sideband frequency is the *difference* between the carrier frequency and the highest modulation frequency. Thus the total bandwidth occupied is equal to *twice* the highest frequency in the modulating signal. For example, if the highest frequency in the modulating signal is 15 kc/s and the carrier frequency is 1000 kc/s, the sidebands will extend from 1015 kc/s to 985 kc/s. The total bandwidth occupied is therefore 30 kc/s.

Modulation Depth

The amplitude-modulated wave is shown graphically in **Fig. 9.1.** Here (a) represents the unmodulated carrier wave of constant amplitude and frequency which when modulated by the audio-frequency wave (b) acquires a varying amplitude as shown at (c). This is the modulated carrier wave, and the two curved lines touching the crests of the modulated carrier wave constitute the *modulation envelope.* The modulation amplitude is represented by either x or y (which in most cases can be assumed to be equal) and the ratio of this to the amplitude of the unmodulated carrier wave z is known as the *modulation depth* or *modulation factor.* This ratio may also be expressed as a percentage. When the amplitude of the modulating signal is increased as at (d), the condition (e) is reached where the negative peak of the modulating signal has reduced the amplitude of the carrier

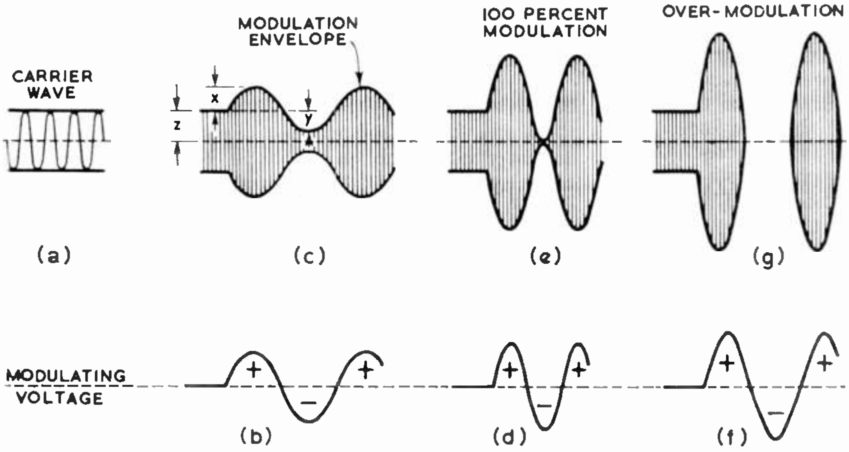


Fig. 9.1. Graphical representation of amplitude modulation: (a) unmodulated carrier wave; (b) modulating signal; (c) modulated carrier wave; (d) (e) 100 per cent modulation; (f) (g) over-modulation.

to zero, while the positive peak increases the carrier amplitude to twice the unmodulated value. This represents 100 per cent modulation, or a modulation factor of 1. Further increase of the modulating signal amplitude as indicated by (f) produces the condition (g) where the carrier wave is reduced to zero for an appreciable period by the negative peaks of the modulating signal. This condition is known as *over-modulation*. The breaking up of the carrier in this way causes distortion and the introduction of harmonics of the modulating frequencies which will be radiated as spurious sidebands; this causes the transmission to occupy a much greater bandwidth than necessary, and considerable interference is likely to be experienced in nearby receivers. The radiation of such spurious sidebands by over-modulation (sometimes known as *splatter* or *spitch*) must be avoided at all costs.

Modulation Power

In the special case of a sinusoidal modulating signal corresponding to a single pure tone, it can be proved mathematically that the effective power in such a wave at 100 per cent modulation is 1·5 times the unmodulated carrier power. Thus, in order to modulate the carrier fully with a sinusoidal wave, the average power in it must be increased by 50 per cent. This extra power must be supplied from the modulator. For example, to modulate fully a radio-frequency stage operating with a d.c. power input of 150 watts, the amount of audio-frequency power required would be 75 watts.

It must not be assumed, however, that the aerial current of a fully modulated transmission will increase by 50 per cent. The relationship between the modulated and unmodulated aerial current for sine-wave modulation is given by:

$$I_m = I_0\sqrt{1 + \frac{m^2}{2}}$$

where I_m = r.m.s. value of modulated aerial current,
I_0 = r.m.s. value of unmodulated aerial current,
m = modulation factor.

Thus, for 100 per cent modulation by a sinusoidal signal:

$$I_m = I_0\sqrt{1+\tfrac{1}{2}}$$
$$= 1{\cdot}226\ I_0$$

In other words, the aerial current will increase by 22·6 per cent.

The position is somewhat different when the modulating signal consists of the "peaky" waveform of speech. Assuming that the peaks drive the transmitter into full modulation, the 22·6 per cent increase will occur at these peaks, but for most of the remainder of the time the modulation depth is much lower; the increase in aerial current will also be much lower. The average modulation depth when the peaks are fully modulated will be of the order of 30 per cent,

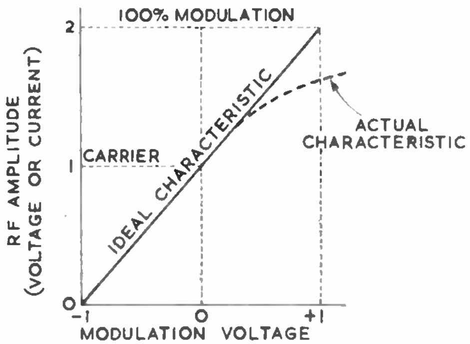


Fig. 9.2. The ideal modulation characteristic is a straight line. Often the actual characteristic is found to be non-linear, and the form of non-linearity shown here is due to the failure of the r.f. voltage (or current) amplitude to rise to twice its normal value at the positive peaks of the modulating signal.

and the average increase in aerial current as seen on a typical ammeter will then be only a few per cent.

Modulation depth may also be expressed in terms of the a.f. power actually supplied and the unmodulated d.c. power supply to the modulated stage, thus:

$$m = \sqrt{2A/W}$$

where A = a.f. power supplied.
W = d.c. input power.

Table 9.1 gives the values, calculated from this expression, for the amount of a.f. power required for various depths of modulation. It will be seen that to produce 70 per cent modulation requires only one half of the a.f. power required for 100 per cent modulation. The corresponding increases in aerial current are shown in the same table.

TABLE 9.1
Effect of Modulation on Aerial Current

Depth of Modulation (per cent)	Ratio: $\frac{\text{a.f. power}}{\text{d.c. power}}$	Increase in aerial current (per cent)
100	0·5	22·6
90	0·405	18·5
80	0·32	15·1
70	0·245	11·5
60	0·18	8·6
50	0·125	6·0

Linearity of Modulation

Ideally, for all modulation depths up to 100 per cent the difference between the amplitude of the r.f. output at the crests and at troughs of the modulation involved should be proportional to the amplitude of the modulating signal; i.e., the modulation characteristic should be linear.

Non-linearity is most often manifest as a flattening of the crests of the modulation waveform, and this causes considerable distortion. It may be minimized by careful design and correct adjustment of the modulated stage, particularly with respect to the amount of r.f. grid drive and the aerial loading. Linear and non-linear modulation characteristics are shown in **Fig. 9.2.**

Bandwidth of a Modulated Wave

For the faithful reproduction of speech and music it is necessary to transmit frequencies in the whole range of the audio spectrum (i.e. approximately 40–15,000 cycles per second). The total bandwidth for this purpose would therefore be 30 kc/s. For a communication system, however, it is the intelligibility and not the fidelity which is of prime importance, and experience has shown that for the intelligible transmission of speech it is sufficient to transmit frequencies up to about 2·5 or 3 kc/s. Thus the transmitted bandwidth need not exceed approximately 5 kc/s. In the overcrowded conditions of the present-day amateur bands it is obviously important to ensure that no transmission occupies a greater bandwidth than is necessary for intelligible communication.

Modulating Impedance

The impedance that an r.f. stage which is being modulated presents to the source of the modulating signal, i.e. the modulator, is called "*modulating impedance.*" It is the ratio of the anode voltage and anode current of the r.f. stage or:

$$Zm = \frac{Va}{Ia} \times 1000$$

where Zm = modulating impedance in ohms
Va = anode voltage of r.f. stage
Ia = anode current of r.f. stage (in mA).

Mathematical Representation of Sidebands

The mathematical equation for a carrier wave of constant frequency which is amplitude-modulated by a signal of constant frequency is:

$$e = E_0 (1 + m. \sin 2\pi f_m.t) \sin 2\pi f_c.t$$

where m = modulation factor,
f_m = frequency of modulating signal,
f_c = frequency of carrier wave,
E_0 = amplitude of unmodulated carrier

This equation may be expanded, giving:

$$e = E_0 \sin 2\pi f_c.t + m\frac{E_0}{2}\cos 2\pi(f_c - f_m)t - m\frac{E_0}{2}\cos 2\pi(f_c + f_m)t$$

Inspection of this expanded form shows that it is made up of three separate terms. The first, $E_0 \sin 2\pi f_c.t$, represents the original carrier,

while $m \frac{E_0}{2} \cos 2\pi(f_c - f_m)t$ and $m \frac{E_0}{2} \cos 2\pi(f_c + f_m)t$

correspond to the lower and upper side-frequencies respectively which are the result of applying a modulating signal of frequency f_m. The total bandwidth of this amplitude modulated wave is $(f_c + f_m) - (f_c - f_m)$ or $2f_m$; i.e. the bandwidth is equal to twice the modulating frequency.

It should be noted from the last equation that the carrier wave is not fundamentally essential to communication since

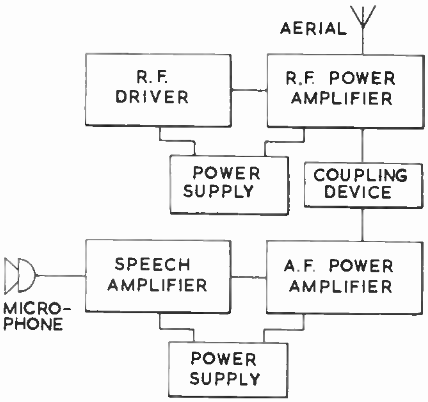


Fig. 9.3. Block diagram of an amplitude-modulated telephony transmitter.

all the intelligence is contained in the sidebands. The carrier wave can therefore be suppressed and need not be transmitted; indeed, it is sufficient to transmit only one of the sidebands. This is known as *single sideband* operation (s.s.b.). Although it requires more complicated equipment than ordinary amplitude modulation, it has the obvious advantage of transmitting the same intelligence within a smaller bandwidth and without a carrier. (See Chapter 10).

Basic Modulation Equipment

In the foregoing paragraphs the fundamentals of the process of amplitude modulation have been reviewed. It thus now becomes possible to specify the modulation equipment required for a communications transmitter. This consists of:

(a) An a.f. power amplifier capable of developing the required power over the minimum frequency range for intelligible speech, and its associated d.c. power unit.
(b) A means of coupling, with correct matching of impedances, the a.f. power to the r.f. stage.
(c) A microphone.
(d) A speech amplifier to amplify the output of the microphone to a suitable level to drive the a.f. power amplifier.

The manner in which these elements are associated in a typical communications transmitter is illustrated in the block diagram shown in **Fig. 9.3.** The term *modulator* is often applied to the whole of the a.f. amplifying section, but strictly speaking it should be applied only to the final stage of the a.f. amplifier.

PRINCIPLES OF FREQUENCY MODULATION

Frequency modulation may be represented graphically as shown in **Fig. 9.4.** Here (A) represents the unmodulated carrier wave, (B) the modulating signal, and (C) the frequency-modulated carrier wave. It will be seen that the frequency of the carrier wave is increased and decreased in sympathy with the amplitude and polarity of the modulating signal, but the envelope amplitude is unaffected. The change in the frequency of the carrier wave is known as the *frequency deviation* or simply as the *deviation*. The total excursion of this frequency is called the *swing:* i.e. the swing is twice the deviation. The carrier frequency, for obvious reasons, is known in f.m. practice as the *centre frequency*. The deviation is proportional to the amplitude of the modulating signal, so that the limits of the swing are determined by the peaks of the modulating voltage. The rate at which the carrier frequency

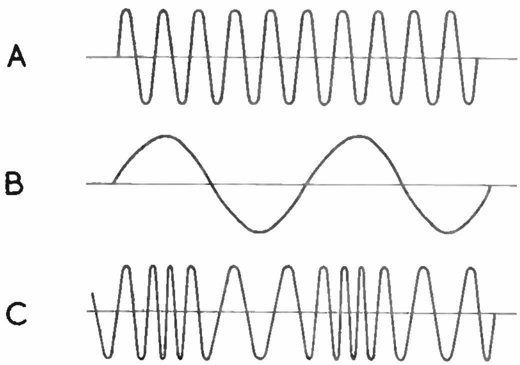


Fig. 9.4. Graphical representation of frequency modulation. A—Unmodulated carrier wave. B—Modulating signal. C—Frequency-modulated carrier wave (i.e. its frequency is varied in sympathy with B).

is deviated is equal to the frequency of the modulating signal. To quote a numerical example, if a carrier wave of frequency 7075 kc/s (i.e. the centre frequency is 7075 kc/s) is modulated by a 3 kc/s tone of specified amplitude to produce a deviation of 2·5 kc/s, the carrier frequency will swing between 7072·5 kc/s and 7077·5 kc/s and back (i.e. the swing is 5 kc/s) 3000 times in one second. If the amplitude of the 3 kc/s tone were doubled the carrier frequency would swing between 7070·0 and 7080·0 kc/s but the rate of variation would still be 3 kc/s.

The ratio of the deviation to the frequency of the modulating signal is known as the *modulation index*. This ratio is obviously not constant since the deviation depends on the amplitude of the modulating signal, and its limiting value, or the ratio of the maximum deviation to the highest modulating frequencies, is called the *deviation ratio*. In the example quoted earlier, the deviation ratio is 2·5 kc/s divided by 3 kc/s or 0·833 for the first given amplitude and 5·0 kc/s divided by 3 kc/s or 1·67 when the amplitude is doubled.

As in the case of amplitude modulation, sidebands are produced during the process of frequency modulation but with the important fundamental difference that a theoretically infinite number of side frequencies may exist. The side frequencies occur in pairs, one on each side of the carrier, and in the case of modulation by a single tone the spacing between the side frequencies on either side of the centre frequency will be equal to the frequency of the modulating tone. In the example just quoted of the frequency modulation of a 7075 kc/s carrier by a 3 kc/s tone, the first pair of side frequencies will occur at frequencies of 7078 kc/s and 7072 kc/s, the second pair at 7081 kc/s and 7069 kc/s, the third pair at 7084 kc/s and 7066 kc/s, and so on.

It should be appreciated that this concept of a number of sidebands and a fixed central carrier is an analytical representation of the effect of varying the frequency of the carrier, and the mental image of the process should not be confused by the supposition that the carrier passes through the frequencies occupied by these sidebands in the course of its frequency deviation.

The amplitudes of the sidebands produced are proportional to the modulation index; i.e. the amplitudes of the higher-order side frequencies (fourth and above) increase from a value which is negligible when the modulation index is 1·0 to appreciable values when the modulation index is greater than 5. The manner in which the sidebands vary is shown in Fig. 9.5 where their variation is plotted for values of the modulation index up to 5. It will also be seen that when

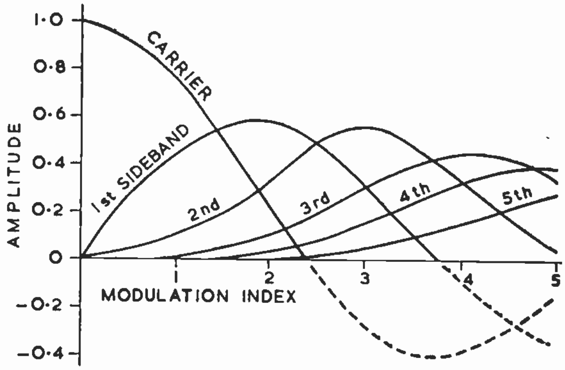


Fig. 9.5. Variation in relative amplitude of the carrier and the low-order sidebands in a frequency modulation system.

the modulation index is 2·4, the carrier itself *disappears*, i.e. its phase reverses compared with the phase when modulation is absent. Further disappearances occur at higher values of modulation index.

It follows therefore that the question of the bandwidth required in the transmitter and receiver of an f.m. system has not such a straightforward answer as in the case of amplitude modulation.

In frequency modulation there is no condition equivalent to over-modulation. An increase in the amplitude of the modulating signal will merely cause an increase in the deviation produced. In the case of large deviations this effect will of course, introduce difficult problems into the design of the r.f. circuits of the transmitter and in the receiver, but these conditions do not occur in amateur communication equipment where the deviation is required by the terms of the Amateur (Sound) Licence to be kept relatively small.

At present the restrictions imposed by the British Post Office on the use of frequency modulation by amateurs are as follows:

(a) The maximum deviation shall not exceed 2·5 kc/s.
(b) The maximum effective modulating frequency shall be limited to 4 kc/s and the a.f. input to the frequency modulator at frequencies higher than 4 kc/s shall be not less than 26 db below the maximum input at lower frequencies.
(c) The carrier frequency shall be at least 10 kc/s within the limits of the frequency band in use.

The deviation ratio permitted is therefore 2·5 kc/s divided by 4 kc/s or 0·625. At this value, the carrier and the first-order sidebands are appreciable, and the second-order sidebands are almost small enough to be neglected while the third and higher orders may be ignored. Therefore, to a first approximation the bandwidth occupied may be considered as equal to twice the deviation or 5 kc/s. In other words, it is roughly equivalent to the bandwidth required for a "communication-quality" amplitude-modulated transmission. In contrast to the large deviations used in high-fidelity broadcasting, for which an international standard of 75 kc/s has been adopted, this restricted mode of operation has come to be known as *narrow-band frequency modulation*, generally abbreviated to n.b.f.m.

AMPLITUDE MODULATION SYSTEMS

As already defined, amplitude modulation consists of variation of the envelope of the r.f. carrier in accordance with the waveform of the intelligence which it is desired to transmit. There are several possible ways of applying amplitude modulation to a carrier wave. They can be classified according to the manner in which the modulating voltage is applied to the stage to be modulated, giving the following broad classifications:

(a) Anode modulation
(b) Control-grid modulation
(c) Suppressor-grid modulation
(d) Screen-grid modulation
(e) Cathode modulation.

Modulation should be applied only to an r.f. stage which is driven from a source of constant frequency, e.g. a crystal oscillator or a stable variable-frequency oscillator followed by at least one isolating stage. This is to prevent the production of spurious frequency modulation such as would result from the direct modulation of a self-excited oscillator. In the case of a triode p.a. stage, neutralization must be perfect, and where neutralization is not employed, as in some tetrode or pentode power amplifiers, great care must be taken to avoid parasitic oscillation which may occur only on the peaks of modulation.

Anode Modulation

Anode modulation of a class C r.f. stage is accomplished by superimposing the modulating voltage on the d.c. anode voltage. For full modulation the anode voltage of the p.a. is swung from zero to twice its normal value, thus causing a corresponding variation in the amplitude of the r.f. output. If the r.f. stage is operated under the correct conditions, the r.f. output current will be proportional to the anode voltage applied to the p.a., and the modulation will be distortionless.

The most straightforward means of anode modulation is the use of a transformer known as a *modulation transformer*

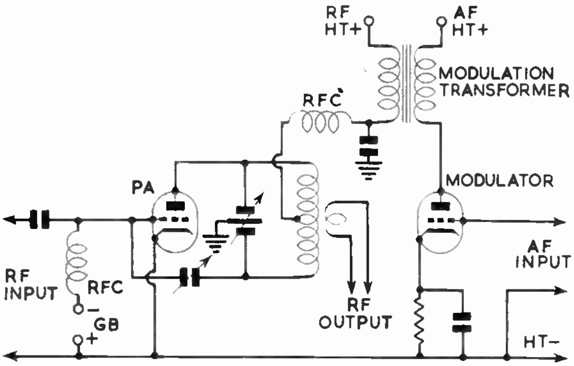


Fig. 9.6. Anode modulation using transformer coupling. The system is sometimes referred to as " transformer modulation."

to couple the modulator to the p.a. stage: the basic circuit arrangement is shown in **Fig. 9.6.** The ratio of the transformer should be such that it will match the impedance of the modulated r.f. stage to the load required by the modulator valve(s). Although in Fig. 9.6 a single triode modulator is shown, it should be realized that triodes, tetrodes or pentodes either in push-pull or single-ended operation, may be used.

In order to obtain linear modulation the r.f. amplifier should be carefully adjusted in accordance with the operating conditions recommended by the manufacturer for the particular valves concerned. In general, it can be said that the r.f. stage must operate under Class C conditions at peak

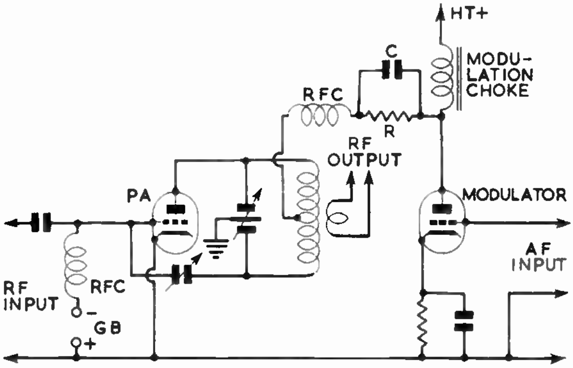


Fig. 9.7. Choke modulation of a class C r.f. amplifier.

output, and the aerial loading must be adjusted to provide the correct modulating impedance.

Although the *effective* power input is modulated, the *average* power input remains constant because each positive excursion of the anode voltage is immediately followed by a corresponding negative one. This means that the anode current, measured by a moving-coil d.c. meter, does not vary during modulation because such a meter measures average current. However, the aerial current is usually measured on a thermal meter which indicates the r.m.s. value and, as was shown earlier, the effective or r.m.s. value of the aerial current does depend upon the amount of modulation. Although the thermal ammeter in the aerial circuit should show a rise of current on modulation, the moving-coil " average-current " meter in the anode circuit should remain steady, and any movement of this meter during modulation indicates incorrect operation.

Anode modulation is the commonest form used, and it is probably the simplest system to adjust and operate. Moreover it is a system which enables the r.f. amplifier to operate at a constant high efficiency, and is capable of full modulation with the least amount of distortion. For most valves the output rating, under anode-modulation conditions, is of the order of two-thirds of the maximum output under c.w. conditions.

Anode modulation may be achieved by other methods which are not so common but nevertheless are sometimes very convenient. These are:

(a) Choke modulation.
(b) Series modulation.

Choke modulation. Choke modulation employs a choke as the coupling impedance between the modulator and the r.f. stage, as shown in **Fig. 9.7.**

The modulation choke must have an impedance at audio frequencies which is high compared with the impedance of the modulated r.f. stage when carrying the combined anode currents of that stage and the modulator. Obviously in this system both valves operate at the same h.t. voltage. The undistorted output voltage from the modulator which is developed across the choke must be less than the h.t. voltage and therefore it cannot modulate the r.f. stage to 100 per cent. For this reason the p.a. stage is run at a lower h.t. voltage than that of the modulator, the excess voltage being dropped across the resistor *R* which is bypassed at audio frequency by the capacitor *C*. The h.t. voltage on the p.a. stage should be reduced by this means to about 60–70 per cent of the modulator h.t. voltage. Since a push-pull arrangement in the modulator is obviously precluded, the modulator must therefore operate under class A conditions. This may necessitate the use of very large modulator valves, or several in parallel, and for this reason choke modulation is only economic for low-power operation, e.g. up to about 15 watts input to the r.f. amplifier.

Series Modulation. In this case the modulator and the p.a. valves are connected in series across one h.t. supply, as shown in **Fig. 9.8.** Thus both will pass the same current, the method of adjustment being to vary the operating conditions (i.e. the amount of grid drive, the aerial loading and the grid bias of the p.a. stage) so that it is operating at a lower anode-to-cathode voltage than the modulator. This system has the advantage that one power unit feeds both the p.a. and the modulator, and that no modulation transformer or choke is required. Valves such as the 807 can be used in both positions in a series-modulation system from

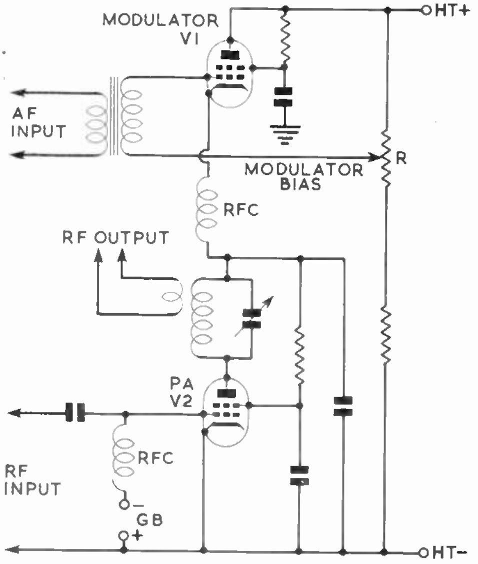


Fig. 9.8. Series modulation. The relative positions of the a.f. modulator VI and the r.f. output valve V2 between the h.t. line and earth can be interchanged if desired. The appropriate negative bias on the control grid of VI with respect to its cathode is obtained by adjusting the potentiometer R.

a 1000 volts supply. The operating conditions of the p.a. stage should be such that there are 500–600 volts across the modulator, and from 400–500 volts across the p.a. For higher power two such valves may be used in parallel as the modulator, and two more either in push-pull or in parallel as the p.a. Although Fig. 9.8 shows the modulator connected to the positive end of the supply, the modulator and p.a. may be interchanged so that the modulator is earthed and the p.a. is connected to the positive supply terminal. It is immaterial which arrangement is used as long as appropriate precautions are taken, e.g. the valve connected to the positive end of the power supply must have a separate heater transformer capable of withstanding the total h.t. voltage, and its grid circuit (a.f. or r.f. as the case may be) must have transformers or capacitors with insulation rated for this voltage.

True anode modulation can be applied only to triodes. The characteristics of tetrodes and pentodes demand that both the screen and anode voltages are modulated to obtain a linear modulation characteristic. This may be done as in **Fig. 9.9** where it will be seen that the screen-grid voltage is obtained from the modulated anode voltage through a series resistance *R*. A potentiometer cannot be used for this purpose. The screen bypass capacitor *C* must have a reasonably high reactance at audio frequencies in order not to bypass the modulation voltages. The modulator load impedance in this case is the ratio of the anode voltage to the sum of the anode and screen currents, and the audio power required is one half of the product of the anode voltage and the sum of the anode and screen currents.

Anode and screen modulation may also be achieved by the use of a special modulation transformer which has two secondary windings. The anode supply is taken through one winding and the screen supply through the other. Such transformers are however not generally available.

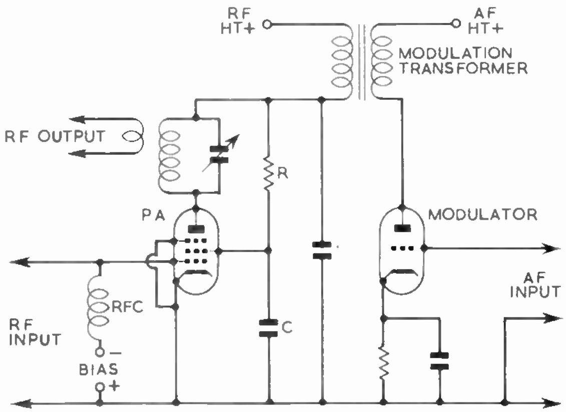


Fig. 9.9. Anode-and-screen modulation. When the modulated stage consists of a pentode or a tetrode, the screen voltage as well as the anode voltage should be modulated by a proportionate amount in order to ensure linear modulation.

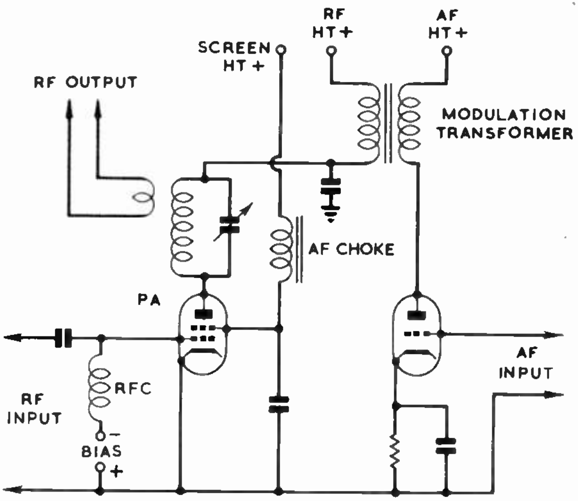


Fig. 9.10. Anode modulation of a tetrode by the use of an inductance in the screen circuit.

Most tetrodes can be satisfactorily anode and screen modulated by the addition of a small inductance in the screen circuit as shown in **Fig. 9.10.** The screen voltage is self-modulated by virtue of the small variation in the screen current which results from the variation (i.e. modulation) of the anode voltage.

The reactance of the inductance should not be less than the impedance of the screen (ratio of the screen voltage and screen current). Generally a small 10–15 H smoothing choke rated at 60 mA is adequate.

Efficiency Modulation Systems

The remaining methods of amplitude modulation are usually grouped together under the general heading of *efficiency modulation systems*. In these systems the efficiency (i.e. the ratio between the r.f. power output and the d.c. power input) is varied by altering the voltage on either the control, screen or suppressor grid, whereas in anode modulation the efficiency is constant and the required variation in r.f. output is obtained by altering the power input. The term *efficiency modulation* should not be taken to imply that this system results in a high efficiency: on the contrary, the overall average efficiency is low. The maximum efficiency for a class C stage is of the order of 70 per cent in anode-modulation systems and this remains constant throughout the modulation cycle. This figure is only achieved, however, on the peaks of modulation in efficiency-modulation systems; the average efficiency is only one half of this figure, and frequently a little lower, say about 30 per cent. The average r.f. output power for a given d.c. input power to the p.a. is appreciably less than for anode modulation. However, the extra r.f. power output which is delivered at modulation peaks is a result of the higher operating efficiency at these peaks rather than the effect of additional audio power from the modulator. In efficiency modulation systems the audio power required is generally insignificant compared with the d.c. input to the r.f. stage.

Grid Modulation

This is probably the most common form of efficiency modulation. It may be applied to practically any type of valve, whether triode, tetrode or pentode. A circuit arrangement is shown in **Fig. 9.11** in which it will be seen that the audio-frequency voltage is applied in series with the r.f. voltage to the grid of the r.f. amplifier. With this system of modulation it is more difficult to make adjustments which will give full modulation with low distortion than in the case of anode modulation. The requirements are as follows:

(a) The r.f. drive must have good regulation, i.e. the r.f. grid voltage must remain substantially constant in spite of the varying grid-current load of the r.f. amplifier. This is generally arranged by providing much more drive than is really necessary and dissipating the excess in a lamp or other resistance load.

(b) The modulator must have a low-impedance output to avoid waveform distortion due to the varying grid load of the r.f. amplifier. A triode operating in class A is preferable, but if a tetrode or pentode is used it is advisable to connect a swamping resistor across the secondary of the modulation transformer.

(c) The grid-bias supply must have good voltage regulation; either battery bias or a voltage-stabilized supply is required.

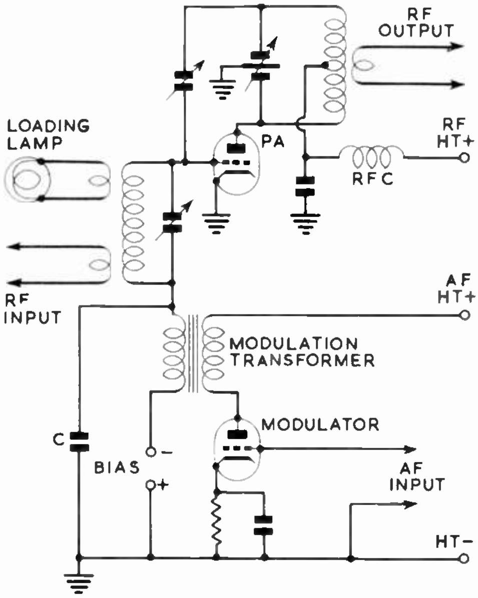


Fig. 9.11. Grid modulation. The lamp is used to absorb an appreciable proportion of the grid-driving power and thus improve the grid drive regulation. The capacitor C serves as an r.f. bypass: it must be small enough to present a high reactance at the highest modulation frequency.

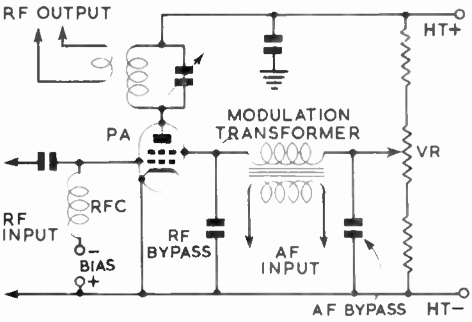


Fig. 9.12. Screen modulation. The screen supply voltage must be obtained from a potentiometer VR to enable the necessary critical adjustment to be made for maximum depth of modulation.

The chief advantage of grid modulation is the relatively low audio power required, a few watts generally sufficing to modulate a 150 watt stage.

The power input to the p.a. valve must be limited to such a value that the maximum permissible anode dissipation is not exceeded, assuming the low average efficiency of 35 per cent. The adjustment generally consists of setting the grid drive (which is about 20–25 per cent of that necessary when maximum efficiency as a c.w. output stage is required), and adjusting the grid bias and the aerial loading until the appropriate amount of aerial current increase is obtained without causing any variation of the p.a. anode current. The audio frequency gain is then adjusted so that the desired peak modulation is obtained: in this system the peak modulation will usually have to be limited to about 90 per cent to prevent the occurrence of distortion.

Screen-grid Modulation

The general circuit arrangement used for screen-grid modulation is shown in **Fig. 9.12.** Popular types of beam tetrodes, such as the 807 and the 813 are well suited to this system.

The main drawback of screen modulation is that the screen-grid voltage must be reduced to a fairly low value (e.g. 100–130 volts in the case of the 807). Unless this is done, the upward excursions of the screen voltage during modulation would be limited by the sharp rise in screen current. The positive peaks of the modulated r.f. output waveform would thus be flattened and the modulation would become asymmetrical.

The low average screen voltage results in a rather low r.f. carrier output power, as in the case of control-grid modulation, and it may therefore be worthwhile to incorporate two r.f. valves in push-pull when using this system. Screen-grid modulation does not call for critical adjustment of the grid drive to the modulated stage but the unmodulated screen voltage must be very carefully adjusted to allow the greatest possible depth of modulation without distortion. This maximum is generally of the order of 75–80 per cent. The audio power required, as in other efficiency modulation systems is quite low, 4–6 watts being sufficient to modulate an input of 150 watts.

Clamp Modulation

The so-called clamp-modulation system is an interesting variation of screen-grid modulation. The purpose of a "clamp" valve is to reduce the screen voltage of a tetrode r.f. amplifier to a low value in the absence of grid excitation, so that it acts as a protective device and allows the exciter to be keyed for c.w. operation without the necessity for using fixed bias on the p.a. stage. A typical circuit is shown in **Fig. 9.13.** The clamp valve V_2 which may be of the KT66 or 6L6 class, is cut off by the operating bias of the r.f. stage V_1 which is developed across R_1. When the excitation is removed, there is no bias on V_2 and hence it passes a large anode current which develops a high voltage-drop across R_2, the screen resistor of V_1. This reduction in the screen voltage of V_1 reduces the anode current to a very low and safe value.

For the purposes of modulation the grid of the clamp valve may be switched from the r.f. bias resistor R_1 to the output of a speech amplifier V_3 so that the screen of V_1 receives the audio output voltage through V_2.

When V_2 is switched for telephony operation its bias should be arranged to be such that its anode voltage, i.e. the screen voltage of V_1, is about one half of the value which it has in the c.w. position. Tuning-up can take place

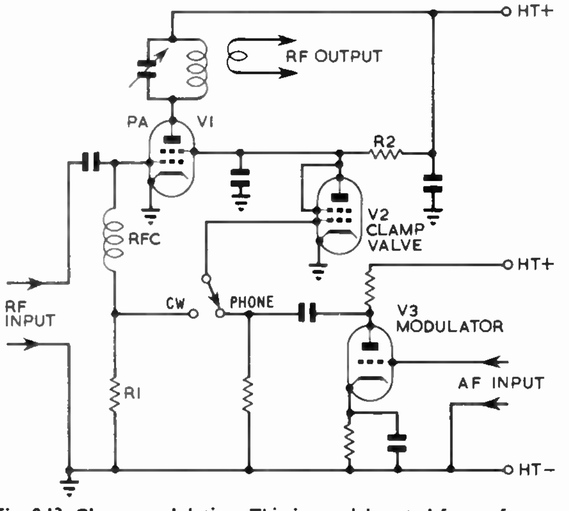


Fig. 9.13. Clamp modulation. This is an elaborated form of screen modulation which enables the transmitter to be easily switched to c.w. operation.

in the c.w. position, but when V_2 is switched to the telephony position the carrier output of V_1 will of course fall. When modulating, the audio gain should be such that the anode current of V_1 on modulation peaks does not exceed about three-quarters of the normal c.w. value.

Series Gate Modulation

Series gate modulation is a form of screen modulation of fairly recent origin. It is simple and effective and appears to justify the claims made for it. The name "series gate" modulation was given to it by the originator, but it does not indicate clearly the mode of operation.

The basic series gate modulator circuit is shown in **Fig. 9.14.** This employs only one valve, a twin triode, but there is, of course, no reason why two separate triodes should not

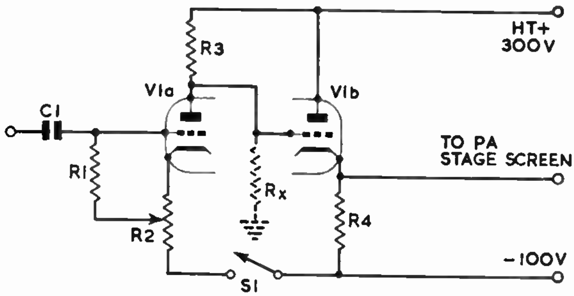


Fig. 9.14. Basic "series gate" modulation circuit.

be used. The circuit is essentially a voltage amplifier V_{1a} direct-coupled to a cathode-follower V_{1b}. Microphone input is applied via C1.

When the slider of R_2 is at the cathode end of its travel, V_{1a} is operating without bias; there is, therefore, a large voltage drop across R_3 and the anode voltage of V_{1a} is reduced to a low value, of the order of 15 volts relative to the chassis. This voltage is direct-coupled to the cathode-follower V_{1b}, and so about 95 per cent of this voltage or 14 volts appears across the cathode load R_4. Thus the screen voltage of the r.f. amplifier is also 14 volts and the r.f. output is low. If the slider of R_2 is moved away from the cathode end, the bias on V_{1a} increases and the voltage drop across R_3 is reduced. The anode voltage of V_{1a}, and hence the grid voltage of V_{1b}, then rises to the order of 250 volts. The screen voltage of the p.a. stage will therefore be increased to somewhat less than this as a result of the cathode-follower action of V_{1b}. Thus the output of the p.a. stage will be increased towards its maximum.

The setting of R_2 therefore makes it possible to vary the output of the r.f. stage from a low level to almost full c.w. ratings. If S_1 is opened, V_{1a} is inoperative and the anode of V_{1a} and the grid of V_{1b} rise towards the anode supply voltage and the screen voltage of the p.a. stage is limited by the flow of grid current of V_{1b} through R_3. The screen potential is then the highest obtainable with this particular circuit arrangement. The maximum screen voltage may be set by the introduction of R_x. R_3 and R_x then form a potentiometer across the h.t. supply which holds the grid voltage of V_{1b} constant.

Modulation may be achieved by applying an a.f. voltage to C_1. Assuming that the bias on the grid of V_{1a} is set (by R_2) to −1V, the d.c. voltage applied to the screen of the p.a. is low and so the r.f. output is low. If an a.f. voltage of 1V peak is now applied to the grid of V_{1a}, it will be amplified by V_{1a} and appear at the p.a. screen and so modulate the low r.f. output to a depth of approximately 95 per cent, the mean d.c. potential of the p.a. screen remaining constant. If the a.f. voltage applied to V_{1a} is increased, grid current will flow in V_{1a} and a negative charge will build up on C_1 proportional to the peak value of the a.f. voltage. This additional d.c. bias applied to the grid of V_{1a} will cause the potential on its anode and the grid of V_{1b} (and of course the mean screen voltage of the p.a. stage) to rise, resulting in increased r.f. output from the p.a. stage. Since the period of time for which this potential is raised depends upon the time it takes for C_1 to discharge through R_1, the time constant of R_1 and C_1 is important and must be considerably longer than the time corresponding to the lowest audio frequency used.

The increased a.f. voltage at the grid of V_{1a} will appear as an amplified voltage at the p.a. screen relative to its previous value, but as the mean screen potential has also been raised the carrier is again modulated to a level of about 95 per cent.

The cathode follower is connected across the h.t. supply and a negative supply of the order of 75 volts to 100 volts. This is necessary to set the operating conditions of the p.a. stage and to provide the additional voltage across the cathode follower so that the p.a. screen voltage can rise to the maximum value required allowing for the voltage drop

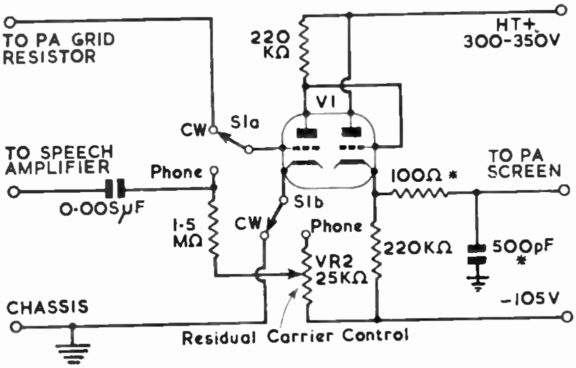


Fig. 9.15. Practical series gate modulator. For p.a. inputs up to 75 watts, suitable valves are the 6SN7 and 12AU7. For higher inputs, a 12BH7 may be used. Alternatively, two 6J5s or triode connected 6V6s are suggested. The 100 ohm resistor and 500 pF capacitor marked with asterisks are the normal screen decoupling components mounted close to the p.a. valveholder.

across the impedance of the valve used as the cathode follower. The negative supply can be the p.a. bias supply.

A practical series gate modulator is shown in **Fig. 9.15.** It will be seen that this is virtually the same as the basic circuit shown in Fig. 9.14 the only additions being the PHONE/C.W. switch S_1, the 500 pF capacitor and the 100 ohm resistor. The latter are the normal p.a. screen decoupling components which of course should be mounted as close as possible to the p.a. valveholder. For inputs to the p.a. of up to 75 watts, recommended valves are the 6SN7 or 12AU7. For higher inputs, a 12BH7 is to be preferred. Alternatively separate valves such as the 6J5 or triode connected 6V6's (or similar) may be used.

The method of adjustment is as follows. The transmitter should be tuned and loaded in the normal manner with S_1 set to c.w. Switch to PHONE and adjust VR_2 i.e. the screen voltage of the p.a., to give an input or " residual carrier " of about one-fifth to one-eighth of the full carrier input under c.w. conditions. The audio gain control in the speech amplifier should then be slowly advanced while speaking into the microphone until the anode current rises on peaks to the value obtained under c.w. conditions. If too small a value of residual carrier is used, the signal tends to become difficult to tune-in initially. When receiving a series gate modulated signal it is preferable to switch off the receiver a.g.c.

The advantages of series gate modulation may be summarized as follows:

(a) Overmodulation on positive peaks cannot occur because as the a.f. voltage applied to C_1 (Fig. 9.14. is increased, limiting will occur in V_{1a} and hence the a.f. voltage applied to the p.a. screen cannot increase.

(b) Splatter caused by the breaking up of the carrier by overmodulation on the negative peaks cannot occur because the screen voltage cannot fall below the value selected by the residual carrier control VR_2. Thus a simple but effective means of speech clipping is available merely by slightly advancing the audio gain control beyond the point specified above. Too much clipping in this manner will, as is usual, tend to decrease intelligibility rather than increase it.

(c) In the cw position of S_1 the modulator functions as a normal clamp circuit and so protects the p.a. under key-up conditions.

The disadvantage is that a negative bias supply is required. This is not necessarily available with the widespread use of a clamp valve on the p.a. stage. A battery or a well-smoothed supply is necessary.

Thus even a 150 watt input r.f. power amplifier may be modulated by a small unit using two twin triodes. The small size and power requirements of such a modulator make it ideal for mobile or portable equipment. It is also a convenient way of modifying a c.w. transmitter for telephony operation with the minimum of expense and addition.

Suppressor-grid Modulation

The gain of a pentode may be controlled independently of the other voltages by varying the voltage applied to the suppressor-grid. This forms the basis of the suppressor-grid modulation system in which the a.f. modulating voltage is added to the suppressor-grid bias, as shown in Fig. 9.16. This form of modulation is one of the simplest to adjust, and is capable of producing up to about 95 per cent modulation with very little distortion, and requires only a small amount of a.f. power. Adjustment is carried out by changing the grid bias in a negative direction by an amount which reduces

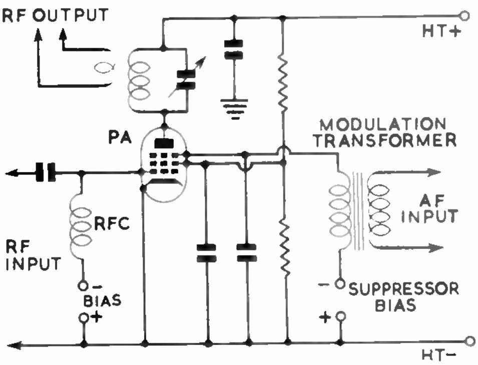


Fig. 9.16. Suppressor-grid modulation. The bias on the suppressor grid needs to be carefully adjusted.

the anode current to one half of that obtained for c.w. working. The peak value of the a.f. voltage applied to the suppressor will need to be equal to this change of suppressor-grid bias, and under these conditions practically full modulation will be obtained.

Grid-leak Modulation

A form of grid modulation which is occasionally used for low-power operation is called *grid-leak modulation.* If, in an r.f. amplifier which obtains its operating bias solely by the grid-leak method, the value of the grid leak is varied, the grid bias and hence the output will also be varied. In practice this may be accomplished by using a valve as the grid leak (**Fig. 9.17**). The usual adjustments should be made to grid drive and aerial loading for approximately one half of the normal c.w. output. Under these conditions the aerial current should rise during modulation by the normal amount.

The power efficiency of this system is very low, being about 25–30 per cent, but on the other hand it is probably the most economical system since it can be made to function successfully with only a carbon microphone and its transformer and one small triode.

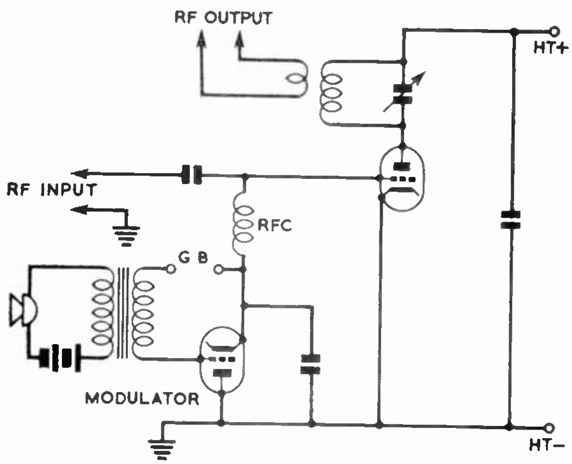


Fig. 9.17. Grid-leak modulation. This system requires the minimum amount of audio equipment but the r.f. efficiency is low.

Cathode Modulation

Modulation may alternatively be applied to the cathode circuit of the p.a. as illustrated in **Fig. 9.18.** This is known as *cathode modulation.* In effect both anode and grid voltages are varied simultaneously with respect to the cathode. Cathode modulation is therefore a compromise between anode modulation at high output efficiency and grid modulation at low output efficiency. By suitable adjustment, any desired proportion of each form may be achieved.

The ratio of the respective amounts of anode modulation and grid modulation is controlled by the value of the grid bias on the p.a., which in turn will determine the amount of audio-frequency power required. The latter determines the amount of anode modulation, while the former determines the amount of control that a given value of modulating voltage will have. The grid bias, of course, will depend upon the value of the r.f. and a.f. voltages. A greater range of adjustment in the grid circuit may be obtained if the grid return can be taken to a variable tap on the secondary winding of the modulation transformer; for this reason a multi-ratio type of transformer is preferable.

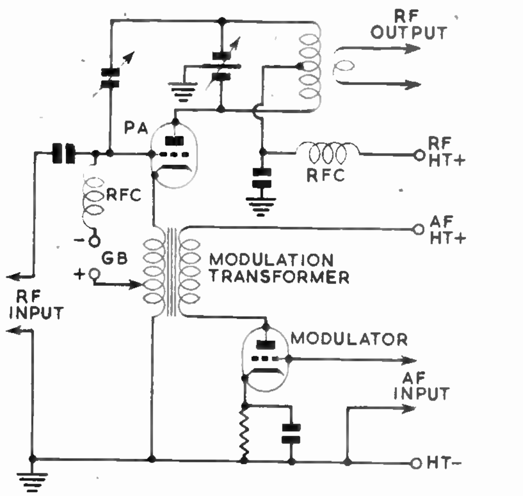


Fig. 9.18. Cathode modulation. This method is a combination of (a) anode modulation with its high output efficiency and high-modulation-power requirement and (b) grid modulation with its low output efficiency and low modulation-power requirement.

The setting-up of such a circuit can be fairly complicated because in addition to the normal adjustments, the proportion of grid and of anode modulation should be adjusted to obtain full modulation. However, once set up the system is capable of very good results.

Low-level Modulation

The modulation systems so far considered are known as "high-level" systems because the modulation is applied to the r.f. stage working at the highest power level. There is no reason why the modulation should not be applied to a low-power stage, the modulated output of which is then amplified. This "low-level" modulation system is used in s.s.b. transmitters and is occasionally found in other amateur equipment. The modulated r.f. voltage can be amplified to increase the output power, but in order to avoid distortion of the modulated wave the r.f. amplifiers following the modulated stage must be of the linear type.

The reader is referred to Chapter 10 (*Single Sideband*) for details of the design and operation of linear amplifiers.

Choice of Modulation System

There is little doubt that anode modulation is the most effective system of amplitude modulation from many points of view. It has, however, the disadvantage that the audio frequency power requirements are the highest, and thus the modulator and associated power supply are expensive and often complicated pieces of equipment.

It may be argued that the overall power efficiency of a transmitter employing one of the so-called efficiency-modulation systems is much higher on account of the smaller amount of audio power required. This may well be a powerful argument in the case of portable, mobile or battery-operated equipment, but such systems are not often adopted for fixed-station work. There is not a great deal of difference between any of the variable-efficiency methods. They are all capable of excellent results, but in general they are harder to adjust correctly than the anode-modulation system. The published valve ratings usually refer to maximum sinusoidal conditions, but owing to the "peaky" nature of the speech waveform the average conditions are appreciably less arduous, and with very great care in adjustment and operation these circuits can be safely operated at somewhat higher ratings than those usually published for a given valve. Perhaps the greatest advantage of variable-efficiency systems is that they present a convenient method of telephony operation to the amateur whose main interest lies in c.w operation.

MICROPHONES

The microphone is the starting point of any electrical system of speech communication, as it is the device which converts the energy of the sound wave of the human voice into electrical energy. The output voltage or current can then be amplified and used to modulate a radio transmitter.

Basic Acoustical Principles

A sound wave is a wave of alternating pressure and rarefaction spreading out from the source, in which the instantaneous air pressure at any fixed point varies above and below the mean atmospheric pressure by an amount which determines the intensity of the sound. The waveform of this pressure variation is a curve whose shape determines the characteristics by which one sound differs from another. For example, a pure sine wave of pressure results in a single tone described as "pure," the frequency determining the pitch. Different musical instruments all playing a note of the same pitch produce waves of varying harmonic content, the proportions of which determine the tonal quality or *timbre* of the sound. Harmonics (or overtones) similarly determine the characteristic differences between the male and female singing voices. The important parts of the waveshape of speech sound are the harmonics of the fundamental pitch and the transient wave-shapes.

The pressure of a sound wave represents the variation about the mean pressure, and is usually measured in dynes per square centimetre. When referring to sinusoidal waves either the peak value or the r.m.s. value may be used as in other alternating quantities, the r.m.s. figure being more generally quoted.

In an electrical circuit an e.m.f. or electrical pressure produces an electrical current, the magnitude of which depends upon the electrical resistance. In an analogous manner an acoustical pressure-wave is accompanied by a movement of the air particles, which may conveniently be measured by the velocity of these particles. This velocity depends upon the *acoustic impedance* of the air space upon which the pressure is acting. This impedance depends in turn on the distributed mass and elasticity of the air, and can be regarded as a "characteristic impedance" similar to the characteristic impedance of a cable or transmission line which is determined by the distributed inductance and capacitance.

The wavelength of an acoustic wave is derived in the same way as that for a radio wave, except that the wave velocity in air is approximately 344 metres/second compared with 300,000,000 metres/second for electromagnetic waves. This wave velocity is a constant figure for any given gas or liquid at a specified temperature and must not be confused with the particle velocity referred to above, which depends upon the acoustic pressure.

Thus, the wavelength corresponding to a frequency of 100 c/s is given by:

$$\lambda = \frac{34{,}400}{100} = 344 \text{ cm.}$$

and that corresponding to a frequency of 10,000 c/s by:

$$\lambda = \frac{34{,}400}{10{,}000} = 3{\cdot}44 \text{ cm.}$$

Microphones can be designed so that their electrical output voltage is proportional either to the acoustic pressure or to the particle velocity, thus giving rise to the terms *pressure microphone* and *velocity microphone*.

The pressure microphone, provided it is small enough to avoid complications due to reflection of the wave, will give a constant output no matter in what direction it is facing relative to the direction of travel of the wave. In practice, however, when the frequency is high enough for a half-wave to approximate to the size of the microphone, or higher, the microphone will be more responsive to sounds from a direction facing the microphone than to those from the sides or the rear. Thus, at 10,000 c/s where the half-wavelength is 1·72 cm. the "front-to-back" ratio would be quite high for a microphone 3 in. in diameter, whereas at 100 c/s, corresponding to a half-wavelength of 172 cm., the same microphone would give the same output regardless of the direction of the sound.

The term *velocity* has no meaning unless it is associated with a direction as well as a magnitude, since sound waves consist of longitudinal movements of the air particles in the direction in which the wave is travelling. The velocity at any point is at a maximum in a direction facing the source, and zero in a direction at right angles to this. Thus a velocity microphone, i.e. one in which the output is designed to be proportional to the particle velocity, will have maximum output when facing towards or away from the source and zero output when at right angles to the source; moreover, the output is independent of any reflection effects referred to earlier, i.e. it is zero in the perpendicular direction, even at the lowest frequencies.

An ideal pressure microphone has a polar diagram consisting of a circle, while a velocity microphone has a diagram like a figure-of-eight, similar to that of a dipole aerial. These diagrams apply in any plane, which means that for a pressure microphone the solid polar diagram is a sphere, while that for a velocity microphone is two spheres touching at the point represented by the microphone.

Basic Electro-mechanical Principles

All practical microphones consist of a mechanical system in which some part moves in sympathy with either the acoustic pressure or the particle velocity, or a combination of both. The pressure microphone comprises a closed chamber containing air at normal atmospheric pressure, part of this chamber being made in the form of a diaphragm which, if it is small and light, will move in sympathy with the variations in air pressure produced by the sound wave outside the chamber above and below the mean pressure of the air inside the chamber.

In a velocity microphone, however, a very light membrane with both sides open to the air is used. When the plane of the membrane is at right angles to the direction of the air movements, it will move with the air, and its velocity will be almost the same as that of the air particles if it is light enough.

To convert a movement of the diaphragm of the pressure microphone or of the membrane of the velocity microphone into an electrical output either the resistance of a circuit may be varied, as for instance by the use of carbon granules, or electromotive forces may be produced by electro-magnetic induction or the piezoelectric effect.

Microphone Characteristics

Brief details are given below of the construction and characteristics of the more common types of microphone. Reference should be made to the manufacturers' literature for more detailed information.

The Carbon Microphone. In this microphone a small capsule filled with carbon granules is attached to a diaphragm, and the effect of the varying pressure of the sound waves is to cause a similar variation in the resistance of the capsule. When connected to a source of constant voltage the microphone thus produces a varying current. The normal resistance of the capsule is generally between 200 and 1000 ohms and the polarizing current should be in the range 5–40 mA according to the particular design of capsule.

The source of current may be a small dry battery or a potential divider connected across the h.t. supply to the speech amplifier, as shown at (A) and (B) in **Fig. 9.19.** To match the low impedance of the microphone circuit to the high impedance of the input grid circuit of the amplifier, a transformer with a large turns-ratio is required; any ratio between 1 : 30 and 1 : 100 should prove satisfactory.

Although these microphones are very sensitive, the frequency response is poor compared with that of most other types. An output of between 5 and 15 volts can be expected across the secondary of the transformer when speaking close to the microphone, depending on the transformer ratio and the polarizing current. To avoid distortion, the latter should be kept as low as possible consistent with the output required. It will be appreciated that a microphone which utilizes the change in pressure between a multitude of contact points cannot be as free from distortion as some of the more elaborate types described below.

The carbon microphone is particularly convenient for portable use as its high output is capable of fully driving valves such as the KT66 and 6L6 directly from the microphone transformer. An alternative method of using a carbon microphone is shown at (C) in Fig. 9.19. Here the microphone acts as the cathode bias resistance of the first valve, the grid being earthed. This form of connection avoids the use of a transformer and a separate polarizing supply, and is capable of somewhat better quality, although the output may be lower.

The Transverse-current Microphone. This is another form of resistance microphone employing carbon granules. It has a mica diaphragm, usually rectangular in shape, behind which is a shallow chamber containing very fine carbon granules. The polarizing current passes between carbon electrodes which are placed at opposite edges of the chamber so that the granules form a thin layer of carbon through which the current passes from one edge to the other. As with the simple carbon microphone a polarizing current and step-up transformer are required. Because the diaphragm is heavily damped all over its surface by the carbon granules, the frequency response is much better than that of the capsule type, although the sensitivity is lower. An output in the order of 0·1–0·5 volt across the transformer secondary can generally be obtained.

The Moving-coil Microphone This microphone is an electrodynamic type. Its construction is very similar to that of a moving-coil loudspeaker, except that the whole construction is very much smaller. The impedance of the coil is generally 20–50 ohms. When the diaphragm is caused to vibrate by the sound pressure an e.m.f. is developed in the coil and no polarizing supply is necessary. A step-up transformer, however, is required to feed the grid of the first amplifier valve. A very good frequency response can be obtained, and this type of microphone is commonly used in broadcasting studios. The output from the secondary of a suitable transformer is usually not more than about 0·5 volt. A miniature moving-coil loudspeaker can sometimes be made to serve as a fairly satisfactory microphone, but its frequency response will not be as good as that of a specially designed instrument.

The Ribbon Microphone. This type of microphone has an output proportional to the velocity of the air particles. The main difference between its operation and that of a pressure microphone is that it has a figure-of-eight polar diagram; it is thus necessary to align its axis with that of the direction of the sound source.

The most common form consists of a very thin strip of aluminium foil (about 1 in. long, 0·1 in. wide and 0·0003 in. thick) supported between the poles of a permanent magnet. As the ribbon vibrates it cuts the magnetic field

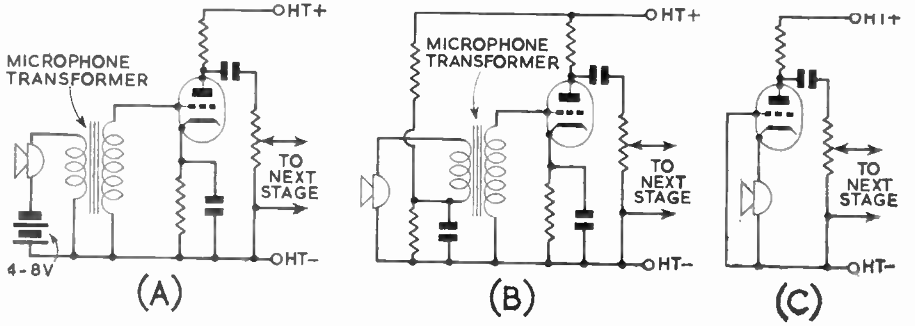


Fig. 9.19. Typical carbon microphone input circuits. In (A) the polarizing current is obtained from a small battery: in (B) a potential divider across the h.t. supply is used as a substitute for the battery: in (C) the microphone is used to modulate the cathode current directly by acting as a variable cathode - bias resistance.

between the poles, and so an e.m.f. is induced in the strip. A step-up transformer is normally placed within the microphone case to transform the very low ribbon impedance to 50–250 ohms to make the microphone impedance suitable for working into a cable, and a further step-up transformer is used at the amplifier input. No polarizing current is required, and the frequency response can be very good, although the sensitivity tends to be rather low.

The Condenser Microphone. The condenser microphone consists of a thin conducting diaphragm which forms one plate of a capacitor whose capacitance varies when actuated

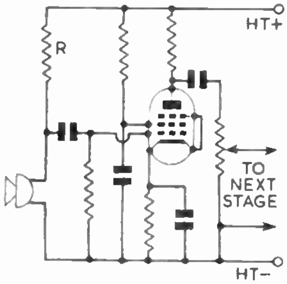
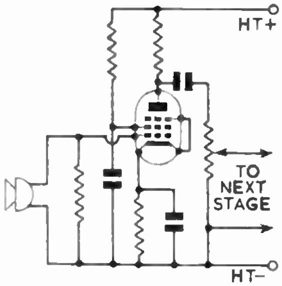


Fig. 9.20. Condenser microphone input circuit. The resistance R must be very high and is usually of the order of several megohms.

by sound waves. A polarizing voltage of 100–200 volts is required in series with a very high resistance across which the output voltage is developed. The usual circuit arrangement is shown in **Fig. 9.20** where R is the resistor across which the output voltage appears. Because the capacitance of any connecting cable will reduce the output, it is necessary for the first stage of the amplifier to be close to the microphone. Consequently the first stage of the amplifier is often built into the microphone housing. Owing to this complication and to the fact that the normal output is generally very low, condenser microphones are not commonly used in amateur stations.

The Crystal Microphone. This type of microphone utilizes the piezoelectric effect whereby any mechanical stress applied to a suitably cut piece of certain materials such as quartz or Rochelle salt generates an e.m.f. between opposite faces. Such microphones are very popular amongst

Fig. 9.21. Crystal microphone input circuit.



amateurs since their frequency response is usually very good, and neither a polarizing voltage nor a transformer is required. They can be connected directly to the grid of the first amplifier stage through a moderate length of screened cable (up to about 8 ft.). A high resistance must, however, be connected across the output at the grid of the valve, as shown in **Fig. 9.21.** This should be 2–5 Megohms; any value lower than 2 Megohms will impair the output at low frequencies.

There are two types of crystal microphone in general use. The most common consists of a diaphragm which acts directly on the crystal (normally Rochelle salt). In the second type, known as the " cell " type, the diaphragm is dispensed with, and the sound waves act directly on a pair of crystals which form the surface of a flat cell. Both types are capable of excellent quality, the latter type having a better frequency response but being less sensitive.

Relative Sensitivities

In commercial practice the output of a microphone is generally specified in decibels relative to a reference level of one volt per dyne per square centimetre of acoustic pressure. For amateur purposes, however, it is more convenient to be able to estimate the actual voltage output which is likely to be obtained from the various types of microphone described above when the microphone is held in the hand of the speaker and is actuated by normal speech. An approximate indication of typical output voltages is given in Table 9.2.

SPEECH AMPLIFIERS

The term *speech amplifier* is generally applied to the stage or stages of an a.f. amplifier whose function is to amplify the low output voltage of a microphone to a level suitable for driving the modulator valves which form the last stage in the audio frequency chain. The speech amplifier may consist entirely of *voltage amplifiers*, i.e. amplifiers which increase the signal voltage without having to supply power into the load, or in some cases the final stage of the speech amplifier may be a *power amplifier*. This is only necessary when the modulator valves are driven so hard that grid current flows.

TABLE 9.2

Voltage Output of Various Microphones

Type	Transformer Ratio	Voltage at Grid
Carbon (capsule)	30–100	5 –15 volts
Transverse-current	20– 40	0·1–0·3 volts
Condenser	—	0·05 volts
Ribbon	10	0·03 volts
Moving-coil	30– 50	0·1–0·5 volts
Crystal (diaphragm)	—	0·05 volts
Crystal (cell)	—	0·01 volts

Voltage Amplifiers

Audio-frequency voltage amplifiers may be divided into two categories, viz. resistance capacitance (RC) coupled amplifiers, and transformer-coupled amplifiers. The basic circuit of an RC-coupled amplifier is shown in **Fig. 9.22.** The voltage amplification of such a stage is given by the expression:

$$\frac{V_{out}}{V_{in}} = \mu \left(\frac{R_L}{R_L + r_a} \right)$$

where μ = amplification factor of V_1
r_a = anode resistance of V_1
R_L = external anode load impedance

It is assumed that the value of R_g, the following grid resistor, is greater than $R_L r_a/(R_L + r_a)$, a condition which is generally fulfilled.

It will be seen from this relationship that in order to achieve high gain, R_L must be large compared with r_a. If, however, R_L is made too large, the d.c. voltage-drop across it will be excessive and a high d.c. voltage will be necessary in order to maintain adequate voltage at the anode of V_1. In practice, R_L is generally made 3–5 times as large as r_a. Grid bias for V_1 is provided by the voltage-drop across the cathode resistor R_1, which is by-passed by C_1, to prevent negative current feedback with resulting loss of gain.

When V_1 is a pentode R_L cannot satisfactorily be made greater than r_a because the latter is normally of the order of several Megohms, and values of R_L between 100,000 and 330,000 ohms are normally used with pentode amplifiers.

The frequency response of such a stage (i.e. the variation of gain over the frequency range) is determined principally by the values of three capacitances:

(a) The cathode bypass capacitor C_1.
(b) The coupling capacitor C_c.
(c) The total input capacitance C_s of the following stage V_2.

The first two affect the response at low frequencies and have negligible effect at high frequencies, while the third affects the high-frequency response.

The capacitance of the cathode bypass capacitor C_1 should be such that its reactance at the lowest frequency at which amplification is desired should be small compared with the value of the cathode bias resistor. This capacitor also serves a useful purpose in that it bypasses a.c. hum voltages between the cathode and the heater; hence it is advantageous to use a somewhat larger capacitor than that dictated by low-frequency response considerations. For most practical purposes a 25 μF capacitor is adequate.

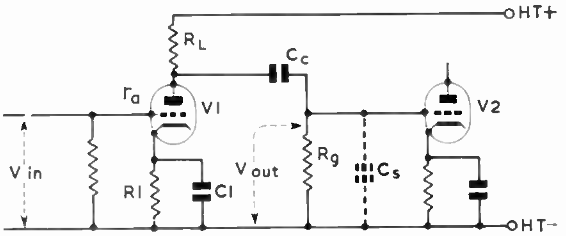


Fig. 9.22. Basic circuit of a resistance-capacity voltage amplifier.

At low frequencies, the reactance of the coupling capacitor C_c increases, and since C_c and R_g form a potential divider the voltage appearing across R_g, which is the input to the following stage, tends to fall. The relative gain at a low audio frequency, with respect to that at frequencies in the middle of the audio range, may be found from the expression:

$$\frac{R_g}{\sqrt{R_g^2 + \dfrac{1}{\omega^2 C_c^2}}} \times 100\%$$

where R_g and C_c are the values of the grid leak and the coupling capacitor respectively of the following stage and $\omega = 2\pi f$. In practice this means that the value of the coupling capacitor must be about 0·5 μF for good low-frequency response. However, for speech, where good response below 300 c/s is not essential, a value of 0·01 μF is adequate for use with a grid leak R_g of 0·1 Megohm or more.

At high frequencies the reactance of C_s falls, and as this capacitance is, in effect, in parallel with the output of the previous stage, appreciable high-frequency loss can occur if its value is too great. The capacitance C_s is composed of the input capacitance of the valve V_2 and the various stray capacitances of the circuit components.

The input capacitance of an amplifier valve with a resistance load is:

$$C_{input} = C_{gc} + (M + 1)\, C_{ga}$$

where C_{gc} = grid/cathode capacitance of the valve
C_{ga} = grid/anode capacitance of the valve
M = stage gain

This reflection of the grid-to-anode capacitance into the grid circuit is known as the *Miller Effect*, and in some circumstances it may make the input capacitance quite high. The relative amplification at high audio frequencies compared with that in the middle of the audio range may be calculated from the expression:

$$\sqrt{\frac{1}{1 + (\omega R C_s)^2}} \times 100\%$$

where R = equivalent resistance of R_L, r_a and R_g in parallel
C_s = total shunt capacitance

This high-frequency loss can be very important in the design of high-fidelity amplifiers, but for speech communication where an upper frequency limit of about 3000 c/s is acceptable it is not usually of great significance.

Resistance-capacitance coupling is a cheap and convenient form of inter-stage coupling and is in almost universal use. It has the added advantage that it is not prone to pick up hum from stray magnetic fields, which can sometimes

Table 9.3

Valve	HT (V)	Rg (Meg-ohms)	Rl (K ohms)	Cc (μF)	V out (V)	Gain
½·12AU7	180	0·1	2·0	·032	24	12
	180	0·22	2·8	·016	33	12
	180	0·47	3·6	·007	40	12
	300	0·1	1·9	·032	44	12
	300	0·22	3·0	·016	68	12
	300	0·47	4·0	·007	80	12
½·12AT7	180	0·22	2·6	·014	18	29
	180	0·47	2·6	·009	19	31
	180	1·00	2·7	·006	20	28
	300	0·22	1·2	·015	22	34
	300	0·47	1·2	·009	23	36
	300	1·00	1·25	·006	24	38
½·12AX7	180	0·22	3·0	·012	24	53
	180	0·47	3·5	·006	34	59
	180	1·00	3·9	·003	39	63
	300	0·22	2·2	·013	54	59
	300	0·47	2·8	·006	60	65
	300	1·00	3·1	·003	79	68

Symbols as in Fig. 9.22 (R_L assumed to be 100 K ohms)

Table 9.4

Valve	HT (V)	RL (K ohms)	R2 (Meg-ohms)	RI (K ohms)	Rg (K ohms)	V out (V)	Gain
EF86/Z729	200	100	0·39	1·0	330	40	106
	300	100	0·39	1·0	330	64	116
	200	220	1·0	2·2	690	36	176
	300	220	1·0	2·2	680	54	188

R2 is value of screen-dropper resistor, assumed to be bypassed to earth by 0·1 μF capacitor. Otherwise symbols are as in Fig. 9.22.

be troublesome with transformer-coupled circuits. It may be used with either triodes or pentodes, and in fact it is the only suitable form of coupling with high-μ valves because of the difficulty of making a transformer suitable for operating with the high load impedances associated with such valves.

Operating conditions for typical RC-coupled amplifiers are given in **Table 9.3** (triodes) and **Table 9.4** (a.f. pentode).

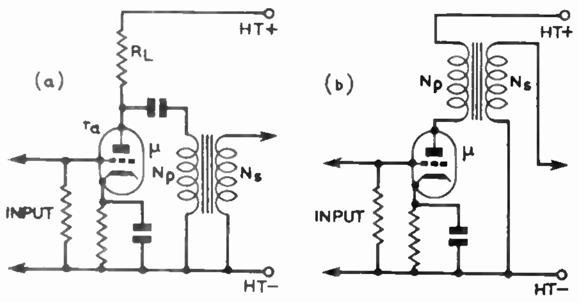


Fig. 9.23. Alternative arrangements of transformer-coupled amplifiers.

Transformer-coupled Amplifiers

Transformer-coupled amplifiers are ordinarily used only when it is desired to amplify power rather than voltage, or when a push-pull output is required from a single-ended input. For most effective operation low-μ triodes operating in class A are generally used because the low anode loads required make the design of the transformer easier.

Two alternative arrangements are shown in **Fig. 9.23**. In (a) a parallel feed to the transformer is shown, in which there is no d.c. flowing through the primary and thus there is no reduction in the impedance of the primary winding. The overall voltage gain of this arrangement is:

$$Gain = \mu\left(\frac{R_L}{R_L + r_a}\right)\left(\frac{N_s}{N_p}\right)$$

In the arrangement shown at (b) in Fig. 9.23 the transformer carries the d.c. anode current of the valve, and to prevent undue reduction of the primary inductance due to this effect the transformer may need a core with a suitable air gap. The voltage amplification, assuming no grid-current load on the secondary, is given by:

$$Gain = \mu\left(\frac{N_s}{N_p}\right)$$

The frequency response of such stages is governed mainly by the characteristics of the transformer. At very low frequencies, the inductive reactance of the primary winding falls, and as this forms either part of or the whole of the anode load of the valve, the amplification will be correspondingly reduced. Thus, a high primary inductance is necessary for good-response at low frequencies. At the upper end of the audio-frequency range the self-capacitance and leakage inductance of the transformer become important, and may cause a resonant rise in the amplification. To maintain a uniform high-frequency response such transformers are generally constructed with sectionalized windings to reduce the leakage inductance and the self-capacitance.

Multi-stage Amplifiers

Several amplifiers may be operated in cascade, in which case the overall voltage amplification is then the product of the individual voltage amplifications of each stage. It is necessary to proportion the amplification of individual stages to avoid overloading. In high-gain amplifiers it is essential to prevent stray coupling between the stages, both by correct layout of components to avoid unwanted capacitive couplings, and also by adequately decoupling the h.t. supply to each stage. A resistance of 10,000–33,000 ohms, and a capacitor of 4–8 μF is normally sufficient.

Gain controls should be introduced as near to the input as possible in the amplifier chain to prevent overloading of the early stages by strong signals. This may often mean inserting the control in the grid circuit of the first stage of the amplifier, in which case the circuit should be adequately screened to prevent the pick-up of hum voltages. In such circumstances it may be preferable to insert the gain control in the second stage.

Push-pull Amplifiers

A push-pull stage requires the two grids of its two valves to be driven by identical voltages of opposite polarity.

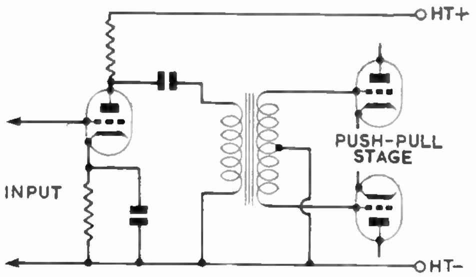


Fig. 9.24. By using a transformer which has a centre-tapped secondary winding, a balanced output suitable for driving a push-pull stage is obtained.

In class A stages no grid current will be produced by this voltage, but in class B stages grid current will be caused to flow, and the driver stage must be able to supply power for this purpose. It should have a low output impedance so that no waveform-distortion of the voltage is caused by the

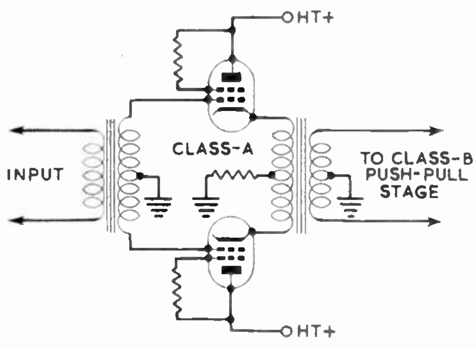


Fig. 9.25. Push-pull cathode-follower driver stage. The cathode output circuit has a low impedance and is suitable for working into a step-up transformer for driving a class B push-pull output stage.

variation of the grid current over the audio cycle. The various methods of obtaining the push-pull driving voltages are discussed below.

Centre-tapped Transformers. The transformer-coupled amplifier using a transformer with a centre-tapped secondary, as shown in **Fig. 9.24**, is the easiest way of obtaining a push-pull output. It can be used for supplying a push-pull grid circuit from a single-ended stage, a step-down ratio often being employed to obtain the required low impedance in the drive circuit when the push-pull stage is of the class B type. **Fig. 9.25** shows another arrangement in which the input transformer is a step-up transformer driving the push-pull stage in class A, while the following push-pull stage (not shown) is driven from the cathodes of the first stage through a transformer having centre-tapped primary and secondary windings. This utilizes the property of the low-impedance output of the cathode follower circuit, and hence it is not necessary for this transformer to have a step-down ratio for driving a class B push-pull output stage. It should be noted that the gain of cathode follower driver stage is only about 0·95.

A transformer with a centre-tapped primary is used almost universally to couple a push-pull power amplifier to its load.

Cathode-coupled driver stages can be connected to the grid circuit of the output stage directly, without the use of push-pull transformers, in the manner shown in **Fig. 9.26.** If the push-pull input to the driver stage is obtained from one of the circuits which follow, no transformers other than the modulation transformer are required.

Phase-splitting Circuits. There are a number of circuits capable of producing balanced voltages of opposite polarity (where grid current is not involved) for driving grid circuits in push-pull without the use of transformers. They all utilize modified forms of *RC* coupling with single or double valves, and are of course essentially voltage amplifiers. Such circuits are generally referred to as *phase-splitting* circuits.

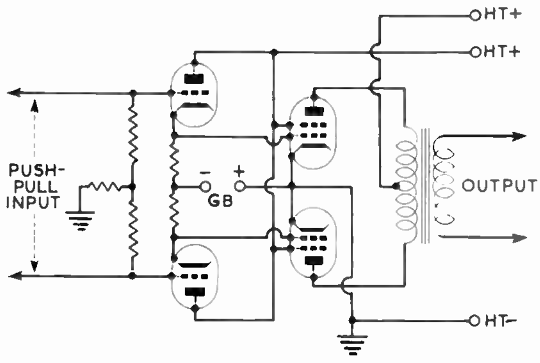


Fig. 9.26. Push-pull cathode-follower driver stage using direct coupling from the cathode load resistances to a class B output stage.

(a) *Paraphase Circuit.* This arrangement employs two similar valves, generally in the form of a double-triode, as shown in **Fig. 9.27.** V_1 acts as a normal voltage amplifier, its output appearing across the series combination $R_1 + R_2$. The part of this output which appears across R_2 is applied to the grid of V_2 and this, due to the phase reversal which occurs in an amplifier, produces an output in R_3 of opposite polarity to that of the voltage across R_1 and R_2. If the output voltages are to be balanced, two conditions are necessary. First that $R_1 + R_2$ should be equal to R_3, and secondly that $(R_1 + R_2)/R_2$ should be equal to the voltage gain of V_2. The use of close-tolerance high-stability resistors

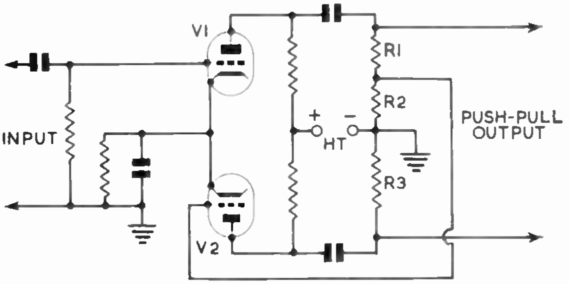


Fig. 9.27. Paraphase phase-splitting circuit. The two triodes may conveniently be in the form of a double-triode.

is recommended for R_1, R_2 and R_3. The value of the cathode bias resistor should be half the normal value for a single valve of the type used.

(b) *Split-load Phase Inverter.* This circuit, shown in **Fig. 9.28**, uses a single valve, the output voltage appearing across

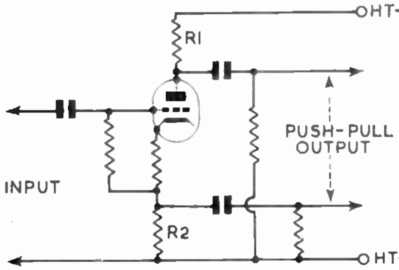


Fig. 9.28. Split-load phase-inverter circuit.

two resistors, one in the anode circuit (R_1) and the other in the cathode circuit (R_2). Since the same current flows in R_1 and R_2, equal voltages are developed across both resistors. The impedances to earth at the two output terminals are, however, different because the anode impedance and the cathode impedance of the valve are not the same.

(c) *Cathode-coupled Phase-splitter.* This circuit (**Fig. 9.29**) uses two similar valves coupled together by a common cathode bias resistor *R*. The signal is applied to one grid only, the other grid being connected to earth through a grid leak. A push-pull output is obtained from the two anodes, but the overall gain is rather less than half that of a single valve under normal *RC* conditions.

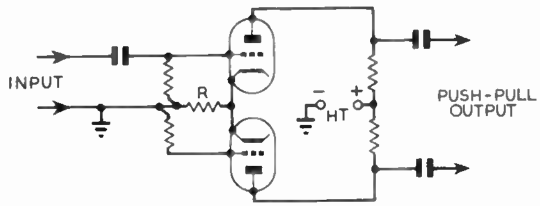


Fig. 9.29. Cathode-coupled phase-inverter circuit.

Balancing Adjustments

The amount of unbalance which may be permitted between the two halves of the output of a push-pull speech amplifier is dependent on the amount of distortion which can be tolerated. In high-fidelity work, accurate balancing is essential, but in amateur transmitting practice this is not so important, and the use of high-stability resistors with a tolerance of $\pm$ 2 per cent or even $\pm$ 5 per cent in the appropriate positions generally ensures an adequate degree of balance.

Nevertheless, it must not be deduced from this that it is unnecessary to check the balance of the input voltages to a push-pull modulator. This may be done by direct measurement of the two halves of the output by a high-resistance a.c. voltmeter (rectifier type), or preferably a valve voltmeter.

The Cathode Follower

Several of the preceding circuits employ fairly high resistances in the cathode circuits of valves. The cathode-follower circuit incorporates a form of negative feedback which arises from the inclusion in the input circuit of the potential developed across the cathode resistor. If the value of this resistor is high enough the signal voltage appearing on the cathode will nearly equal the voltage applied to the grid. Such a circuit is therefore called a *cathode follower*, and although the output voltage is less than the input voltage, it appears across an effective resistance which is very low and is therefore capable of delivering substantial current without waveform distortion.

The Anode Follower

Another circuit of this type which can be used as a phase-splitter is known as an *anode follower*. In this case the feedback is arranged so that the output voltage from the anode is similar to the input grid voltage, the effective output impedance again being very low. Whereas in the case of the cathode follower the output voltage is of the same polarity as the input voltage, in the anode follower the output voltage from the anode is of opposite polarity to the input voltage.

Fig. 9.30 shows how this circuit can be used to produce a balanced push-pull output. The resistor R_1 is made

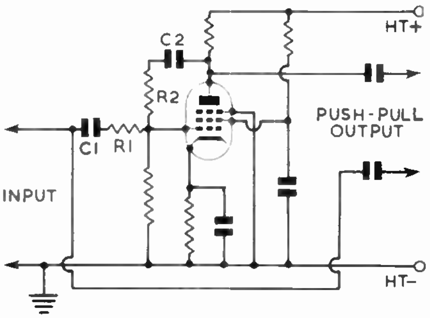


Fig. 9.30. The anode follower used as a phase inverter.

equal to R_2, and the capacitor C_1 equal to C_2. The voltage appearing on the anode will be almost equal to the input voltage if the stage gain, in the absence of feedback, is made very high. This is achieved by the use of a pentode with a high anode resistance load. The push-pull output is taken from the anode and the input terminals respectively through suitable blocking capacitors. The output impedance of the half taken from the anode will, however, probably be much lower than that taken from the input, its impedance being equal to $2/g_m$, where g_m is the mutual conductance of the valve which is used.

Negative Feedback

Negative feedback is commonly used in audio frequency amplifiers, but it is not often applied to the modulators of amateur transmitters. Negative feedback consists of feeding back to an earlier stage a proportion of the output of the amplifier in such a phase relationship that the feedback voltage has an opposite polarity to the input voltage which produces it. Since the effect of the feedback voltage is thus to reduce the effective grid voltage of the earlier stage, the overall result is to reduce the gain, but in return the distortion, the hum and the noise are reduced; in addition

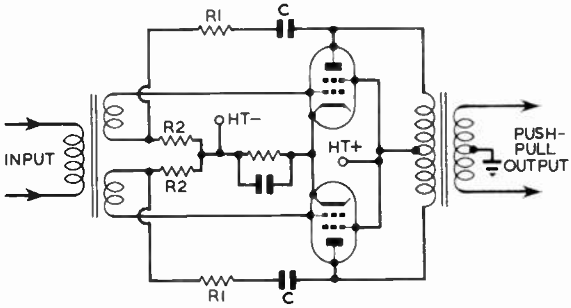


Fig. 9.31. Application of negative feedback to push-pull driver stage.

there are a number of other advantages of which the following are the most important:

(a) Less change in amplification due to changes in supply voltages and valve characteristics.
(b) Improvement of frequency response.
(c) Reduced harmonic and inter-modulation distortion.
(d) The ability to increase or decrease the output and input impedances of the amplifier, if desired.

It is this last property which can be of great value in modulator design, as it presents a convenient means of providing a low-impedance driver stage for a push-pull modulator working under class B conditions.

The feedback voltage may be obtained in various ways, the two most commonly used being:

(a) The provision of a separate secondary winding on the output transformer.
(b) A fairly high-resistance potentiometer across the output of one stage, the tapping point of which is returned to the input either of the same stage or a suitable earlier one.

A typical circuit showing the use of negative feedback in the push-pull driver stage is illustrated in **Fig. 9.31.** The magnitude of the feedback voltage is governed by the proportions of the potential dividers $R_1 R_2$, suitable values for which are 150,000 and 33,000 ohms respectively. The two blocking capacitors C may have a value of 0·1 μF and should be rated at about twice the h.t. voltage used. It will be seen that when a voltage at one of the grids is increased the corresponding anode voltage will be decreased, and hence a feedback voltage opposite to that of the input voltage will be added to the grid, thus providing negative feedback.

Overall Negative Feedback

The use of negative feedback need not be confined to the a.f. equipment only, but may be applied over the whole chain from the radio frequency output of the transmitter right back to the first stage of the speech amplifier. The basic arrangement of such an application is shown in **Fig. 9.32.** Here *L* is a small coil loosely coupled to the output of the transmitter. The voltage developed across this coil is then rectified, and the resulting voltage may be fed back to the speech amplifier.

Such a scheme has the advantage that it takes account of distortion appearing anywhere in the transmitter, a particularly valuable feature since most of the distortion occurs in the modulation transformer and the modulation system. It can also reduce hum introduced from the h.t. supply and from the a.c. used for the valve heaters. It should be pointed out, however, that negative feedback cannot reduce the distortion resulting from over-modulation once the r.f. output has been reduced to zero, any more than it can reduce the distortion in an ordinary a.f. amplifier due to overloading which is severe enough to cut off the anode current.

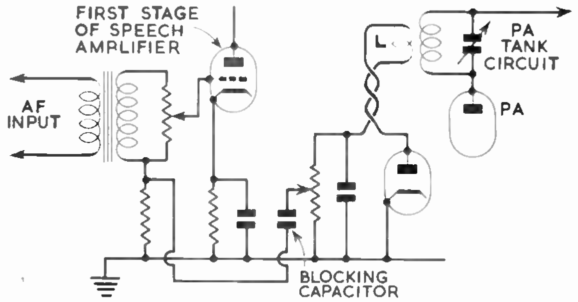


Fig. 9.32. Overall negative feedback. Part of the modulated r.f. output voltage is rectified and fed back to the first stage of the speech amplifier.

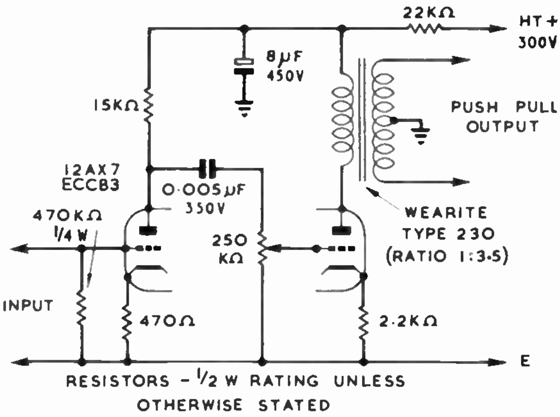


Fig. 9.33. Simple speech amplifier using a double triode valve.

Simple Speech-amplifier Circuit

The circuit of a simple speech amplifier is shown in **Fig. 9.33.** This uses the two halves of a type 12AX7 twin-triode valve in cascade. The low frequency response is restricted by the use of a low value (0·005 μF) of coupling capacitor. At maximum gain, an output of 80 V peak-to-peak is available at the secondary of the centre-tapped transformer for an input of 50 mV at the first grid. The h.t. consumption is 3 mA at 300 V.

The particular characteristics of some valves commonly used in speech amplifiers are given in **Table 9.5.**

THE MODULATOR STAGE

In a voltage-control type of modulation system, such as grid modulation, it is sufficient to use a voltage amplifier as the modulator stage since only a negligible amount of power is required. Other systems, such as the very popular anode modulation, require considerable power output from the modulator stage which must therefore be designed specifically as a *power amplifier*.

One of the dominant factors in the design of a power amplifier is the anode-circuit efficiency. It is uneconomical, in general, to operate such an amplifier in class A owing to the inherently low efficiency of this method, and the majority of modulators are arranged to operate in class

TABLE 9.5

A Selection of Valves Commonly Used as Speech Amplifiers

Type	Class	Heater V	Heater A	E_a	E_{g2}	g_m (mA/V)	r_a	μ	Base
12AT7	Double-triode	6·3 12·6	0·3 0·15	250	—	5·5	10,000 ohms	55	B9A
12AU7	Double-triode	6·3 12·6	0·3 0·15	250	—	2·2	7,700 ohms	17	B9A
12AX7	Double-triode	6·3 12·6	0·3 0·15	250	—	1·6	62,500 ohms	100	B9A
6AM6	Pentode	6·3	0·3	250	250	7·5	1 M ohms	—	B7G
6BR7	Pentode	6·3	0·15	250	100	1·25	2·3M ohms	—	B9A
6BS7*	Pentode	6·3	0·15	250	100	1·25	2·3M ohms	—	B9A
EF80/Z719	Pentode	6·3	0·3	170	170	7·4	0·4M ohms	—	B9A
EF86/Z729	Pentode	6·3	0·2	250	140	1·8	2·5M ohms	—	B9A
EF37A*	Pentode	6·3	0·2	250	100	1·8	2·5M ohms	—	Octal

* Grid top cap

AB or class B. This naturally leads to the use of a push-pull circuit incorporating a pair of valves in order to keep the distortion to a reasonable level.

As already explained the grids and anodes of the valves in a push-pull circuit are connected to opposite ends of balanced circuits, generally in the form of centre-tapped transformers. Because of the close magnetic coupling between the two halves of the windings the grid and anode voltage excursions of one valve are equal and opposite to those of the other; thus the momentary increase in the anode current of one valve during modulation is associated with a decrease in anode current of the other. These changes in anode current are combined in the output transformer so that the current in the secondary load represents an effective addition of the two anode current changes. Any even harmonics produced will be automatically balanced out and there will be little such harmonic current in the output. For this reason a greater distortionless output power can be obtained from two valves working in push-pull than from the same two valves connected in parallel.

The various modes of operation of valves as power amplifiers have been described in Chapter 2 (*Valves*), but the salient differences in the operating conditions are summarized here in order to emphasize the widely differing characteristics of amplifiers operating in each class.

Class A Amplifiers

Class A operation may be defined as that in which the values of grid bias and alternating input voltage are such that anode current flows during the *whole* of the cycle of input voltage. In order that there should be no distortion in the output waveform it is necessary for the grid bias to be chosen so that equal positive and negative excursions of input voltage cause equal positive and negative changes in the anode current. This implies that the operative portion of the valve characteristic should be as nearly linear as possible. The anode current in the presence of a signal remains at the same steady value as in the absence of a signal, since the equal positive and negative changes in anode current do not affect the average value. Thus there is a continuous dissipation of power in the valve, and hence the efficiency of a class A amplifier will always be low.

Class B Amplifiers

A class B amplifier is one in which the grid is biased approximately to the cut-off voltage, i.e. in the absence of any input signal the anode current is very low. Anode current therefore only flows for approximately half of the full cycle. In order to obtain high power-efficiency, class B amplifiers are commonly operated with a large input voltage, the peak value of which may be greater than that of the grid bias, causing grid current to flow over the part of the cycle where the grid potential becomes positive. Because this grid current flows for only part of the cycle, the driver stage must have good regulation to avoid distortion of the input wave by the loading effect of the grid current.

Because the anode current flows for only half of each cycle, class B operation at audio frequencies demands the use of two valves in push-pull in order that both halves of the input waveform may be reproduced without distortion. Such a push-pull class B amplifier can have a high anode efficiency since, in the absence of an input signal, the anode current is very low. However, owing to the variation in anode current with the input voltage, an anode power supply with good regulation is essential. Provided that the requirement of low grid-circuit impedance is met, class B operation enables full advantage to be taken of the portion of the anode-current/grid-voltage curve which lies in the positive grid region. This part of the curve is usually fairly linear. The lower portion of the characteristic is badly curved, but the curved portions of the two valve characteristics compensate each other when they are combined in the output circuit. Thus waveform distortion can be kept reasonably low, while still retaining high anode efficiency, so that class B amplification is well suited for use in high-power audio amplifiers.

Special valves with a high amplification factor, which are suitable for operating at zero grid bias, have been developed for this class of operation, typical examples being the DA41 and the DA42 which when used in pairs will give an output of up to 200 watts.

Class AB Amplifiers

In class AB operation, which is an intermediate stage between class A and class B, the grid bias is insufficient to reduce the anode current to zero, and hence anode current flows for more than half of the input cycle but for less than the whole cycle. In class AB1, the input signal is not permitted to be large enough to drive the grids positive, whereas in class AB2 grid current is allowed to flow, and consequently the driver stage must be capable of delivering appreciable power. Since the fixed grid bias is higher than in class A, larger signals can be handled without grid current, and the standing anode current is less. The efficiency and maximum power output are therefore greater than for class A. Operation in push-pull is, of course, still necessary to avoid distortion arising from the use of the curved part of the valve characteristic.

To summarize the foregoing, as the mode of operation is changed from class A through class AB to class B, the reduction in standing anode current and anode dissipation under no-signal conditions results in a considerable increase in the power-handling capacity and the anode efficiency. Against this, however, must be set the fact that both the h.t. and grid-drive requirements become more critical. A greater input voltage is required and, in the case of class AB2 and class B operation, the driver stage must be designed to supply the grid power and to have good regulation. Since the anode current varies over a wide range the anode power supply must also have good regulation.

As a general rule class B operation is liable to produce more distortion than class A, but it must be remembered that the overall distortion depends on many factors, and provided that the regulation of the power supplies and the driver stage is good, and particularly if negative feedback is employed, the overall distortion can be reduced to a very low level.

Ultra-linear Operation of Tetrodes and Pentodes

In ultra-linear or " distributed load " operation, the screen supply is obtained from taps on the primary winding of the output transformer as shown in **Fig. 9.34**. As the position of the tap is varied so that the ratio $a : b$ changes from zero to 100 per cent, the operating conditions of the output valves

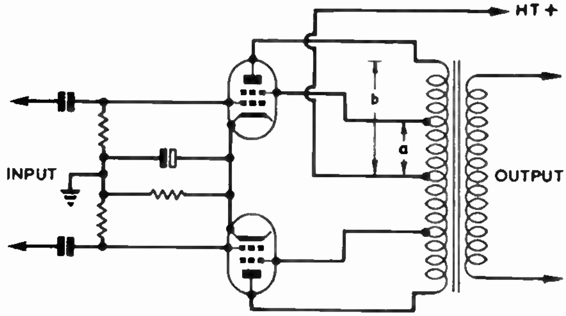


Fig. 9.34. Basic arrangement of an ultra-linear output stage.

change from those of a tetrode to those of a triode.

Ultra-linear operation thus provides a compromise between the high efficiency of tetrodes and the low distortion and uncritical load impedance of triodes. The ratio of $a : b$ is not very critical and can be between 20 per cent and 50 per cent with an optimum value of about 40 per cent.

Ultra-linear operation is now almost universal in high fidelity audio amplifiers but its use is not common in amateur radio modulators. It has undoubted advantages for this service, i.e. the matching of the modulator to the r.f. stage is not so critical and as the screen supply is obtained automatically, a high-wattage screen potentiometer or stabilizer valves are not required. It should be noted that as the anode and screen voltages are effectively equal, the output valves must be so rated.

The Woden multi-ratio modulation transformer may be used for ultra-linear operation as the tapping points 2 and 5 on windings 1–4 and 3–6 respectively provide a ratio of about 40 per cent if the anodes are connected to 3 and 4 and a ratio of about 60 per cent if the anodes are connected to 1 and 6.

The Use of Transistors in Modulation Equipment

The main features of transistors, viz.

(a) small space occupied;

(b) higher overall efficiency resulting from absence of heater;

(c) much smaller components possible as a result of the low supply voltage;

makes their use highly advantageous in portable or mobile equipment.

It is, however, probably true to say that a standard practice for the use of transistors in modulation equipment for fixed station is only now becoming established. Undoubtedly, the features referred to earlier can be usefully employed in fixed stations, for example, a simple one- or two-stage head amplifier for a crystal microphone, complete with battery can be built into the microphone case.

A speech amplifier for an output stage which does not run into grid current may be made in a very small screened box, the necessary low voltage supply being obtained from the rectification and smoothing of the 6·3 volts-heater supply of the output stage.

Such applications are dealt with in Chapter 3 (*Semiconductors*).

Audio outputs of 30–50 watts may be obtained from transistors operating at a supply voltage of 12 volts. Such an application is obviously very suitable for mobile use and a typical modulator is described later in this chapter.

Outputs of 80–100 watts may be obtained from relatively cheap transistors, but the large heat-sinks and the low-voltage, high-current power supply required (24 volts at 4–5 amps) may or may not be considered a disadvantage.

MODULATOR DESIGN AND CONSTRUCTION

To design any particular modulating equipment it is necessary first of all to know (a) the voltage output of the microphone to be used and (b) the maximum audio output power required to modulate the transmitter fully. The intermediate stages can then be designed to provide the necessary voltage gain. The initial step in the design is the choice of a suitable output stage which will deliver the calculated amount of power.

Power Required for Full Modulation

As stated at the beginning of this chapter, the audio power required for 100 per cent modulation of a triode amplifier when using the anode modulation system is one half of the d.c. power input to the r.f. amplifier, but in practice it must be remembered that the modulation transformer will not be free from losses: a typical efficiency is about 90 per cent. For this reason, about 10 per cent more audio power than the theoretical value should be provided. Thus, in order to anode modulate fully a p.a. stage having an input of 150 watts an audio power of 75 watts + 10 per cent, say 83 watts, is required from the modulator valves. In a tetrode or pentode p.a. stage in which both the anode and the screen grid are modulated, a further 10–20 per cent of audio power is required for the screen grid. Hence it is necessary to use modulator valves capable of delivering about 100 watts of audio power in order to fully modulate such stages when the input is 150 watts. Since a small factor of safety is always desirable, a suitable figure on which to base such a design could be reckoned as 120 watts.

Modulation of Transistor R.F. Amplifiers

It is difficult to obtain absolute linearity when amplitude modulating transistor r.f. amplifiers operating in Class C.

Results can be greatly improved in many cases by the simultaneous assymetric modulation of the driver stage in addition to the normal modulation of the final stage.

The audio power required is the normal amount to be expected plus a small percentage for the driver stage.

Typical Modulator Valves

The approximate economic ranges of audio power which may be obtained from different types of valves operating in various circuit conditions are given in **Table 9.6.** It will be seen that the maximum requirements of the British amateur station may be most conveniently met by using tetrodes or pentodes of 25 or 35 watts anode dissipation operating in class AB1, AB2 or B.

Basic operating conditions of valves commonly used as modulators in this country are given in **Table 9.7.** Reference to this table shows that under class AB2 or class B conditions there is considerable variation in both anode and screen-grid currents between zero and maximum signal conditions. There will consequently be a similar variation in the bias if this is obtained from the voltage drop across the cathode

TABLE 9.6

Modulation Power obtainable under various circuit conditions

Type of Valve	Class A	Class AB1	Class AB2	Class B
Noval-based	5 watts	12–15 watts	25–30 watts	—
25 watt tetrode/pentode (single-ended) ...	10 watts	30 watts	50–70 watts	—
25 watt tetrode/pentode (double-ended) ...	15 watts	30–50 watts	80–120 watts	120 watts
35 watt tetrode/pentode (single-ended) ...	25 watts	60 watts	100 watts	150 watts
40/50 watt triode	20 watts	40 watts	100 watts	150–200 watts

resistor, and also in the screen grid voltage if this is obtained through the usual dropping resistor.

The effect of such voltage variation is to cause the operating conditions of the valves to be continuously changing between zero and maximum signal levels, thereby preventing the full output of the valves from being obtained.

It must be emphasized therefore that in order to obtain the maximum output from tetrodes or pentodes operating in class AB2 or class B it is essential that all supply voltages, anode and screen grid and grid bias, are maintained as nearly constant as possible. As a general rule, all supply voltages should not vary by more than 5 per cent between zero and maximum signal conditions.

It will be appreciated, therefore, that fairly stringent requirements are placed upon the design of the power unit for such amplifiers. In particular, the grid bias voltage should be obtained from a small separate supply.

These conditions are not encountered to the same extent when triodes, and particularly those designed for zero bias operation, are used. However, such valves require a higher anode voltage and more grid driving power for the same power output.

A compromise which is fairly commonly used is a pair of triode-connected 807s operating in class B with zero bias. In this arrangement, shown in Fig. 9.35, the input taken to the screen grids to which the control grids are connected through 22,000-ohm resistors. This circuit is

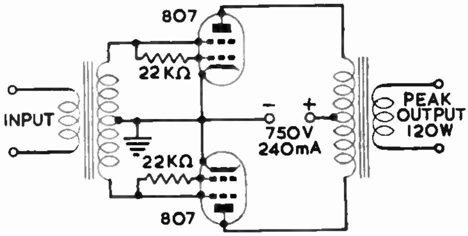


Fig. 9.35. A pair of beam tetrodes, each with its two grids connected together, may be used as zero-bias triodes in a class B amplifier.

capable of an output of 120 watts and requires an anode supply of 750 volts at 240 mA. These figures are similar to those for a pair of 807s operating in class AB2. However, in the latter condition stable sources of anode and grid bias voltages are essential (although a grid-drive power of

TABLE 9.7

Characteristics of Typical Modulator Valves

Type	Class	Heater (V)	Heater (A)	V_a	V_{g2}	I_a (No signal) (mA)	I_{g2} (No signal (mA)	I_a (Max. signal) (mA)	I_{g2} (Max. signal) (mA)	Peak Input Voltage (grid-to-grid) (V)	Load Resistance (anode-to anode) (ohms)	Power Output (watts)
6AM5 ...	A	6·3	0·2	250	250	2 × 11	2 × 1·6	2 × 13	2 × 4·1	32	24,000	4
6AQ5 ...	AB1	6·3	0·45	250	250	2 × 35	2 × 2·5	2 × 39·5	2 × 6·5	30	10,000	10
6V6 ...	AB1	6·3	0·45	285	285	2 × 35	2 × 2	2 × 46	2 × 7	35	8000	14
EL84 ...	AB1	6·3	0·76	300	300	2 × 40	2 × 3	2 × 72	2 × 8	70	8000	17
6L6 ...	A	6·3	0·9	270	270	—	—	2 × 72·5	2 × 8·5	35	5000	18·5
6F6 ...	AB2	6·3	0·7	375	250	2 × 54	2 × 8	2 × 74	2 × 18	55	10,000	19
6L6 ...	AB1 *	6·3	0·9	360	270	2 × 44	2 × 2·5	2 × 50	2 × 8·5	45	9000	24·5
KT55 ...	AB1	52	0·3	200	200	2 × 110	2 × 7·5	2 × 113	2 × 22·5	29	2000	25
5763 ...	AB2	6·3	0·75	300	225	2 × 13·5	2 × 1·1	2 × 70	2 × 9	71	4500	25
6BW6 ...	AB2	6·3	0·45	315	285	2 × 35	2 × 2	2 × 77·5	2 × 8	80	5000	30
KT66 ...	AB1	6·3	1·27	390	275	2 × 52	2 × 2·5	2 × 62·5	2 × 9	70	8000	30
KT88 ...	AB1	6·3	1·6	330	330	2 × 85	2 × 7·5	2 × 95	2 × 20	80	3400	40
6L6 ...	AB1 *	6·3	0·9	360	270	2 × 44	2 × 2·5	2 × 102·5	2 × 8	72	6000	47
807 ...	AB1	6·3	0·9	600	300	2 × 40	2 × 3	2 × 75	2 × 9	60	10,000	47·5
KT66 ...	AB1 *	6·3	1·27	480	375	2 × 40	2 × 1·5	2 × 87·5	2 × 9·5	80	5000	50
KT88 ...	AB1 *	6·3	1·6	450	345	2 × 50	2 × 4	2 × 120	2 × 17·5	75	4000	65
EL37 ...	AB1 *	6·3	1·4	400	400	2 × 50	2 × 6	2 × 138	2 × 36	60	3250	69
KT77 ...	AB1 *	6·3	1·4	600	300	2 × 40	2 × 4	2 × 100	2 × 10	94	11,000	72
KT88 ...	AB1 *	6·3	1·6	600	330	2 × 50	2 × 3	2 × 125	2 × 16	70	5000	100
EL34 ...	AB2 *	6·3	1·5	800	400	2 × 25	2 × 3	2 × 91	2 × 19	67	11,000	100
807 ...	AB2 *	6·3	0·9	750	300	2 × 26	2 × 2·5	2 × 120	2 × 10	92	7000	120
807† ...	B	6·3	0·9	750	—	2 × 6	—	2 × 120	—	100	6650	120
6146 ...	AB1	6·3	1·25	750	195	2 × 12	2 × 1	2 × 110	2 × 13	100	8000	120
KT88† ...	B	6·3	1·6	750	—	2 × 10	—	2 × 140	—	360	6000	150
DA41† ...	B	7·5	3·1	1000	—	2 × 22	—	2 × 140	—	220	7000	175
TT21 ...	AB1 *	6·3	1·6	1250	300	2 × 28	2 × 1	2 × 130	2 × 13	71	15,000	200

* Fixed bias. The figures shown here relate to pairs of valves operating in push-pull. † Triode-connected.

only 0·2 watt is required), whereas the zero-bias class B circuit does not require any grid bias or screen-grid supplies, but it does require a speech amplifier having a low-impedance output and capable of supplying 5·5 watts. Other valves may, of course, be used in this arrangement, e.g. the GEC type KT88 is capable of an output of 150 watts at an anode voltage of 750 volts and a grid driving power of 7 watts.

The Modulation Transformer

The modulation transformer is a most important component, for its object is to superimpose the output of the modulator upon the d.c. supply to the r.f. stage in order to produce the required modulation: it must also convert the modulating impedance of the r.f. amplifier to the load impedance required by the modulator valves. To avoid distortion the iron core should be of generous proportions, and modulation transformers should always be used well within their ratings.

The choice of a suitable modulation transformer should receive careful attention. Two conditions must be fulfilled: (a) the ratio must be correct in order to transform the impedance of the modulated r.f. amplifier to the optimum load required for the modulator, and (b) the transformer must be capable of handling the maximum power output over the audio-frequency range without undue distortion. Multi-ratio modulation transformers, of which the Woden types UM1 and UM3 are excellent examples, are frequently used, since they enable adjustments of the matching conditions to be made. The use of miscellaneous transformers, such as mains transformers, which are not designed for audio work is not recommended as they rarely have the correct impedance ratio, and magnetic saturation of their cores is likely to occur even with relatively low values of direct current.

The Driver Stage

When the basic design of the output stage has been determined, the next step is the choice of a suitable driver stage appropriate to the manner in which the output valves are to be run. As discussed earlier, operation in class B or class AB2 is usual, in which case medium-power triode valves or triode-connected tetrodes such as the 6V6 are commonly chosen as drivers. It is advisable to employ a driver stage which is capable of providing about 50 per cent more power than the estimated power required to drive the output stage. If the output valves are to operate in class A or class AB1 the driver stage needs only to be a voltage amplifier using either transformer coupling or *RC* coupling to the output stage with one of the centre-tapped transformer or phase-splitting circuits already described. The voltage-amplifying stages preceding the driver stage must have sufficient gain to produce the necessary amount of drive to the output stage when supplied with the normal output voltage from the microphone. It is good practice to make the first stage a high-gain amplifier (e.g. a pentode), but if further amplification is required one or two stages of medium gain are preferable to a second high-gain stage in order to avoid instability.

Frequency Range

The frequency range of the modulator should be restricted to the minimum necessary for intelligible communication, i.e. to about 300–3000c/s. The low frequency response may be controlled by a suitable choice of grid coupling capacitor and grid leak (see voltage amplifiers, page 9.13.) However, this is not of overriding importance as the l.f. response inevitably tends to be reduced by any transformers used in the speech amplifier and particularly by the modulation transformer.

A typical modulator valve for amateur use—the G.E.C. KT88. ***(Photo by courtesy of M-O Valve Company).***

It is the reduction of the upper frequency limit which is far more important as this defines the overall bandwidth of the transmission. The upper frequency limit may be restricted by the inclusion of shunt capacitors (value 0·0005-0·01μF) across the anode loads of any RC coupled valves in the speech amplifier or by connecting a capacitor of similar value and of suitable working voltage across the secondary winding of the modulation transformer. A more satisfactory method of reducing the bandwidth is by the use of a low pass filter.

It must be emphasized however that a modulator should be capable of satisfactory, i.e. stable and distortion-free, operation over the full audio frequency range. An amplifier which is stable only because its frequency range is significantly restricted cannot be considered to be a satisfactory design and will probably cause trouble sooner or later.

Modulator Power Supplies

The power supplies required by class A or class AB1 modulators are relatively simple in view of the small variation in current drawn. A capacitor input smoothing circuit is generally adequate.

In class AB2 or class B operation, the anode, screen grid and grid bias voltages should not vary by more than 5 per cent between zero signal and maximum signal conditions; in other words, the dynamic regulation of the power supply must be very good.

In order to achieve this, a choke-input smoothing circuit is necessary. The choke and transformer should be of generous design with low d.c. resistance. The choke should have the minimum inductance consistent with satisfactory operation of a choke-input filter and the largest possible value of smoothing capacitance should be used (at least 32 μF).

The regulation may be further improved by the use of low impedance rectifying elements such as gaseous rectifier valves or semiconductor rectifiers.

A separate well regulated supply should be used for the screen voltage, but care must be taken to ensure that the screen volts cannot be applied in the absence of anode volts; alternatively it may be obtained from the anode supply by utilizing the voltage drop across gas stabilizer valves such as the VR150 or S130. The negative bias voltage may be conveniently obtained from a small separate supply which uses a reversed 240 volts/6·3 volts heater transformer fed from the heater circuit.

Reference should be made to Chapter 17 (*Power Supplies*) for details of the design of power units.

Construction Practice

No matter how good the theoretical design of a modulator may be its virtues will be wasted unless the construction is given careful thought. In addition to the usual practical points contributing to good workmanship, it is necessary to ensure the correct choice of components suitable for the appropriate working conditions in regard to their structural features as well as their circuit values. It is particularly important that full consideration be given to the layout of the components and the supply leads in order to avoid unwanted feedback between points of the circuit having high signal-potential differences and between low-level input circuits and power supply circuits. This matter of unwanted couplings between different parts of the circuit forms one of the most difficult problems in the construction of audio-frequency apparatus. In addition care should be taken to avoid circulating earth currents; the use of a common earth rail which is connected to chassis at only a single point, can also sometimes be helpful.

The general layout of the modulator is usually determined to a large extent by its power rating. As an example, consider a modulator suitable for a 25 watt transmitter. This would need to have a maximum output of about 15 watts, and would most likely use a pair of small tetrodes operating in class AB1. All the components, including the modulation transformer and power supply would be of quite small dimensions, and such a modulator could easily be accommodated on a chassis measuring 14 in. × 12 in. A suggested layout is shown at (A) in **Fig. 9.36.** The main requirement in such a scheme is that both the modulation transformer and the low-level input stages should be situated as far as possible from the mains transformer and smoothing chokes.

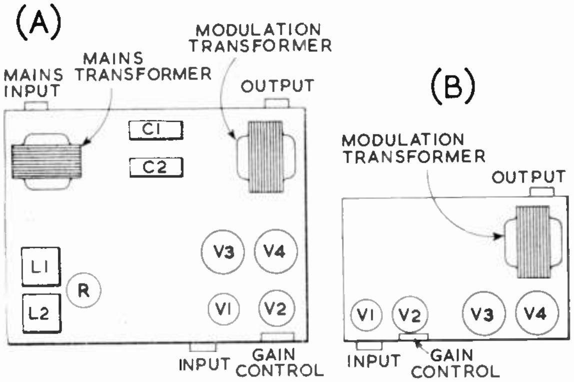


Fig. 9.36. Suggested alternative layouts for a 15-watt modulator. In (A) the power supply is incorporated with the amplifier-modulator on a 14 in. × 12 in. chassis. In (B) the power supply is excluded and the size of the modulator chassis can be considerably smaller. V1, V2 speech-amplifier valves; V3, V4 modulator valves; C1, C2 smoothing capacitors; L1, L2 smoothing chokes; R power-supply rectifier.

The stray magnetic field from a power-supply transformer or from a filter choke can induce mains-hum pick-up in an a.f. transformer if their cores happen to be oriented in such a way that there is mutual coupling between them. Such coupling is best minimized by a trial-and-error method, moving one or other of the transformers to various positions and placing it at various angles while the equipment is in operation and thus finding the setting where the hum pick-up is eliminated. (*Caution:* Extreme care should be taken when working on " live " equipment.) Likewise there may be stray coupling between the input and output transformers of the a.f. amplifier itself, and the result may then be an appreciable amount of feedback (either positive or negative) which could seriously affect the performance. This is rather more difficult to detect and to cure, but if the transformers are widely separated and if their cores are arranged to be mutually at right angles, the centre of one transformer being located on the magnetic axis of the other, the stray coupling will quite probably be negligible.

The layout shown at (B) in **Fig. 9.36** suggests an alternative arrangement in which the power supply is not included thus permitting a smaller chassis to be used.

Turning now to the other extreme and considering a modulator with an output power of up to 120 watts, the modulation and h.t. transformers, the smoothing chokes and the capacitors become large and heavy, and so it becomes necessary to separate the modulator and the power supply to avoid an unduly heavy unit. Generally the most convenient way to do this is to use one chassis to accommodate the modulator and speech amplifier and the power supply for the speech-amplifier stages, and a second chassis to accommodate the power supply for the output stages. A typical layout for such a scheme is shown at (A) in **Fig. 9.37.** The power supply for the speech-amplifier stages need only use a small transformer (e.g. a standard 250 volt 60 mA type), a small choke and a two-section electrolytic capacitor. If a separate heater transformer for the output stage is required, it is advisable to mount this on the modulator chassis as it reduces to a minimum the number of interconnecting wires required. Another suitable layout for a high-power modulator using a driver stage coupled with a driver transformer is shown at (B) in Fig. 9.37.

In order to keep the microphone leads short when the transmitter is remote from the operating position, it is sometimes convenient to make the first one or two stages of the speech amplifier as a *head-amplifier* unit separate from the remainder of the modulator. Such a head amplifier, which can have its own gain control, may be combined

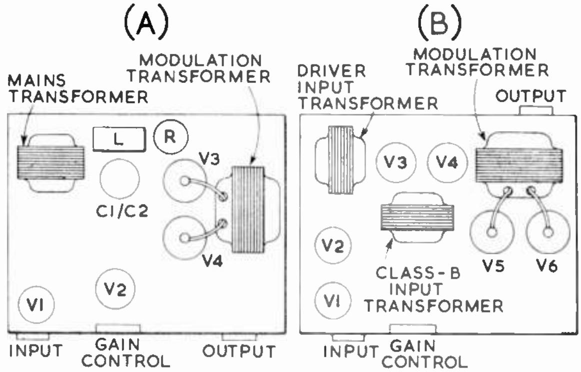


Fig. 9.37. Suggested alternative layouts for a high-power modulator. In (A) the small power supply required for the speech amplifier is included. In (B) the power supplies are excluded as the space is needed for the driver-input and class B input transformers. A suitable chassis size is 14 in. × 12 in.

with any other transmitter controls at some suitable position on the operating desk.

The actual mechanical form of construction is not very important, the most convenient being the familiar inverted-tray type of chassis. Quite successful amplifiers may be built on a simple chassis consisting of a flat sheet of metal supported on a wooden frame. Where rack mounting is desired an alternative form of construction which can be especially recommended for high-power equipment is to mount the heavy components such as transformers and chokes directly on the panel, the smaller components such as the valves being supported on brackets mounted on the transformers. In this way the use of a chassis is avoided.

GENERAL DESIGN FEATURES

Without doubt the most critical stage in any a.f. amplifier is the first voltage-amplifying stage. Its grid circuit is very prone to hum pickup and the signal voltage in it is quite low, resulting in a poor signal-to-hum ratio. Consequently it is advisable to screen the microphone input jack and all the wiring, together with the microphone transformer (if any) and the grid resistor, right up to the grid of the valve. If one of the double-ended older valves is used, i.e. one in which the grid connection is at the top cap, it should at least have a screened top cap and preferably a complete screening can. For similar reasons the gain control should be inserted *after* the first valve rather than in its grid circuit.

Another troublesome fault frequently encountered in the input stage is *microphony* which is apparent as a ringing sound when the valve is tapped and is due to the vibration of the valve electrodes. It can usually be cured by substituting one of the special valves such as the GEC type Z729, Mullard EF86 (B9A base), and Mullard EF37A (octal base) which are provided with a very high degree of internal bracing for reducing the relative movement between the electrodes.

By comparison with the input stage, the remainder of the amplifier is relatively straightforward. If the layout is logically arranged and if adequate decoupling is included the biggest bugbear of amplifier construction—undesired feedback—can be avoided. Any layout which brings the input circuit of the amplifier near to the output circuit, or which brings the wiring of an anode circuit near to the grid circuit of the same valve should always be avoided.

The Output Stage

If single-ended valves are used in the output stage, it is advisable to mount them in ceramic or other good-quality valveholders, as the high voltage appearing at the anode pin has been known to break down the insulation of ordinary moulded bakelite valveholders. Stopper resistors for the grid circuits and/or anode and screen circuits, where used, should be placed as near as possible to the appropriate valve pins. Grid stoppers must not be used, for obvious reasons if the output stage runs into grid current, i.e. if it operates in class AB2 or class B. Such resistances should have the lowest value that will ensure the prevention of parasitic oscillations. Typical values are 10 ohms for anode stoppers, 100 ohms for screen stoppers and 1000 ohms for grid stoppers The grid stoppers of valves with top-cap grids should be attached directly to the grid connector.

Heater Wiring

If the heater winding of the power supply transformer has a centre tap it should be earthed and the valve heaters wired with twisted flex or screened cable to reduce the electric and magnetic fields. In any case the grid and anode leads should be kept as far away from the heater wiring as possible. The heater wiring is preferably laid along the corners formed by the ends and sides of the chassis, and if this is done screened cable will probably be unnecessary. If in spite of these precautions the hum from the heater wiring still persists it is sometimes beneficial to disconnect the centre tap from earth and connect a small potentiometer or " humdinger " across the heater circuit, the slider being connected to earth. A careful adjustment of this potentiometer will generally balance out the last traces of hum.

A major difficulty in tracing hum in new designs is that it may be arising from several different causes, so that changes which reduce hum from one source may have little effect on others. It is also very easy to produce misleading symptoms: for example, clipping on extra smoothing capacitors may actually cause the hum level to rise because incorrect earthing of the capacitor may inject some of the ripple current into a signal circuit—but this would not indicate that the unit had excessive smoothing. Smoothing capacitor connections to the earth " bus " line are, in fact, important and should be the mirror image of the h.t. positive distribution system (for this reason common earth connections in multiple section capacitors are sometimes to be avoided).

The importance of using twisted leads for a.c. heater wiring is well known and correct and incorrect methods are shown in Chapter 4—(*H.F. Receivers*). It is also necessary to see that twisted leads, sensibly routed, are used for pilot lamps and a.c. primary wiring. While a.c. leakage from the heater circuit usually produces 50 c/s hum as opposed to the 100 c/s hum from full-wave h.t. ripple, it is worth noting that electronic conduction between cathode and heater due to the heater acting as an " anode " with low applied a.c. potentials (the heater voltage), can cause either 50 or 100 c/s hum. The remedy for this type of hum is to put a positive bias of about 50 volts on the heater line, provided that the cathode/heater insulation will stand this. The bias potential saturates the electronic conduction.

Where equipment develops hum after being in use for some time, it can usually be traced to defective valve inter-electrode insulation, or faulty electrolytics.

Earthing

The actual earthing of the various points of the circuit should be carefully planned. Indiscriminate earthing of components to various points on the chassis is liable to set up circulating currents in it. It is better to take all the earth connections belonging to each stage to one point, as in r.f. technique. An alternative method particularly applicable to a.f. amplifiers is to run a busbar of thick tinned copper wire (10 or 12 s.w.g.) along the whole length of the chassis to which all earth connections may be made. One end of this busbar should be reliably connected to the chassis. It is important that all connections to the windings of input transformers should be made by means of twisted wires going through the same hole. On no account should leads to low-impedance transformer windings be taken through separate holes in a steel chassis.

Choice of Components

As in any other electronic device, the use of components of adequate rating is essential for reliable operation and long life. A power rating of ¼–½ watt is adequate for many resistors in audio equipment, but anode resistors will often need to be of a higher rating. The actual dissipation should be checked in each case from a knowledge of the current which the resistor carries, and a reasonable factor of safety allowed. This is also important for the cathode bias and screen-voltage potential dividers in which the power to be dissipated is sometimes quite high. All capacitors should have a reasonable safety factor in regard to the working voltage. Grid-coupling capacitors should be of the mica or ceramic type because even a very slight leak may have disastrous effects on the following valve.

Metering

It is helpful to be able to measure the anode or cathode current of each valve in the modulator since this provides an immediate check on the operating conditions of each stage. A low value resistor can be wired in series with each valve, a single low-reading milliammeter being switched across each resistance. The resistance should be considerably higher than that of the meter in order not to affect its accuracy appreciably. As an alternative, a meter may be plugged into each stage by means of a closed-circuit jack which completes the circuit when the plug is removed.

The two valves in the output stage should preferably have independent anode current meters of their own, although a single meter for the combined anode currents is commonly employed. If separate meters are not incorporated, some means whereby a voltmeter can conveniently be connected across the two cathode resistors should be provided in order to check the balance of the two output valves when setting-up the modulator and during operation: the two currents should not differ normally by more than 10 per cent.

Interconnections

All interconnections should be as short and direct as possible, with emphasis on the adequate bonding together of the chassis of the modulator and its power unit and the transmitter with thick copper wire. Circuits of either 200 or 600 ohms impedance are commonly used for distribution of audio-frequency energy, but for the short cable lengths commonly encountered in amateur equipment this is not important, and audio input and output circuits of any convenient impedance up to several thousand ohms can be connected with good quality screened flexible cables.

Microphone-to-modulator. Microphones of the moving-coil or carbon type are preferably connected to the modulator or head amplifier by screened twin flex but coaxial cable is often used for crystal microphones. In order to reduce hum pick-up a preferred scheme for crystal microphones is to use screened twin flex with the screening connected to the microphone case at one end and, together with the "earthy" microphone lead, to the earthed chassis at the other end. The other lead connects the live side of the microphone to the grid of the valve. As far as possible the screening should be continuous from the microphone to the grid of the first valve. If coaxial cable is used the appropriate coaxial plugs and sockets will be required. When screened twin cable is used it should be terminated in a screened two-pin connector.

The use of coaxial cable for low-impedance moving-coil microphones is bad practice since it may result in serious hum pick-up: a screened twisted-pair cable is much to be preferred.

Modulator-to-transmitter. Owing to the high voltages which exist across the secondary of the modulation transformer, it is necessary to use a good quality heavily insulated wire for making the connections from it to the transmitter. The leads should be as short as possible, and preferably placed clear of metalwork and other leads. A point often overlooked is that in a transmitter which is also used for telegraphy, the secondary winding of the modulation transformer should be either short-circuited or disconnected from the power amplifier h.t. supply altogether during telegraphy operation. If this is not done the high voltage induced in the windings when the current through the secondary is keyed is liable to cause quite severe key clicks and may result in the breakdown of the transformer insulation.

Radio-frequency Pick-up

Instability in a phone transmitter can often be caused, particularly when working on the higher frequency bands, by the existence of r.f. fields or currents in the vicinity of the high-gain a.f. amplifiers, the first voltage-amplifying stage being usually the most susceptible. Rectification of the modulated r.f. voltage takes place, and the resulting a.f. voltage is fed back into the transmitter through the modulator. At some frequency the phase will probably be such as to produce positive feedback, and the whole system will oscillate at an audio frequency. This form of instability can best be cured by excluding the r.f. fields from the a.f. equipment. The problem is similar to that of the suppression of one of the forms of TVI. Short, direct and screened wiring between the modulator, transmitter and power circuits, in conjunction with the liberal use of r.f. chokes and bypass capacitors of 0·001–0·01 μF at each end of the power line is generally sufficient to effect a cure.

The effect of an r.f. choke may be achieved very simply by slipping a small ferrite bead over the grid lead of a valve as close as possible to the grid pin on the valveholder. The ferrite bead increases the inductance of the lead at that point without, of course, increasing the resistance and so acts as an r.f. choke. The bead should be fixed in position by an adhesive.

SPEECH CLIPPING AND VOLUME COMPRESSION

Before considering speech clipping in detail it is necessary to examine further some aspects of radio telephony which were discussed earlier in this chapter.

Over-modulation, or the breaking up of the carrier wave by excessive audio output power from the modulator, gives rise to severe harmonic distortion and to the production of spurious sidebands. This may cause the total bandwidth occupied by an over-modulated transmission to extend to 50 kc/s or more. While this excessive bandwidth may not be apparent to a distant listener to whom it may only appear as a badly distorted signal of little more than normal bandwidth, it can cause extreme inconvenience to stations in the immediate vicinity. These spurious sidebands, which only occur on the peaks of modulation, may be received at considerable strength at stations within a radius of a few miles. This form of interference, or "splatter" as it is usually known, obviously prevents the reception of weak signals on frequencies adjacent to the over-modulated transmission and is an offence under the terms of the amateur transmitting licence.

For intelligible communication by speech, it is not necessary to transmit audio frequencies higher than about 3 kc/s. The total bandwidth of an a.m. communications transmitter need not therefore exceed 6 kc/s. This important point is perhaps not appreciated as much as it should be. There is unquestionably no justification in the over-crowded amateur bands of today for telephony transmissions which occupy a bandwidth greater than that really necessary for intelligible communication.

Although a pure sine wave and a speech waveform may have the same peak value, due to the peaky nature of the speech waveform, the average value of the latter is lower than that of a sine wave. This is shown in **Fig. 9.38**, where (A) represents a pure sine wave having just sufficient amplitude to modulate a given transmitter to a depth of 100 per cent. (B) is a typical speech waveform also with just sufficient amplitude to give 100 per cent modulation on peaks. In the latter case, if the audio level is conscientiously adjusted to give 100 per cent modulation on peaks which may not occur very often, the average modulation depth will be much less than 100 per cent, probably of the order of 30 per cent. If

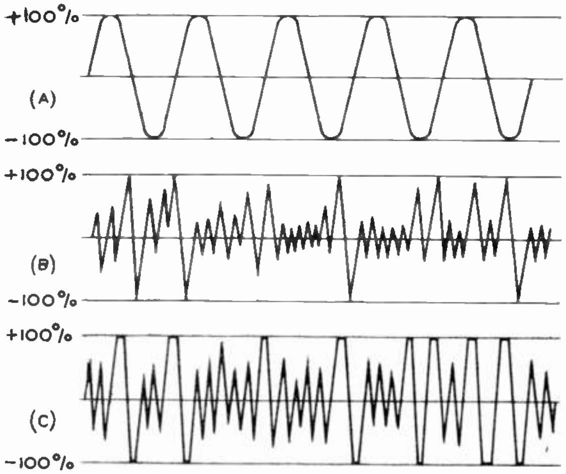


Fig. 9.38. Speech clipping. (A) represents a sinusoidal modulation signal the peak amplitude of which is sufficient to modulate a given carrier wave to a depth of 100%. (B) shows a typical speech waveform of equivalent peak amplitude to (A). Full 100% modulation is reached relatively infrequently and the average modulation level is low. In (C) the same speech waveform has had its peaks removed by clipping, and the mean amplitude of the modulating voltage has been increased so that 100 per cent modulation is again reached. As compared with (B), there is greater average speech power in the clipped waveform (C).

the audio level is increased in order to increase the average modulation depth, then overmodulation will obviously occur on the peaks.

Principle of Speech Clipping

The process of speech clipping consists simply of increasing the audio input level and at the same time clipping or limiting the output level at the value corresponding to 100 per cent modulation and at the same time increasing the audio level as shown in Fig. 9.38C. Thus overmodulation is prevented and the average level is increased.

The degree to which this process can be carried is limited by the amount of distortion of the speech waveform that can be tolerated, and if taken to excess the quality of the speech will deteriorate and intelligibility will be diminished. Furthermore the squaring of the individual peaks as shown at (C) in Fig. 9.38 would introduce very high audio-frequency

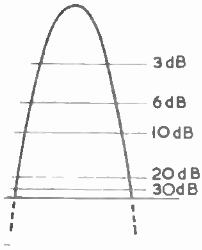


Fig. 9.39. The degree of speech clipping is expressed in decibels. The diagram shows the corresponding amounts by which the maximum amplitude of the waveform must be reduced.

components which will defeat the object of the operation by again increasing the bandwidth occupied. Any speech-clipping system must therefore be followed by a low-pass filter to restrict the audio bandwidth to the required 3 kc/s. The degree of speech clipping is defined as the ratio of the peak level to the clipped level, as shown in **Fig. 9.39.** Small amounts of clipping, of the order of 4–6 db, will give an appreciable increase in audio strength with only a slight loss of naturalness. When the clipping is increased to about 10 db distortion becomes noticeable, while at 20–25 db some intelligibility is lost.

In Fig. 9.38 both positive and negative peaks are shown clipped (*symmetrical clipping*). This is by no means essential but is sometimes preferable. Either peak may be clipped (*asymmetrical clipping*) provided that the phasing of the audio chain is adjusted so that it is the *un*clipped peak that *increases* the carrier amplitude. Asymmetrical clipping results in the displacement of the effective zero line of the speech waveform. Normal intervalve couplings adjust themselves to give equal areas above and below the zero line, and asymmetrical clipping will upset this equality, giving a momentary shift in the zero axis. It is advisable to make all succeeding coupling capacitances small enough so that the time constant of the coupling circuits is short enough to allow rapid restoration of the normal zero position. It is quite possible for this momentary displacement of the zero axis, owing to the presence of a.c. intervalve couplings, to cause the intended level to be exceeded and therefore permit over-modulation in spite of the speech-clipping action. A similar effect can occur even in symmetrical clipping circuits owing to the asymmetrical characteristic sometimes experienced in speech waveforms. It may be obviated by ensuring that there are the fewest possible couplings after the clipper stage, i.e. the clipper should be as late as possible in the audio chain.

In principle speech clipping may be effected equally well at either high or low level. High-level clipping, as its name suggests, is introduced between the modulator and the modulated r.f. stage, while low-level clipping is introduced at some point preceding the modulator. The necessary filter for attenuating the higher audio-frequency components may also be inserted at either high or low level, but the filtering action must, of course, always take place *after* the clipping.

High-level clipping has the disadvantage that the associated circuit components must have a correspondingly high voltage rating, but on the other hand the filter would then also serve to remove any high audio frequencies that might be produced anywhere in the whole a.f. chain from the microphone input to the modulation transformer. Another disadvantage of high-level clipping is that the severe voltage transients which it causes may break down the insulation of the modulation transformer and possibly also damage the modulator valves, but to some extent these troubles could be encountered even without clipping in the event of extreme over-modulation. A reasonable factor of safety should therefore be allowed in the design of such equipment.

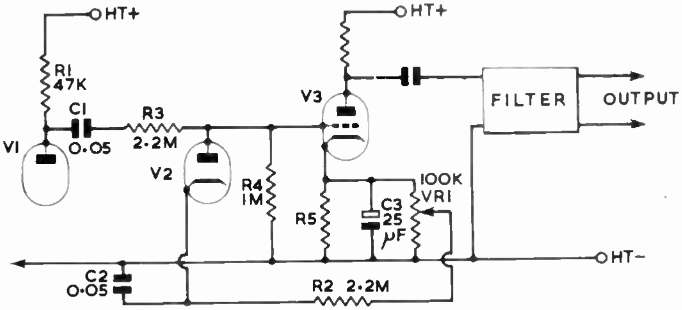


Fig. 9.41. Low-level asymmetrical shunt clipper.

Low-level Clippers

The limiting of the peaks that occur in speech waveforms resembles the limiting of impulsive interference by a conventional noise limiter. The circuits employed are very

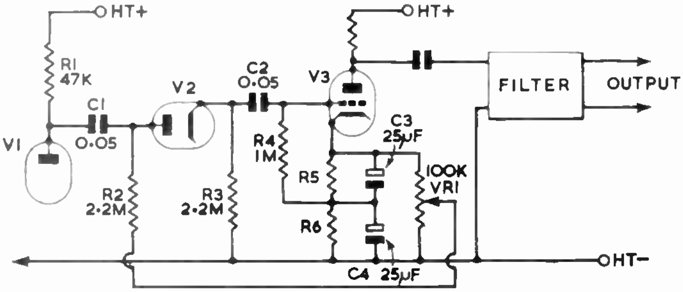


Fig. 9.40. Low-level asymmetrical series clipper.

similar and consist of a series or shunt arrangement of diodes suitably biased to give the desired clipping level. A typical low-level asymmetrical series clipper is shown in **Fig. 9.40.** If the anode of V_2 is held at, say, + 5 volts by means of potentiometer VR1 signals at all amplitudes up to 5 volts peak will be passed on unchanged, but any *negative* peak which exceeds − 5 volts will cause the diode to become non-conducting and the peaks will be clipped.

The corresponding shunt diode clipper is shown in **Fig. 9.41.** In this case the diode V_2 is normally non-conducting, but it conducts when the *positive* peak voltage at its anode exceeds the positive bias applied to the cathode from VR1 and thus all positive peaks which exceed this value are by-passed by the diode.

The corresponding symmetrical series and shunt clipping circuits are shown in **Figs. 9.42** and **9.43** respectively. The series circuit includes an extra pair of diodes to equalize the load on the previous stage. Whenever either of the series diodes stops conducting and in effect disconnects the following stage, its partner conducts and presents an equivalent dummy load.

The diodes in low-level clipping circuits may be double-diodes of the 6H6 or 6AL5 class (or their equivalent); miniature metal rectifiers or semi-conductor diodes may be used. The latter have three advantages; (a) no heater power is required, (b) a more compact layout is possible, and (c) there is no risk of hum being introduced into the clipper stage from the heater circuit.

The sharpness of the clipping action depends upon the characteristics of the diodes used. Thermionic diodes provided that the input is large enough, give almost perfect limiting, i.e. the clipped peaks are flat-topped. In the case of the smaller metal rectifiers, such as the copper-oxide instrument type, the limiting action is not so sharp and the output waveform resembles a "compressed" signal rather than a limited one. Assuming that the circuit is suitably adjusted to prevent over-modulation, this is not necessarily a condition to be avoided since it makes the succeeding filter somewhat simpler, the filter not then being required to remove such a wide range of undesired harmonics. A low-level clipper is sometimes combined with the first stage of the speech amplifier, the combination being a

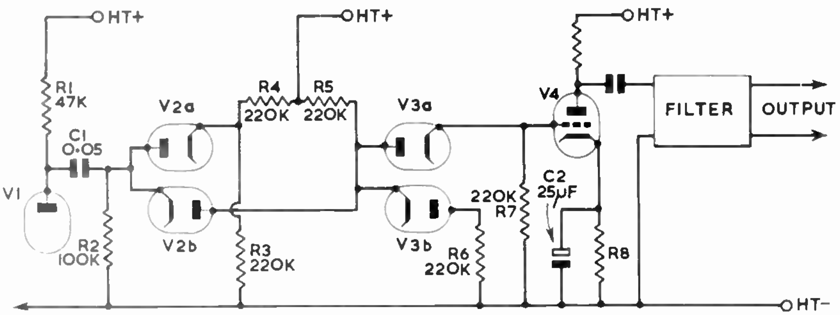


Fig. 9.42. Low-level symmetrical series clipper. This circuit provides a constant load for the preceding stage.

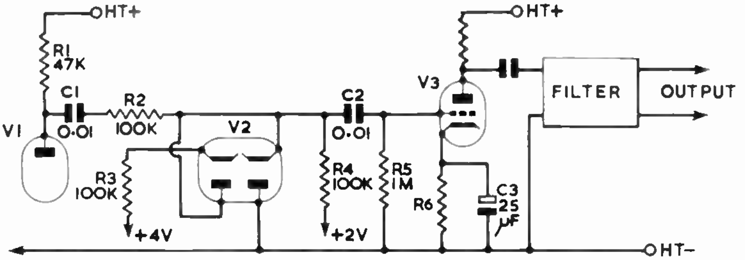


Fig. 9.43. Low-level symmetrical shunt-clipper.
The small positive bias voltages applied through R3 and R4 to the cathodes of the double-diode V2 can be conveniently obtained from taps on the cathode bias resistor R6 of the following amplifier stage V3.

separate unit from the rest of the audio equipment in the form of a pre-amplifier. A typical example of such an arrangement is shown in **Fig. 9.44.** In this circuit the 6AM6 and 6J6 are voltage amplifiers and the 6AL5 acts as a symmetrical shunt clipper.

An alternative form of clipping circuit is shown in **Fig. 9.46.** This uses a twin-triode valve in a cathode-coupled arrangement and has the advantage that it provides a certain amount of gain. It must be followed by the usual low-pass filter.

High-level Clippers

The high-level series-diode negative-peak clipper is the commonest arrangement. The circuit is shown in **Fig. 9.45.** The diode *V* conducts only when its anode is positive with respect to its cathode. Therefore the modulated anode voltage cannot swing the anode of the r.f. amplifier negative. For use in a 150-watt transmitter, the diode *V* may be a rectifier of the 500 volt 250 mA class with the two anodes connected together. It should be noted that the filament transformer *T* must have a primary-to-secondary insulation suitable for at least twice the anode voltage of the r.f. stage. Such a clipper used in conjunction with a good filter makes a very effective "splatter" suppressor. It has the advantage that relatively few extra components are required, although the filter components are likely to be rather bulky if entirely satisfactory performance is to be achieved.

Volume Compression

A *volume compressor* is an automatic gain control which causes the gain of an audio amplifier to vary inversely with the output level. In other words it reduces the gain when the level is high and increases it when the level is low. The volume range of the signal is automatically reduced or "compressed," and the intelligibility of a transmitted signal will therefore be increased by the consequent rise in the average level of the speech power.

Volume compression is often used for increasing the

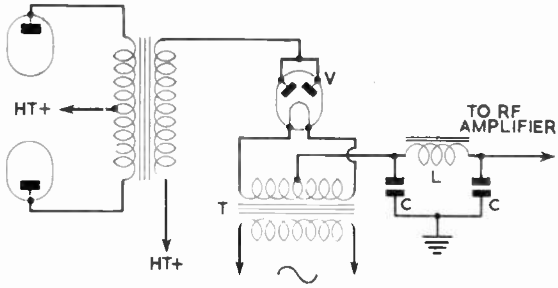


Fig. 9.45. High-level series-diode negative-peak clipper. The transformer T must be insulated for at least twice the anode voltage of the r.f. stage. V may be a 500 V, 250 mA rectifier. Suitable values of L and C can be calculated from the basic filter design equations (see text).

average depth of modulation in commercial radio telephony and also in sound recording, but its application in Amateur Radio is not as common as speech clipping.

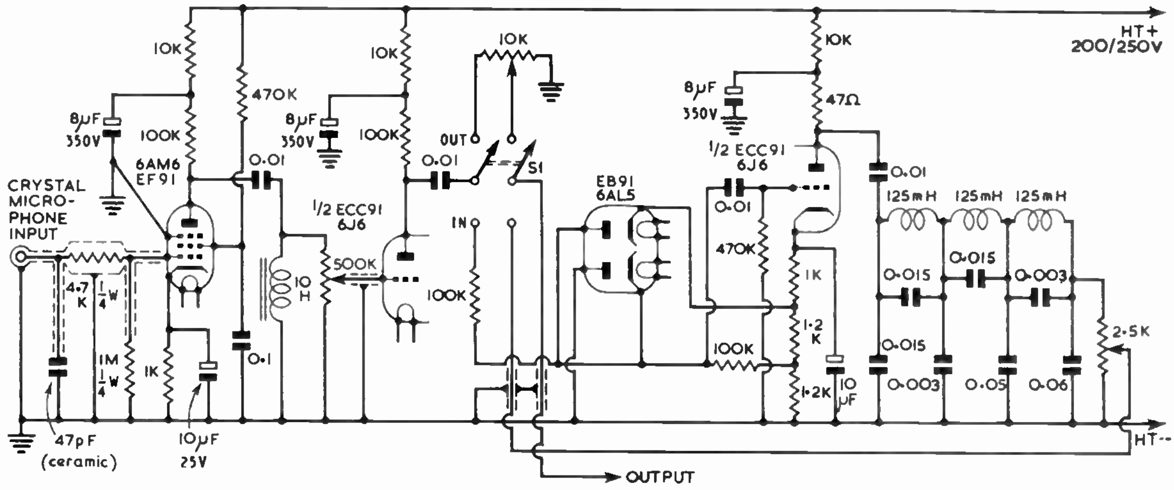
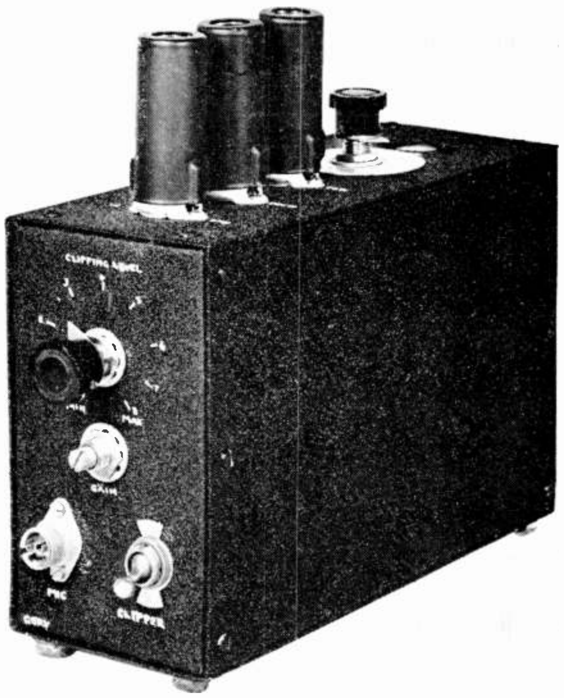


Fig. 9.44. Combined low-level symmetrical shunt clipper, speech amplifier and filter.



Speech amplifier clipper-filter unit, the circuit of which is shown in Fig. 9.44.

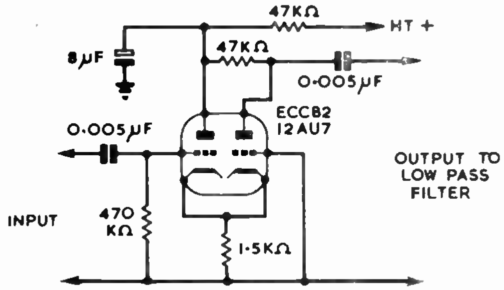


Fig. 9.46. Cathode-coupled triode limiter.

Such a circuit may be used to increase the average modulation depth of a telephony transmitter, without causing over-modulation. There is a practical limit to the amount of compression of the volume range because if it is carried too far any hum or background noise in the speech input will be increased to an unacceptable level. The increased gain at low input levels will also increase the risk of trouble due to overall feedback. The maximum amount of volume compression which can normally be used is that which will reduce gain by about 20db at peak level. As in the case of speech clipping, the received signal sounds louder because the average modulation power has been raised.

Volume compression is achieved by rectifying a portion of the output from the a.f. amplifier, passing the resultant rectified voltage through a smoothing circuit which eliminates the audio-frequency components, and then applying the smoothed output to the control grid or suppressor grid of one of the early voltage-amplifying stages in the chain. A simple circuit is shown in Fig. 9.47. A fraction of the output from the audio amplifier is applied through the transformer to the double-diode V_1. The resulting rectified voltage which is developed across R_3 is smoothed by the RC network composed of $C_2R_1C_3R_5C_4$. The smoothed voltage is then applied to the suppressor grid of V_2 which is the first stage of the speech amplifier. The voltage at which the circuit begins to operate is controlled by applying a positive bias to the cathodes of V_1 by the potentiometer R_2 which may be conveniently supplied through R_1 from the h.t line of the speech amplifier.

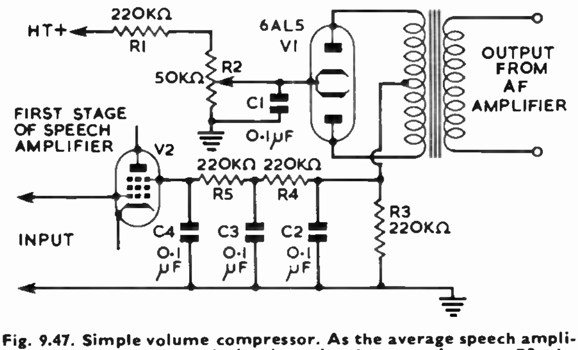


Fig. 9.47. Simple volume compressor. As the average speech amplitude rises above a certain level, set by the potentiometer R2, the gain of the amplifier V2 is reduced by the rise in the suppressor-grid bias.

The audio input to the compressor circuit may be obtained (a) directly from the modulator output, (b) from the input transformer of a class B output stage, or (c) from a separate amplifier valve fed from the last stage of the speech amplifier. Owing to the load presented by the rectifiers it is not usually satisfactory to feed such a circuit directly from a voltage-amplifier.

The distortion which results from volume compression is not so troublesome as that caused by the clipping of a signal, and therefore a low-pass filter is not essential after the variable-gain stage. Distortion of a less severe type can, however, be produced in the variable-gain stage, but this can be kept to a minimum by the use of two valves connected in push-pull which will cause even harmonics to be cancelled.

Clipping-v-Compression

A volume compressor circuit may not necessarily be as effective as a properly designed and adjusted speech clipper and filter system; however it is somewhat simpler in application because a low-pass filter is not required.

Since the object of clipping and compression circuits is to raise the average modulation level, it is important that the transmitter and its power supplies should be capable of operating at the higher average modulation level which will result. As already pointed out, while simple speech circuits without clippers operate at low average-signal conditions owing to the presence of the peaks, a speech transmitter with volume compression or speech-clipping circuits should preferably be designed as if it were required to operate with a sine-wave modulation signal. These are the conditions which are usually quoted in valve data.

It cannot be too strongly emphasized that the use of a speech clipper or a volume compressor is no cure for modulation troubles which may exist due to overloading of the modulator stage, bad impedance matching, etc.

FILTERS

Every speech-clipping stage must be followed by a filter to remove the harmonics generated by the action of speech clipping on the waveform. Such filters must obviously be put into the circuit after the clipping has taken place. Transmitters which only include volume compression do not introduce waveform distortion of the same type as clippers, but such transmitters, and even those without either clipping or compression can, with advantage, incorporate a low-pass filter to remove any modulation frequencies above about 3 kc/s, thus restricting the total transmitted bandwidth to about 6 kc/s.

The filter by itself will not prevent splatter resulting from over-modulation, but assuming over-modulation is avoided by adequate gain control of the amplifier stages, a filter to restrict the bandwidth is a worthwhile addition.

Filter Design

The complete design of filters is outside the scope of this Handbook, but the following information should enable the amateur to design and make satisfactory filters without very much trouble.

Before the design can be commenced, three factors must be known:

(a) The "cut-off" frequency, i.e. the frequency separating the regions of high attenuation and low attenuation.

(b) The amount of attenuation required in the attenuating band.

(c) The impedances between which the filter is to operate.

It is upon the first two of these factors that the degree of splatter-suppression achieved will depend. A recommendation of the IARU Congress (Paris), May 1950, is that the response of the modulator in an amateur telephony transmitter at 4 kc/s should be 26db below the response at 1 kc/s. A suggested method of achieving this is by the use of two filters, one in the speech amplifier with an attenuation of 20db at 4 kc/s and a second between the modulation transformer and the r.f. amplifier having an attenuation of 6db, the second filter having a higher cut-off frequency than the first. It would be reasonable to consider these attenuation figures as being the minimum acceptable, particularly in the case of a high-power transmitter.

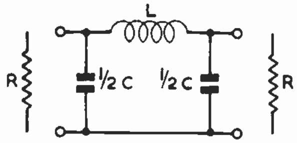


Fig. 9.48. Low-pass prototype or constant-k filter. It is also referred to as a pi-network filter. Its function is to attenuate all voltages having frequencies above a certain value determined by the filter components. See text for design equations.

The operating impedance of the filter is not critical, and normally considerable variation is possible. In the case of a high-level filter, it should be designed to work from the modulating impedance of the r.f. amplifier.

The simplest form of low-pass filter is the pi-network illustrated in Fig. 9.48. This is known as a "prototype" or "constant-k" filter. If the impedance of the two circuits which are coupled by the filter are assumed equal, the design equations are as follows:

$$R = \sqrt{\frac{1000\,L}{C}} \qquad L = \frac{R}{\pi f} \qquad C = \frac{1000}{\pi f R}$$

where R = terminal impedance (ohms)
L = inductance in filter (mH)
C = total capacitance in filter (μF)
f = cut-off frequency (kc/s)

Thus L and C may be calculated for any chosen values of R and f. In a low-pass filter for the purpose described f could have a value of 3·3 kc/s.

Such a filter by itself has an attenuation curve which does not rise very steeply beyond the cut-off frequency,

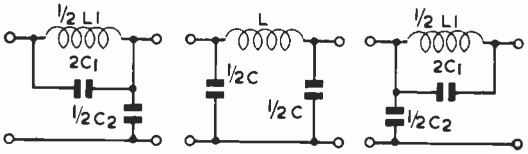


Fig. 9.49. An improved low-pass filter having a more sharply-defined cut-off frequency than the simple filter shown in Fig. 9.48. The diagram illustrates the three separate sections comprising a single-section "prototype" filter preceded and followed by an "m-derived" section. See text for design equations.

and would only give an attenuation of about 6db at 4 kc/s and about 30db at 10 kc/s. It would, however, be suitable for a high-level filter if a filter of higher attenuation is included in the speech amplifier.

The simplest method of increasing the attenuation of a filter is to connect several identical sections in cascade. Three sections of the type just referred to give an attenuation of about 20db at 4 kc/s and about 60db at 10 kc/s. Thus a three-section constant-k filter in the low-level amplifier plus a single section of similar type for the high-level filter would just about meet the specified requirement. If a steeper attenuation/frequency curve is desired, use may be made of what are called "m-derived" half-sections. The arrangement of a filter having a single constant-k section and an m-derived half section at each end is shown in Fig. 9.49. The design equations for the m-derived half sections at the ends are:

$$L_1 = mL \qquad C_1 = \frac{1 - m^2}{4m}\,C \qquad C_2 = mC$$

The general attenuation characteristic of such section has a very sharp peak of attenuation followed by a reduced attenuation remote from the cut-off frequency, as compared with the constant-k sections which have a slow but steadily increasing attenuation. The quantity m in the above equations is determined by the separation between the frequency of peak attenuation and the cut-off frequency. A convenient value of m for design purposes can be taken as 0·6, and these equations then become:

$$L_1 = 0{\cdot}6L \qquad C_1 = 0{\cdot}27C \qquad C_2 = 0{\cdot}6C$$

A filter consisting of a single constant-k section and two m-derived half-sections would give an attenuation of about 30db at 4 kc/s.

It will be seen from the above equations that the value

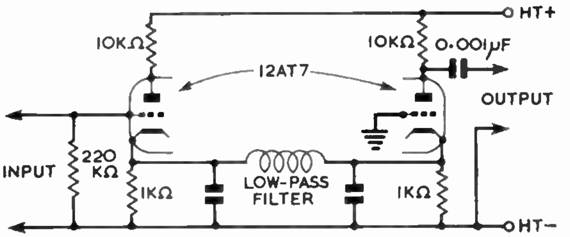


Fig. 9.50. A suggested circuit arrangement for a low-impedance type of low-pass filter. The use of a cathode follower and a grounded-grid stage provides relatively low terminal impedances. The values of the filter components can be calculated from the basic filter design equations (see text).

of inductance required is proportional to the terminal impedance of the filter. High values of inductance introduce difficulties owing to the presence of self-capacitance, and for this reason it is often preferable to avoid filters of higher impedance than, say, 10,000–15,000 ohms. Such filters, however, may not provide adequate load impedance for voltage amplifiers, and a means of using a low-pass filter with a terminal impedance as low as 500–1000 ohms is shown in **Fig. 9.50.** The filter is connected between a cathode follower and a grounded-grid stage.

Filter Construction

As the values of *L* and *C* required in a filter depend upon the choice of impedance and the cut-off frequency, these values may be selected in order to utilize existing components. Obviously the use of close-tolerance components enables the filter to fit the exact requirements, but this is sometimes rather a luxury in amateur practice.

The layout of filters is not usually very critical. However, care should be taken to avoid undesired couplings between filter coils by suitable spacing. The mounting of such coils with their axes at right-angles will reduce the coupling between them, provided that the axis of one coil intersects the axis of another at its centre.

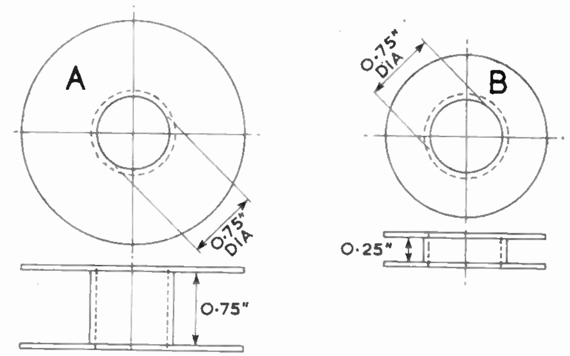


Fig. 9.51. Dimensions of bobbins for inductances wound according to Table 9.8. The inductances are all air-cored.

Capacitors. The capacitance values required rarely coincide with those easily obtainable, and series or parallel arrangements are usually necessary. Good quality paper capacitors are satisfactory, although mica ones are better, but if these are not readily available the extra expense involved is rarely justified, provided that the working voltage of the capacitors is adequate. In a high-level filter the voltage rating of the capacitors should be at least twice the anode voltage of the r.f. amplifier.

Inductances. In the case of a high-level filter, the inductances must be capable of carrying the h.t. current of the modulated stage, and they therefore tend to be of larger dimensions than those intended for low-level filters where the current may be only a few milliamperes.

Either air-cored or iron-cored inductances may be used, but air-cored ones are simpler to design and make. This applies particularly to inductances for high-level filters.

Inductances with an accuracy of 5–10 per cent may be made by reference to Table 9.8 which shows the number of turns of wire required to give the stated values of inductance when wound on one or other of the two bobbins illustrated in **Fig. 9.51.** For the larger bobbin (A) the wire size is 32 s.w.g. (enamelled), and inductances wound with this wire will carry 200 mA without overheating, thus being suitable for high-level filters. The largest coil listed (1 Henry) will take about 1 lb. of wire and will have a resistance of about 260 ohms.

For the smaller bobbin (B) the wire size required is 40 s.w.g. (enamelled), and the largest size shown will need about 2¼ oz. of wire and will have a resistance of about 850 ohms. These inductances are suitable for low-level filters. In either case inductances of intermediate values may be found by simple interpolation.

TABLE 9.8

Winding Data for Air-cored Inductances

Approximate Inductance (mH)	Number of Turns	Outside Diameter	
		Bobbin A	Bobbin B
80 100 120 150	2030 2240 2400 2620	2½ in.	1¼ in.
180 200 250 300	2800 2930 3200 3450	3 in.	1½ in.
350 400 450 500 600 700	3660 3850 4030 4200 4570 4840	3½ in.	1¾ in.
800 900	5160 5480	4 in.	2 in.
1000	5800	4 in.	2⅛ in.

For bobbin A the wire size is 32 s.w.g. enamelled.
For bobbin B the wire size is 40 s.w.g. enamelled.

Typical A.F. Filters

The filter shown in **Fig. 9.52** is a single-section constant-*k* filter suitable for use at high level, feeding an r.f. amplifier having an input of 150 watts. Assuming an anode supply of 1000 volts and a current of 150 mA, the terminal impedance required is 1000 volts/0·15 amp, which equals 6700 ohms. The values of the components shown will give acut-off frequency of 3·3 kc/s.

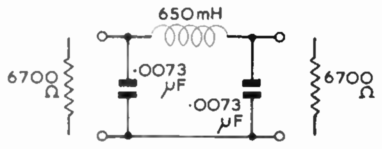


Fig. 9.52. Single-section "prototype" filter suitable for feeding a 150 watt modulated r.f. stage.

Fig. 9.53 shows a filter having a single constant-*k* section with *m*-derived end sections which would be suitable for use at low levels in a voltage amplifier. It is designed for

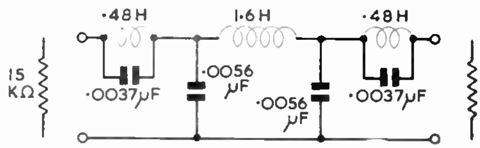


Fig. 9.53. Low-pass filter suitable for use at low levels in a voltage amplifier having an output impedance of 15,000 ohms. The cut-off frequency is 3 kc/s. The filter comprises a "constant-*k*" section with an "*m*-derived" section at each end.

a terminal impedance of 15,000 ohms and has a cut-off frequency of 3 kc/s.

Resistance-capacitance Filters

By careful design, a filter of the type just described may be made to have a characteristic which is flat to 2·5 kc/s or so and then falls rapidly to the order of – 50db at 4 kc/s.

It is considered by some that such a filter gives a too unnatural sound to the voice and that a filter with a much slower fall-off in its characteristic is preferable.

Such a characteristic may be obtained from a resistance capacitance filter. A single section *RC* filter may be considered as a potentiometer composed of the resistance and the reactance of the capacitor. The ratio of the potentiometer therefore, is dependent on the frequency of the input voltage.

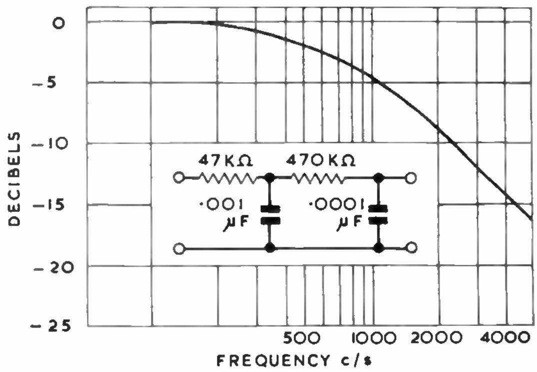


Fig. 9.54. Typical two stage RC filter and approximate Loss/Frequency characteristic.

A two-section *RC* filter is shown in **Fig. 9.54** which also shows its approximate loss/frequency characteristic.

The attenuation per octave may be roughly doubled by the combination of such a filter with a cathode follower in the arrangement shown in **Fig. 9.55.**

This type of filter is obviously not as effective as the conventional *LC* type, particularly when used after heavy speech clipping, but it has the merit of simplicity.

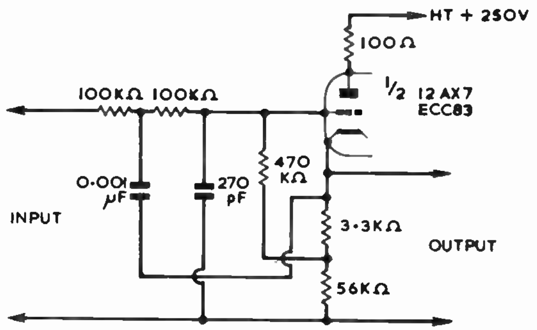


Fig. 9.55. An RC low-pass filter combined with a cathode follower. This arrangement has an attenuation of 10-12 db per octave.

ADJUSTMENT AND MONITORING OF AMPLITUDE MODULATED TRANSMITTERS

The importance of thoroughly checking the overall performance of a telephony transmitter cannot be overemphasized. Unfortunately it is a task which can present considerable difficulties, especially in the case of a transmitter with a power input as high as 150 watts. Nevertheless, it should be the aim of every amateur to test his modulator as far as the facilities available will permit before applying its output to the transmitter, and also to test the complete system with some form of dummy load before using it to radiate a signal.

The principal tests which should be applied to any telephony system are as follows:

(a) *On the modulator:*
- (i) maximum undistorted power output,
- (ii) frequency distortion (i.e. variation of gain over the required frequency range),
- (iii) amplitude distortion (i.e. introduction of harmonics and other unwanted frequencies),
- (iv) hum level.

(b) *On the complete system:*
- (i) depth of modulation,
- (ii) overall fidelity.

Testing the Modulator

After the wiring of the modulator has been thoroughly checked the heaters should be switched on, and after a few minutes the h.t. may be applied. The voltage on the various electrodes of each valve should be measured with a high-resistance voltmeter having a sensitivity of at least 10,000 ohms-per-volt, and the results can then be compared with the design figures.

In order to avoid possible damage to some of the components the h.t. voltage should on no account be applied to the output stage unless it has a suitable load. This should consist of a non-inductive resistor having a dissipation rating at least equal to the expected power output of the modulator. It should have a resistance approximately equal to the modulating impedance of the r.f. stage, or alternatively of such a value that it can be matched to the modulation by suitable adjustment of the taps of a multi-ratio modulation transformer.

The ideal set-up for the overall performance checking of any audio amplifier is shown in **Fig. 9.56.** A signal of known frequency and known voltage is fed into the amplifier from an a.f. oscillator. The amplifier output is dissipated in the resistive load *R*, the voltage across the load being measured by the voltmeter *V*. This must be an a.c. voltmeter suitable for use over the audio frequency range. Many multi-range test meters are suitable for this purpose. A preliminary calculation to find the approximate value of the output voltage is advisable in order to set the range correctly and avoid damaging the voltmeter. It will be found convenient to plot a graph of the calculated power output against the voltmeter reading for the load resistance selected, so that the actual output power from the amplifier can be determined under the various conditions of test.

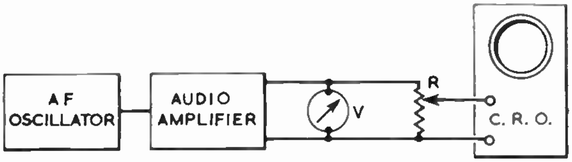


Fig. 9.56. Ideal arrangement using a test oscillator and an oscilloscope for checking the overall performance of an audio amplifier.

The output voltage is observed on the cathode-ray oscilloscope the input to which may be tapped down the resistance load if necessary, to avoid overloading the oscilloscope. The maximum undistorted power output at any frequency is readily found by increasing the gain control in the amplifier, or by increasing the output of the oscillator until distortion becomes apparent on the output waveform. It is advisable to check the output of the oscillator with the oscilloscope to be certain that there is no inherent distortion of the input waveform. Such distortion tests should be made at low and high frequencies as well as at medium frequencies. With the gain control of the modulating system left at the same setting, the input signal should be removed and the residual hum measured on the voltmeter and observed on the oscilloscope. The hum voltage should preferably be less than 1 per cent of the maximum undistorted signal output voltage. If it is more than this the origin of the hum may be traced by applying the oscilloscope to the output of each stage in turn.

By varying the frequency of the oscillator and measuring the modulator output voltage, keeping the input voltage constant, the frequency response of the modulator can be determined. Any low-pass filters which it is intended to use should be included in the equipment when such tests are made.

Another method of estimating distortion is to apply the input and output voltages of the modulator to the horizontal- and vertical-deflector plates respectively of the oscilloscope, with the internal time base inoperative. The gain in the horizontal- and vertical-deflection amplifiers should be adjusted so that vertical and horizontal lines of equal length are obtained when each is tested separately. When both inputs are applied simultaneously, the resulting pattern will be a straight line inclined at 45 degrees to the horizontal and vertical axes. Any phase shift originating in the amplifier, which will be most noticeable at the lower and upper frequency limits, will cause this straight line to open out into an ellipse. Distortion produced in the modulator will show up as a curving at the ends of the straight line or at the tips of the ellipse. Examples of the patterns which may be observed in this way are shown in **Fig. 9.57.**

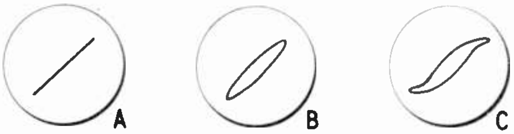


Fig. 9.57. Oscilloscope patterns produced by applying the input and output voltages of the modulator to the X- and Y-plates respectively. The straight-line trace A shows the complete absence of phase-shift or distortion. The simple loop shown at B indicates slight phase-shift but no distortion. The twisted loop C indicates both phase-shift and distortion.

Before beginning such tests it is advisable to apply the oscillator voltage to both of the oscilloscope amplifiers to verify that there is no appreciable distortion in these amplifiers.

Simpler methods are available when an oscilloscope and audio oscillator are not available. The output of the modulator can be monitored by means of a pair of headphones connected across a low resistance in series with the load resistance, as shown in **Fig. 9.58.** The value of *r* should be adjusted to give a comfortable signal strength in the headphones; generally 2–3 ohms is sufficient. As a safety precaution it is very important that one side of the headphones should be earthed, and that the resistor *r* should be of a very reliable type and very securely connected across the headphones. Such a simple testing method will enable any serious distortion to be heard when the modulator is driven either from the microphone to be used or from a musical signal derived from a broadcast receiver or a suitable recording.

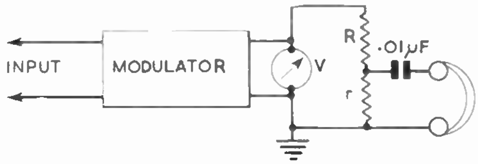


Fig. 9.58. A simple method of checking the audio performance of the modulator by a direct listening test. Special care must be taken to ensure that no dangerously high voltage ever reaches the headphones.

Testing the Complete System

Before the output of the modulator is applied to the transmitter, it is essential to make certain that the r.f. amplifier which is to be modulated is itself operating satisfactorily. Neutralization, if used, should be as accurate as possible. The correct values of grid bias and grid drive should be applied, and the aerial coupling should be adjusted so that the stage is drawing the required anode current. This ensures that the modulating impedance of this stage is of the correct value. As suggested earlier, the first trials of any telephony transmitter should be carried out with some form of dummy aerial to avoid causing unnecessary interference to other stations.

When it is certain that the r.f. stage is operating satisfactorily, the modulation may be applied and steadily increased up to the maximum amplitude which the system in use is capable of accepting. Almost 100 per cent should be possible in the case of anode modulation, but 75 per cent is as much as can be expected without distortion for the various forms of grid modulation.

Measurement of Modulation Depth

The cathode-ray oscilloscope is the best instrument for measuring the depth of modulation since it enables a quick visual check to be made. There are two distinct methods

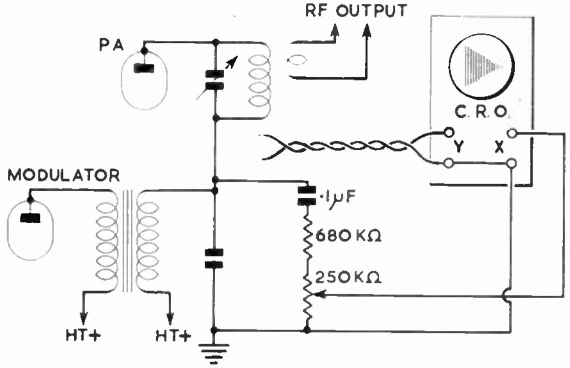


Fig. 9.59. Trapezium test for checking the modulated output from the r.f. power amplifier. Here r.f. voltage is applied to the *Y*-plates and a.f. voltage is applied to the *X*-plates. No timebase is required in the oscilloscope.

of measuring modulation depth with an oscilloscope; one is dependent on the use of a timebase and the other is not.

The method which does not require a timebase is illustrated in **Fig. 9.59.** A small fraction of the modulated output voltage is fed to the *Y*-plates of the oscilloscope by a pick-up loop loosely coupled to the output tank circuit while the a.f. modulating voltage is fed to the *X*-plates. With a steady modulating signal a trapezoidal or wedge-shaped pattern will be produced on the screen. **Fig. 9.60** shows the variations in this pattern for various degrees of modulation. The modulation depth is calculated as follows:

$$m = \frac{P - Q}{P + Q} \times 100 \text{ per cent}$$

where P and Q are the lengths of the vertical sides of the pattern. P and Q can be measured with reasonable accuracy by means of a pair of dividers. For 100 per cent modulation a triangular pattern is produced.

The oscilloscope pattern can also be used to examine the overall performance in other respects besides the actual modulation depth. If the operating conditions are correct so that there is no distortion in the modulation process, the sloping sides of the trapezoidal pattern will be perfectly

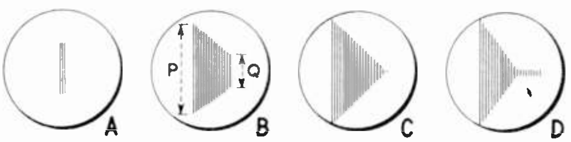


Fig. 9.60. Typical oscilloscope patterns obtained by the method shown in Fig. 9.59. The vertical line A shows the unmodulated-carrier amplitude. In B the carrier is modulated 50 per cent while in C it is modulated 100 per cent. Where the sloping edges of the pattern are flattened, as in D, the carrier is over-modulated.

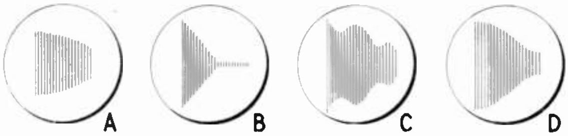


Fig. 9.61. Fault conditions in anode modulation: A—insufficient drive: B—over-modulation: C—instability in power amplifier: D—incorrect matching of modulator to p.a.

straight: any curvature of these sides indicates non-linear modulation. Typical patterns for various fault conditions in both anode- and grid-modulation systems are shown in **Figs. 9.61** and **9.62** respectively. The audio voltage applied to the *X*-plates may be derived directly from the audio input to the amplifying system of the transmitter instead of from the output of the modulator, in which case any distortion in the audio part of the system will become apparent. The apparatus required for this test is very simple and consists only of a small cathode-ray tube and its associated power supply. The whole apparatus can be conveniently built up on a 3-in. panel for rack mounting.

Details of a modulation monitor will be found in Chapter 19 (*Measurements*).

The alternative method, which requires the use of a timebase, produces the envelope of the modulated wave on the screen. In the arrangement shown in **Fig. 9.63** the modulated r.f. voltage is applied to the *Y*-plates as in the previous method, and the normal timebase which is of course applied to the *X*-plates is adjusted so that the audio-frequency

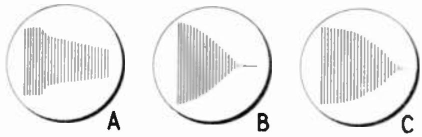


Fig. 9.62. Fault conditions in grid modulation. A—excessive drive: B—over-modulation: C—power amplifier insufficiently loaded.

components of the envelope are suitably displayed. A rectangular patch of constant height will be observed when there is no modulation, and this will show the appropriate change of shape during modulation. The type of patterns displayed on the cathode-ray tube are shown in **Fig. 9.64.** By measuring the height R corresponding to a modulation peak and the height S of the unmodulated carrier, the depth of modulation can be calculated directly:

$$m = \frac{R - S}{S} \times 100 \text{ per cent}$$

The patterns resulting from incorrect operation of the transmitter are more difficult to interpret in this method, and for this reason the trapezium test is usually found more satisfactory.

Whichever method is used the results are easier to interpret when a constant sinusoidal input is applied to the modulator, but for many purposes the signal obtained from a sustained whistling into the microphone is sufficient.

In the absence of a cathode-ray oscilloscope the most convenient way to determine the depth of modulation is to measure the aerial current with and without modulation. A rearrangement of the formula shown on page 9.2 gives the expression for modulation percentage with a sinusoidal signal as:

$$m = \sqrt{2\left[\left(\frac{I_m}{I_0}\right)^2 - 1\right]} \times 100 \text{ per cent}$$

This method of measurement assumes that there is no distortion in the modulating system. If such distortion is present it may even cause the aerial current to decrease on modulation, and unreliable results are frequently obtained when using this method. Moreover the formula does not apply to modulation by speech. For these reasons the first method of using the cathode-ray oscilloscope is much to be preferred.

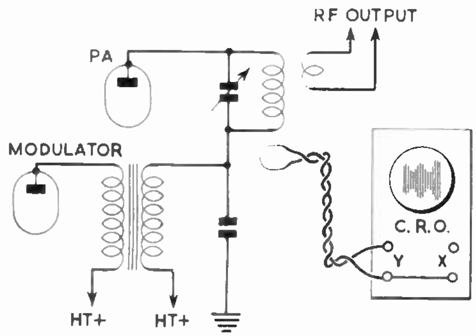


Fig. 9.63. Alternative methods for checking the modulated output from the r.f. power amplifier. The test signal is applied to the Y-plates, while the X-plates are supplied with a linear timebase voltage of any suitable frequency.

In an anode-modulation system the modulation percentage may be obtained approximately by measuring the a.c. voltage across the modulation choke or the secondary of the modulation transformer. If the test signal has a sinusoidal waveform, the modulation percentage is given by:

$$m = \frac{\text{a.c. modulation voltage (r.m.s.)}}{\text{d.c. voltage applied to p.a.}} \times 141 \text{ per cent}$$

If, however, the test signal is not sinusoidal or if for any reason it is more convenient to measure the *peak* value of

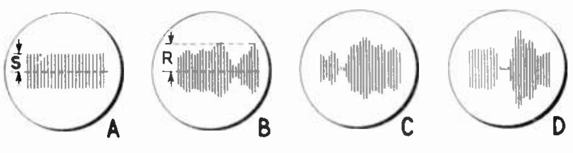


Fig. 9.64. Typical oscilloscope patterns obtained by the method shown in Fig. 9.63. A—unmodulated carrier: B—50 per cent modulation: C—100 per cent modulation: D—over-modulation. Since the time-base frequency is independently fixed the patterns obtained with a normal speech input are constantly varying in position on the horizontal axis.

the modulation voltage instead of the *r.m.s.* value, the factor 141 in this formula must be reduced to 100.

The formulae given here are only applicable to the use of a sine-wave modulation voltage. In the case of speech the rise in average aerial current from the measured modulation voltage for 100 per cent modulation will be considerably less.

Downward Modulation

Downward modulation is the name given to a condition, sometimes occurring in amateur telephony transmitters, in which the power level of the modulated carrier is found to *diminish* as the modulation amplitude increases. It is easily detectable by connecting a small lamp in the aerial lead or coupling it to the p.a. tank coil; if downward modulation is occurring the lamp will be seen to decrease in brilliance as the modulation level is raised.

This spurious condition is somewhat similar to the "under-modulation" phenomenon in v.h.f. transmitters, and is caused only by incorrect adjustment or bad design. The main features to be suspected are as follows:

(a) Incorrect grid drive and/or bias on the p.a. stage,
(b) Power-amplifier stage not being run under the valve maker's recommended conditions, or alternatively lack of cathode emission,
(c) Presence of parasitic oscillations in the p.a. stage,
(d) Incomplete neutralization in the case of triode p.a.,
(e) In the case of tetrode or pentode p.a., incorrect time-constant of the screen circuit; the value of the screen decoupling capacitor should be such as to give a suitably high reactance (e.g. higher than the screen dropping resistor) at the highest modulation frequency in use. If too great a time-constant is used, the screen potential will not be able to "follow" the modulating voltage; generally this implies that the screen decoupling capacitor should not be greater than $0{\cdot}002\mu F$,
(f) Grossly incorrect matching between the output stage and aerial due to incorrect design or adjustment of the pi-coupler,
(g) Inadequate regulation of the h.t. supply to the p.a. stage,
(h) Modulation transformer or modulator being grossly over-run,
(i) Aerial coupling too tight.

The remedy for all these defects is fairly obvious. In this respect, the value of checking the linearity of any modulated amplifier by means of an oscilloscope cannot be over-emphasized.

Modulation Monitors and Over-Modulation Indicators

Even if no means of measuring the exact depth of modulation is available, some indication of the presence of over-modulation is essential. There are several ways of doing this.

In a class B or class AB2 modulator the anode current of the modulator varies appreciably with the amplitude of the signal input. If a device to measure the actual modulation percentage can be borrowed a mark can be made on the scale of the modulator anode-current meter corresponding to 100 per cent modulation, so that the gain control can always be adjusted to ensure that this value is not exceeded. Similarly a high-resistance a.c. voltmeter connected to the output of the modulator can be calibrated to show percentage modulation from the formula shown above. It should be noted, however, that sluggishness in the operation of such a meter may give rise to a false impression.

Since the average anode current of an anode-modulated amplifier is constant, any variation in the reading of the anode milliammeter may be taken to indicate over-modulation provided that the modulating system is free from distortion such as that due to insufficient grid drive or grid bias on the modulated stage which will also cause the meter needle to flicker. Any instability of the r.f. stage can likewise cause variations in the anode current, and consequently the results of such a test are not always easy to interpret.

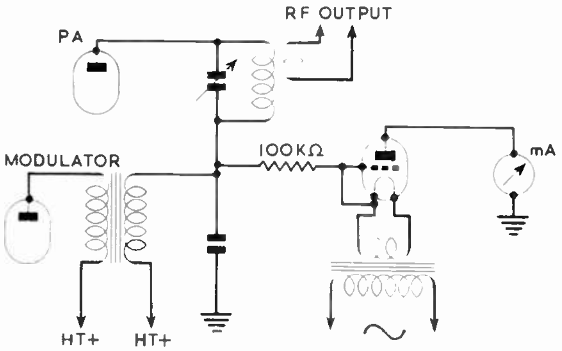


Fig. 9.65. Over-modulation indicator. The meter needle will flicker when the depth of modulation exceeds 100 per cent.

The sudden interruptions of the carrier which occur in over-modulation as a result of the reduction of the p.a. anode current to zero are the principal cause of interference to local receivers, and it is therefore particularly important that this should be avoided. A simple means of showing the presence of this effect is shown in **Fig. 9.65.** The milliammeter in series with the rectifier will show a reading whenever there is a peak of modulation which carries the instantaneous voltage on the anode of the modulated r.f. stage below zero into the negative region. The diode cannot conduct as long as the negative peak of the audio output voltage is less than the d.c. voltage applied

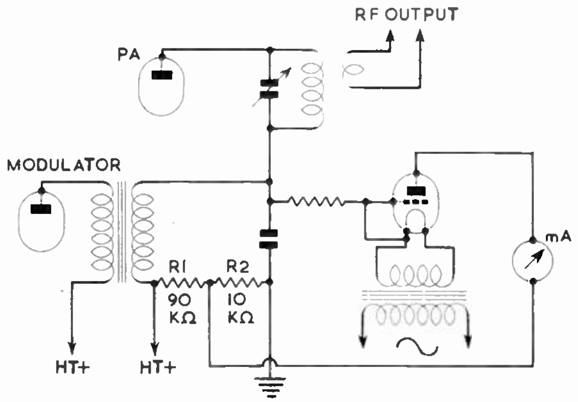


Fig. 9.66. Over-modulation indicator biased to show when the depth of modulation exceeds 90 per cent. The critical percentage can be altered by suitably proportioning the potential divider R1-R2.

to the r.f. stage. Both the rectifier and its heater transformer must be capable of withstanding a voltage equal to the sum of the h.t. and peak a.f. voltages. Any high-voltage low-current rectifier or an old transmitting triode with the grid and cathode strapped together will serve as the diode. A low-reading milliammeter (e.g. one with a full-scale deflection of about 5 mA) is a suitable indicator.

It should be noted that if a mercury-vapour rectifier valve is used, the glow of the mercury vapour when the valve passes current is an indication of the occurrence of over-modulation. If such a valve is used, it is therefore worthwhile mounting it so that it can be seen from the operating position.

If an indication of any modulation depth less than 100 per cent is required, the meter may be returned to a point on a potentiometer connected across the h.t. supply instead of to earth. For example, if a maximum of 90 per cent is to be indicated, the voltage across the earthed section of the potentiometer should be 10 per cent of the d.c. voltage. Such an arrangement is illustrated in **Fig. 9.66.**

Taking everything into account, the use of a speech-clipping system which will automatically avoid over-modulation is to be recommended in preference to a non-limited system which has to be continuously monitored to avoid over-modulation. Nevertheless the monitoring of the speech quality by means of headphones, using the station receiver (generally referred to as *side-tone*), provides a valuable overall check. It is necessary, however that the receiver should have a gain control in the first r.f. stage and that it should be well screened in order to prevent overloading and the resultant distortion of the signal. Such a monitoring receiver cannot be considered satisfactory unless it is possible to reduce the r.f. gain so that the transmission can be rendered completely inaudible.

FREQUENCY MODULATION SYSTEMS

Frequency-modulated transmitters may be divided into two broad classes:

(a) Those which use a quartz crystal to determine the centre frequency.

(b) Those which operate by the direct frequency variation of the master oscillator.

Transmitters in the first category are, in general, complicated in design and may involve the production of a phase-modulated signal which is then converted to a frequency-modulated one, the use of specially cut crystals or complicated chains of phase shifters, balanced modulators, frequency multipliers and mixers. However, they are capable of providing large deviations at extremely high degrees of stability of the centre frequency. For details, the reader is referred to the standard works on frequency modulation.

In the second category, the frequency of the master oscillator or v.f.o. is varied directly by placing a variable reactance across the tuned circuit of the oscillator. This of course implies that the frequency stability of the oscillator itself must be as high as possible. The stability of the present-day variable-frequency oscillator as used in amateur transmitters is usually adequate and this method of producing frequency modulation is therefore the one best suited to amateur use.

The Variable-Reactance Valve Modulator

The basic circuit of the variable-reactance or reactor type of modulator is shown in **Fig. 9.67.** Essentially it consists of a valve connected across the tuned circuit of an oscillator in such a way that it behaves like a variable capacitance or inductance. In the circuit shown, a potentiometer consisting of a capacitor C_2 and a resistor R in series is connected from anode to earth and also, of course, across the tuned circuit.

The centre point is taken to the grid of the valve. Provided that the resistance of R is large compared with the reactance of C_2 at the resonant frequency of the tuned circuit, the grid will be fed with a voltage which is very nearly 90° out of phase with the voltage at the anode. A similar phase-shift but of the opposite sign may also be obtained by interchanging C_2 and R, in which case the reactance of C_2 would need to be large compared with the resistor R.

Since the anode current of a valve is in phase with its grid voltage, the current flowing through the reactance valve is 90° out of phase with the voltage across the tuned circuit due to the 90° phase-shift introduced by the combination of C_2 and R. Since the valve current also flows through the tuned circuit, the current through the tuned circuit is 90° out of phase with the voltage across it, which is equivalent to the result produced by connecting a reactance across the tuned circuit. This reactance may be either capacitive or inductive according to the configuration of the RC_2 potentiometer. In the arrangement shown in Fig. 9.67, the reactance will be inductive, while if R and C_2 are interchanged, the reactance will be capacitive. In either case, the effect of this reactance will be to change the resonant frequency of the tuned circuit.

The value of the reactance thrown across the tuned circuit and hence the change in resonant frequency will obviously depend of the value of the anode current of the valve;

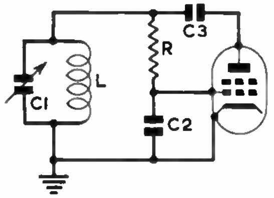


Fig. 9.67. Basic circuit of a variable-reactance frequency modulator. L-C1 is the tuned circuit of the master oscillator.

in other words it will depend on the voltage applied to the grid of the valve. Thus the variable reactance valve presents a simple means of varying the frequency of an oscillator in sympathy with an a.f. voltage applied to the grid of the reactance valve. The amplitude of the grid voltage will govern the magnitude of the change in frequency; i.e. the deviation produced will be determined by the amplitude of the modulating signal and the rate at which the oscillator frequency is varied will be equal to the frequency of the voltage applied.

A reactance-valve modulator may be used with any type of oscillator. It is normally connected directly across the tuned circuit of the oscillator as shown in Fig. 9.67, although in the case of a series-tuned circuit, such as the Clapp oscillator, it may be connected across the tuning capacitor.

Almost any tetrode or pentode valve will function as a reactance in this manner and since the valve is a voltage-operated device, a voltage-amplifying type is required rather than a power type (the long grid base of the variable-μ valve is required for optimum performance). Alternatively, a hexode or similar multi-grid valve may be used, and the r.f. voltage from the reactance potentiometer and the a.f. voltage from the speech amplifier can then be applied to different grids. In this way, there is less risk of instability which could be caused by feedback between the r.f. and a.f. circuits. It will be realized from the foregoing account of the operation

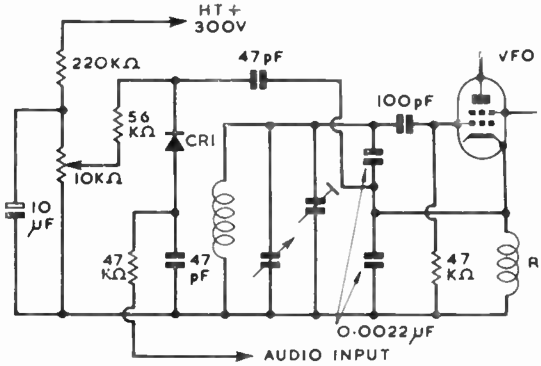


Fig. 9.68. Variable capacitance diode frequency modulator. CR1 may be type SX761, SVC1 or similar.

of reactance modulators that the sensitivity of the reactor stage will be dependent on the mutual conductance of the reactance valve, i.e. the higher the mutual conductance the greater will be the change in anode current and hence in deviation for a given output voltage to the grid. The valve must therefore be chosen with regard to the degree of a.f. amplification available or intended. In the interests of overall stability and ease of adjustment a valve having low or medium mutual conductance is generally to be preferred.

In the potentiometer arrangement shown in Fig. 9.67, the capacitor C_2 may conveniently consist of the input capacitance of the reactance valve plus the stray capacitance at this point, although of course no adjustment of C_2 is then possible and the control must be effected by varying the resistor R. The capacitor C_3 is merely a d.c. blocking capacitor.

The h.t. supply to the reactor and associated circuits, i.e. the speech amplifier and the v.f.o. must be very well smoothed to prevent the production of spurious amplitude and frequency modulation at the h.t. ripple frequency and also any unwanted change of mutual conductance of the reactance valve. Likewise great care must be taken with the layout to prevent hum pick-up.

The Variable Capacitance Diode Frequency Modulator

The variable capacitance diode is a small area *p-n* junction diode which exhibits a change in junction capacitance with change of reverse voltage. This change in capacitance is due to the change in thickness of the "depletion layer" or "charge-free" zone between the *p* region and the *n* region as the reverse voltage is varied. The capacitance varies very approximately as the cube root of the reverse voltage applied to the diode.

Such a device connected across a tuned circuit therefore presents a very convenient method of changing the resonant frequency of the tuned circuit by variation of a voltage. If the voltage applied to the diode is varied at audio frequency, frequency modulation of an oscillator whose frequency is controlled by the tuned circuit may be achieved.

All *p-n* diodes have this characteristic and most may be used for this particular application. Special diodes are also available in which the effect is closely controlled in manufacture.

A typical circuit is shown in Fig. 9.68. As the capacitance/voltage characteristic of these devices is not linear, a small standing bias of the order of 1 volt to 1·5 volts is applied. This is adjusted by the 10K ohms potentiometer to give an approximately equal change in capacitance for equal changes of applied voltage in each direction, i.e. for each sideband. The change in capacitance for a 1 volt change in bias is of the order of 2·5pF. The audio modulating voltage is applied " on top of " the standing bias. The peak modulating voltage must be restricted so that it does not drive the diode into the positive region, i.e. it must not cause the diode to conduct. It should be noted that the diode must be connected so that the bias applied is in fact a " reverse " voltage.

In setting up this circuit, it is advisable to follow the procedure laid down for adjustment of the reactance valve modulator described later in this chapter.

Multi-band Operation

It is necessary to emphasize here a point made earlier in the chapter. If a v.f.o. operating on, say, 3·6 Mc/s is deviated to give a swing of 5 kc/s and if this is followed by a frequency doubler, the swing at 7·2 Mc/s is 10 kc/s, and so on. Therefore the deviation must be adjusted so that the maximum allowed value is not exceeded on any band. To meet this requirement the exciter unit can of course be designed so that it operates with constant frequency multiplication on all bands. This implies, however, a different v.f.o. frequency for each band, which would be an unwelcome complication unless the only form of modulation contemplated is n.b.f.m.

The simplest method of calibrating the deviation control is to read off from the reactor characteristic the input voltage required at the grid of the reactor to give the desired deviation of the v.f.o. for each band. It is then necessary to know the *peak* output voltage of the speech amplifier and adjust it accordingly. The output may be measured with a peak voltmeter, or alternatively a sufficiently accurate estimate may be made by multiplying the peak output voltage of the microphone by the gain of the speech amplifier. Thus it is possible to arrive at a fairly reliable calibration of the deviation control for each band in terms of an arbitrary scale calibrated say, 0–10.

Speech Clipping in F.M. Transmitters

As in amplitude modulation, speech clipping may be applied to an n.b.f.m. transmitter with the same advantages as apply to a.m. Low-level clipping must be used since there is no equivalent to high-level clipping in an f.m. transmitter, and the clipper/filter circuits must therefore precede the reactance valve.

It should be noted that the use of a low pass filter is implied in the licence conditions covering the use of n.b.f.m.

TYPICAL MODULATORS

Details follow of various types of amplitude modulator having outputs from 15 watts to 140 watts and a reactance valve type of frequency modulator.

It is not essential to follow exactly the arrangements shown and considerable variation is possible provided that the general guidance given earlier in this chapter is followed. In particular, a logical sequence of stages should be adopted and long leads in grid circuits (especially in the input stage) must be avoided.

Modulator Sub-unit

This modulator sub-unit consists of a two-stage speech amplifier and all the components of a push-pull output stage with the exception of the modulation transformer and the top resistor of the screen supply potentiometer.

The speech amplifier is a twin-triode type 12AX7, the two halves of which are connected in cascade by resistance-capacitance coupling. The gain control potentiometer is immediately after the first stage. The second half of the 12AX7 is coupled to the output stage by means of a push-pull input transformer.

The overall gain of the sub-unit is about 1600; thus, depending on the output of the microphone used, output valves up to the KT88 in size may be used with whatever size of modulation transformer and power supply is available.

The sub-unit is built on an aluminium chassis 6¾ in. × 3¾ in. × 1½ in. connections to the screen supply, heater supply

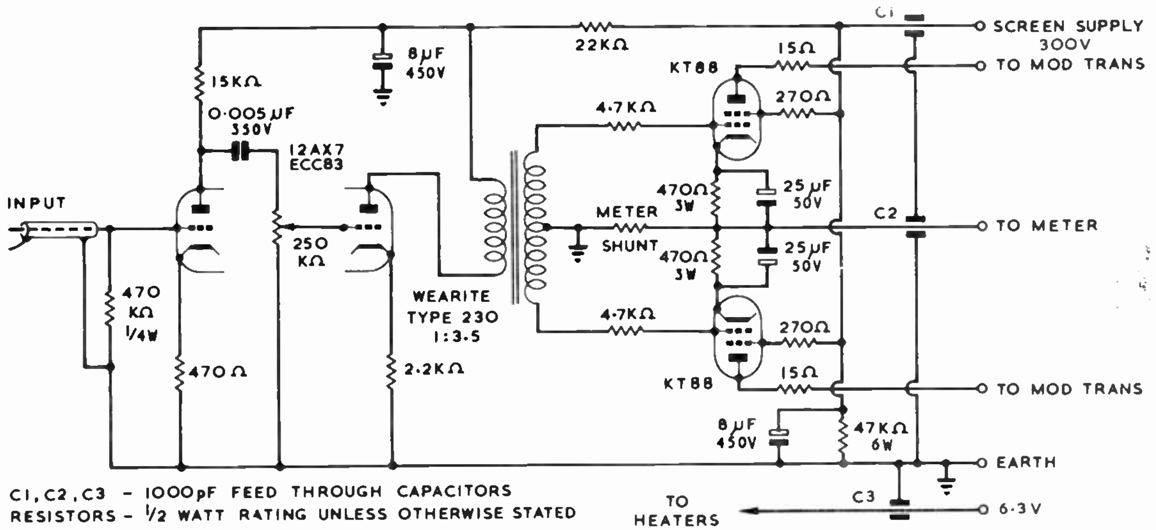
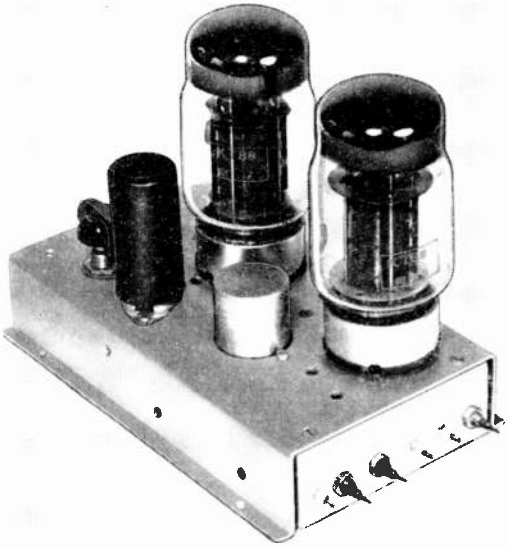


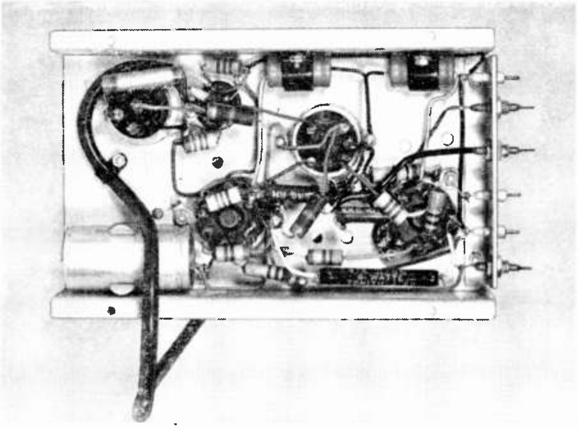
Fig. 9.69. Modulator sub-unit.



General view of the Modulator Sub-Unit.

and to a meter to monitor the total cathode current of the output stage are taken via 1000 pF feed-through capacitors at one end of the chassis. Connections to the modulation transformer are taken via small feed-through insulators. The input to the first valve is via a short length of coaxial cable.

The component values shown in the circuit diagram (Fig. 9.69) apply to the use of KT88's in class AB1 as output valves. The measured operating conditions are as follows:



Underside view of the Modulator Sub-Unit.

Input (at 1000 c/s) to first grid, 50 mV.
Input to grids of output stage, 43V + 43V.
Output (in 6600 ohm resistive load on secondary of Woden UM1 modulation transformer), 33 W.
Top resistor of screen supply potentiometer, 7·8 K ohms.
Screen voltage, 300V.
Total cathode current of output stage, 130 mA.
H.T. input, 500V at 145 mA.

A Modulator for a 25 Watt Transmitter

This modulator is conventional in design and uses miniature valves in all stages. It may be driven to full output by a crystal microphone. The circuit is shown in Fig. 9.70.

The first stage uses a high gain a.f. pentode type EF86. This is *RC* coupled to a paraphase phase-splitter (twin triode type 12AT7) by a low-value coupling capacitor (0·001μF) and the gain control which is a 500 K ohm potentiometer.

The output stage consists of two valves type EL84 in push-pull and operating in class AB1. Cathode bias is used and grid, screen and anode stoppers are provided. The modulator shown in the photograph uses a " surplus " modulation transformer rated at 20 watts and having a ratio of 1 : 1 but a small multi-ratio transformer may be used.

This unit was designed to be part of a 25 watt v.h.f. transmitter, both a.f. and r.f. sections operating from a common 300 volt h.t. supply. In order to restrict the output to the required value, the screen voltage of the output valves is reduced by a 5600 ohm resistor in the screen supply and the overall h.t. by means of a 1000 ohm resistor in the 300 volts line. For c.w. operation, the modulator may be switched off by switch S which opens the h.t. supply and also short circuits the secondary of the modulation transformer.

The modulator is constructed on an elongated *U*-chassis 12 in. × 3 in. × 2 in. deep and the layout of the main components and valves may be seen in the photographs.

The total power requirement is 300 volts at approximately 85 mA and 6·3V at 2·1 A. The h.t. may be obtained from a simple power supply with a capacitor-input filter. Under these conditions the output is 12·5 watts; by removal of the 5600 ohm and 1000 ohm resistors referred to earlier the output may be increased to 17 watts. The h.t. supply is then 300V at 125 mA.

A Compact High-Power Modulator

This modulator which provides an output of 100 watts uses four tetrodes type KT66 in parallel push-pull in the output stage. It is built on a chassis 17 in. × 10 in. × 2 in. which also houses all the power supplies necessary. The front panel is 19 in. × 8¾ in. The circuit diagram is given in Fig. 9.71.

The output valves operate in class AB1 in the ultra-linear mode. The modulation transformer is a Woden type UM3 and the screen supply is obtained from taps 2 and 5, the anodes being connected to terminals 3 and 4. Grid stoppers (10,000 ohms) and screen stoppers (220 ohms) are used in the output stage. The total cathode current of the parallel pairs of output valves is measured by a 0–150 mA meter which can be switched (switch S3) to either pair of valves.

The speech amplifier, which is coupled to the output stage by means of a small push-pull input transformer (ratio 1 : 3), consists of a triode-connected a.f. pentode type Z729 (EF86 or 6267) which is resistance-capacitance coupled to a twin-triode type 6SN7, the two sections of which are also *RC* coupled. The gain control (a 1 Megohm potentiometer) is immediately after the first stage. The speech amplifier is generously decoupled at r.f. and a.f. to prevent instability and is built into a small die-cast box $4\frac{11}{16}$ in. × $3\frac{11}{16}$ in. × $2\frac{1}{4}$ in. This box is mounted through an opening cut in the chassis so that the coaxial input plug protrudes through a small hole in the front panel.

The h.t. supply for the output stage (650 volts) is obtained from a conventional bi-phase full-wave rectifier arrangement which is composed of series-connected silicon diodes type SX635. A capacitor-input smoothing circuit is used and the output capacitor consists of two 160μF 450 volts working electrolytic capacitors in series. A 100 K ohm resistor is connected across each to equalise the voltage across them. The h.t. supply for the speech amplifier is obtained from a resistive potentiometer connected across the 650 volts supply. The grid bias for the output valves is obtained from a small 250-0-250 volts transformer, one 6·3 volts secondary winding of which is connected to a 4 volts supply obtained from the main heater transformer. The bias is adjusted by means of the two 10,000 ohm potentiometers.

The on-off switch S1 controls the mains input to both the h.t. and heater transformers while the supply to the h.t. transformer is also controlled by switch S2. For remote control purposes, a relay is connected in parallel with S2. For c.w. operation, the secondary of the modulation transformer is shorted by S4.

The following controls are mounted in a row at the bottom of the front panel; mains on-off switch, pilot lamp, h.t. on-off switch, microphone input plug, bias control (a), secondary winding shorting switch and bias control (b). Above these are the gain control, meter, meter switch and output terminals. The mains input plug and a two-pin plug for the relay operating supply are mounted on the rear chassis drop.

The setting-up procedure is simple; before the h.t. is applied, the bias potentiometers are adjusted to give maximum bias (at least −200 volts), h.t. is then switched on and each bias is slowly reduced to give a standing cathode current of 70 mA under no-signal conditions in each parallel pair of output valves. The values of bias voltage obtained should agree to within about 5 per cent.

High-Power Modulator

This modulator is capable of an output of 140 watts, obtained from a pair of tetrodes type TT21 operating in class AB1 at an anode voltage of 1000 volts with a screen voltage of 300 volts and a grid bias of approximately −40 volts.

As will be seen from **Fig. 9.72**, the circuit is simple and straightforward. The speech amplifier consists of two twin triodes; the first (12AX7) is a self-balancing floating paraphase phase inverter which feeds a push-pull voltage amplifier (12AU7).

In the interests of simplicity, no balancing arrangements for the output stage are provided, but the bias should be adjusted so that under zero-signal conditions the anode current of each output valve does not exceed 35 mA at an anode voltage of 1000 volts (i.e. the anode dissipation must not exceed 35 watts).

The modulator is built on a chassis which measures 17 in. × $6\frac{1}{2}$ in. × 2 in.

The operating conditions of each output valve are as follows:

Anode voltage	1000	1000	1000	1000V
Screen voltage	300	300	300	300V
*Grid voltage**	−40	−40	−40	−40V
Anode current	35	56	80	92mA
Screen current	4	5	9	17mA
Anode dissipation	35	35	30	22W
Screen dissipation	1·2	1·5	2·7	4·9W
Output†	0	40	100	140W
Anode load (a-a)		16,800	16,800	16,800Ω
Distortion		0·6	1·4	4·4%

*Set to give anode current of 35 mA per valve at anode voltage of 1000V.

† Measured in resistive load of 7000Ω.

Input to first stage (gain control at max., output of 100W) = 25mV.

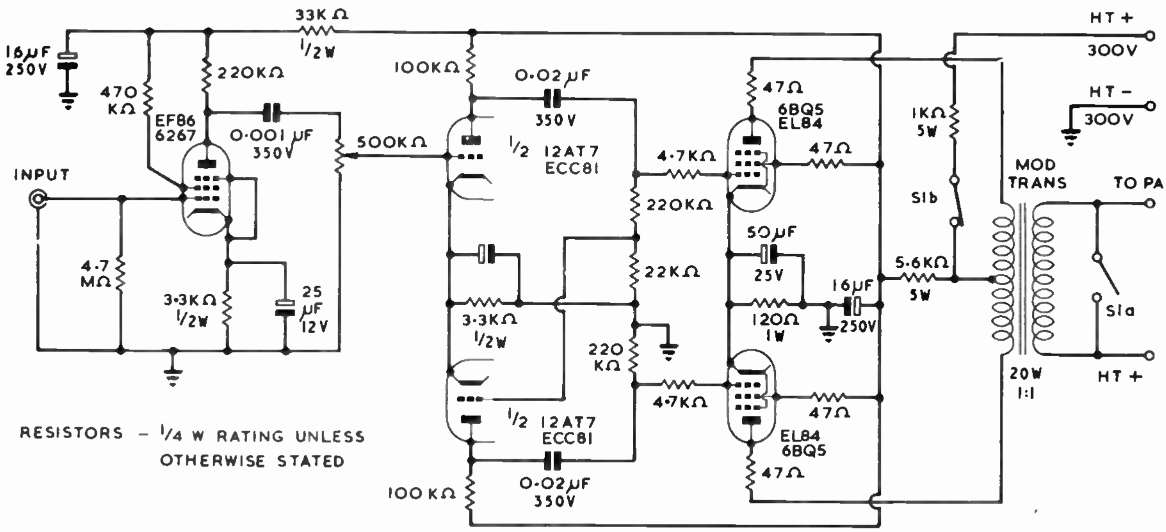


Fig. 9.70. Modulator for a 25 watt transmitter. A 0·1µF decoupling capacitor should be connected between EF86 screen and earth.

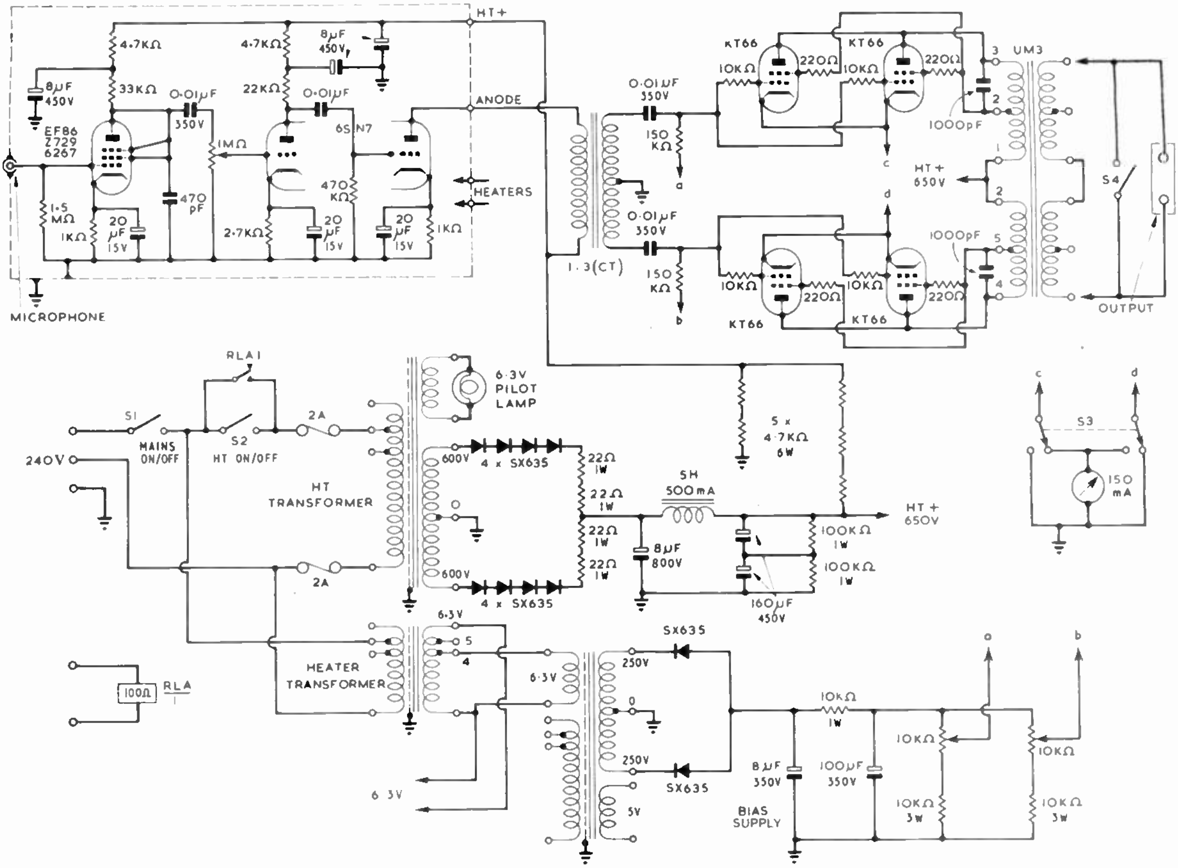


Fig. 9.71. Compact high power modulator.

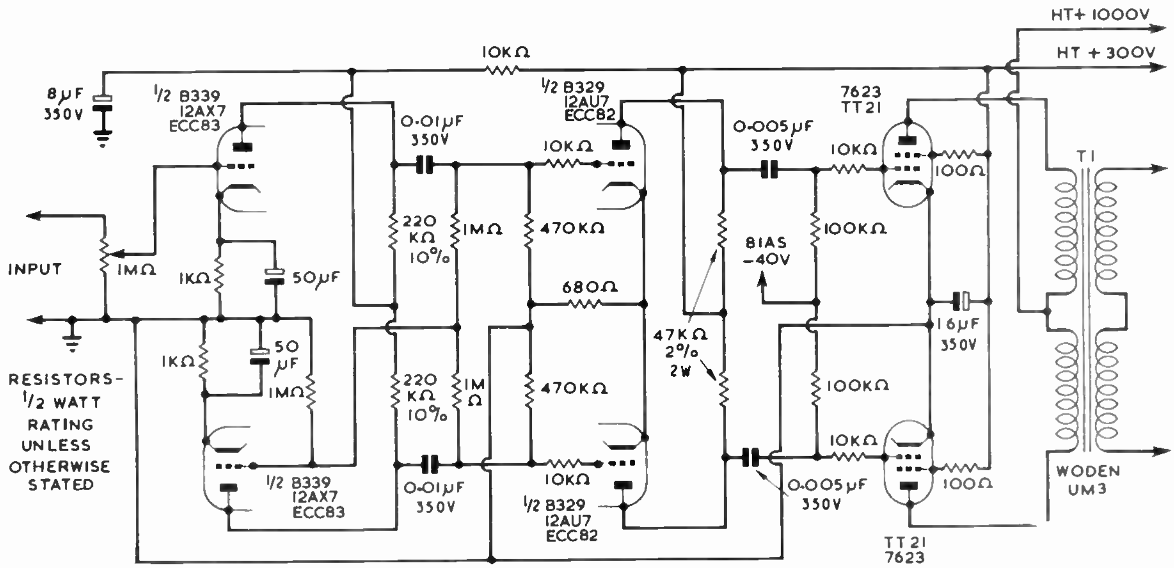


Fig. 9.72. High power modulator.

A 45 Watt Transistor Modulator

The circuit of a modulator employing a pair of power transistors in class B in the output stage is shown in Fig. 9.73. It gives an output of 45 watts and has sufficient gain to allow the use of a crystal microphone. Operation is from a 12 volt supply (earthed positive lead, maximum current drain 7·0 amps) and so the unit is suitable for mobile or portable use.

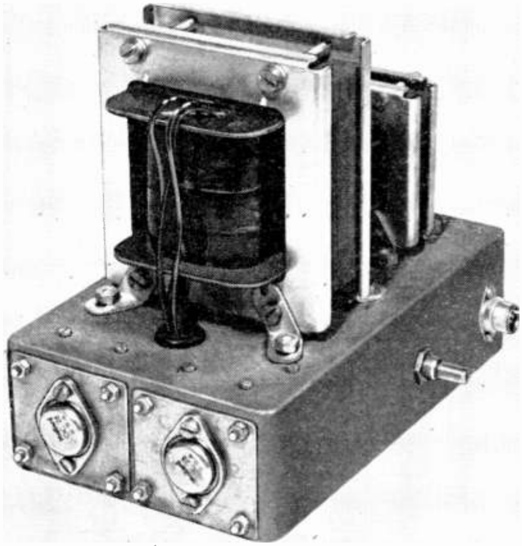
A total of six stages arranged in negative feedback pairs are used. The first two pairs of stages have both d.c. and a.f. feedback applied for d.c. stabilization and improved linearity. This is accomplished by coupling the first stage base of each pair to the second stage emitter through a parallel *RC* network; the first stage collector is directly coupled to the second stage base. The gain control is between the first two pairs of stages and so is not in a feedback loop.

The single-ended class A driver stage and the push-pull output stage also form a feedback pair. Emitter resistance and base potentiometer biasing is used for both stages. the negative supply resistor R1 of the base potentiometer for TR5 being connected to one collector of the output stage. Forward biasing of the output transistors is such as to reduce cross-over distortion to the minimum value. The common emitter resistance R2 is 0·1 ohm and consists of 67 turns of 22 s.w.g. copper wire close-wound on a high value 3 watt carbon resistor. The correct collector for attachment of R1 must be found by trial and error; this must be carefully done with the modulator suitably loaded as the last two stages will oscillate if the incorrect collector is chosen.

Several types of transistor may be used; in the following list, they are given in order of preference.

TR1.	GET 106.
TR2., TR3., TR4.	GET104, GET103, OC72, OC71.
TR5., TR6., TR7.	GET573, GET572, OC28 series (OC29, OC36, OC35, OC28), OC25

The multi-ratio modulation transformer matches the optimum collector-to-collector impedance of the output stage to impedances of approximately 4000, 6000, 8000, 13,000 and 15,000 ohms, due allowance having been made for imperfections in the transformer. Details of both this and the driver transformer are as follows:



General View of the 45 watt Transistor Modulator.

Modulation Transformer (T2)

Core 1$\frac{3}{8}$ in. stack of No. 4 laminations (0·015 in. Silcor) M and E Alloys Ltd.

Secondary Winding (wound first) 2200 turns of 30 s.w.g.

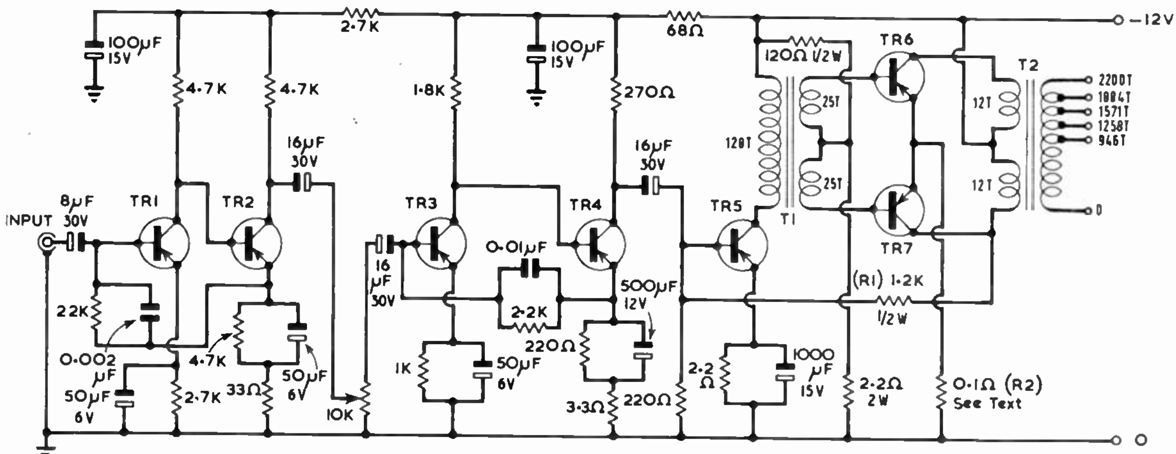
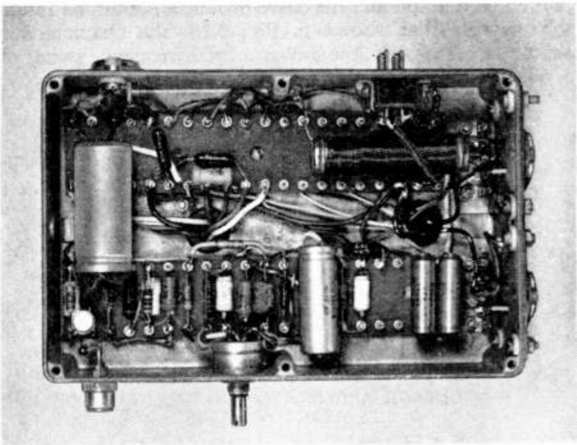


Fig. 9.73. 45 watt Transistor Modulator.



Underside view of the 45 watt Transistor Modulator.

Lewmex wire in 12 layers; one layer of thin paper every fourth layer.

Taps at 946 turns, 1258 turns, 1571 turns and 1884 turns.

Start, taps and end of winding brought out with 7/0·0076 wire, stuck to winding with adhesive Melinex tape.

Two layers of paper over secondary winding.

Primary Winding Two layers, each of 12 turns of 14 s.w.g. Lewmex wire, separate windings.

Wind on three layers of empire tape, insert laminations (interleaved, i.e. no air gap) and wax-impregnate.

Driver Transformer (T1)

Core 1 $\frac{1}{16}$ in. stack of No. 101A laminations (0·015 in. Silcor) M and E Alloys Ltd.

Primary Winding 128 turns of 18 s.w.g. Lewmex wire, in 6 layers, one layer of paper after third layer.

One layer of paper over primary winding.

Secondary Winding Two layers, each of 25 turns of 16 s.w.g. Lewmex wire, separate windings.

Wind on three layers of empire tape, insert laminations (interleaved, i.e. no air gap) and wax-impregnate.

The unit, with the exception of the output transistors, is built inside an Eddystone die-cast box measuring 7¾ in. × 4½ in. × 2 in. TR5 is bolted to the side of the box by 4B.A. nylon screws, the insulating washer being smeared with silicone grease. Each output transistor is mounted on an 18 s.w.g. brass plate, measuring 2 in. × 2¼ in. These plates are mounted at one end of the box and are insulated from the box by mica plates slightly larger than the brass plates. The mica plates should be smeared with silicone grease to ensure good thermal contact with the box. If metal screws are used to hold down the plates, they must of course be insulated by fibre washers and bushes. The transistor mounting arrangements are clearly shown in the photograph.

Apart from ensuring that R1 is connected to the correct transistor, no setting up is required.

Narrow-Band Frequency Modulator

The circuit of a typical reactance valve modulator is shown in Fig. 9.74. This consists of two stages, a speech amplifier V_1 and a reactor valve V_2. The speech amplifier is intended for use with a crystal microphone or other low-output type and is therefore designed to have a high gain. The microphone input is shielded right up to the grid of V_1. It would be advisable to reduce the gain of the first stage if a high-output microphone is used. The reactor stage V_2 is a pentode having a variable-μ characteristic. The phase-shift network consists of the resistor R_9 and the stray capacitance between the grid of V_2 and earth, i.e. the input capacitance of V_2 plus the wiring and valveholder capacitances. The a.f. input to the reactor is controlled by the potentiometer R_7. This control is, of course, the deviation control. A certain amount of top cut is introduced by the capacitor C_8 shunting R_7. C_8 and C_{10} are blocking capacitors. The output lead which connects to the tuned circuit of the v.f.o. should be rigid and as short as possible to prevent any unwanted variation of the frequency. Alternatively coaxial cable may be used if the length of the lead exceeds an inch or two. With care and by the use of miniature components, the whole unit can be made on a very small chassis so that it may be located conveniently close to the v.f.o. The screening afforded by the chassis should be as effective as possible and the h.t. supply (150 volts) preferably should be stabilized by a voltage regulator tube such as the VR150/30.

In setting up the reactor stage the first step is to determine its characteristic, i.e. the relationship between the voltage applied to the grid of the reactor and the resulting deviation of the frequency of the v.f.o. This may be done quite simply by isolating the reactor and applying a variable d.c. voltage to its grid as shown in Fig. 9.75. The frequency of the v.f.o. must first be measured by an accurate frequency meter, zero voltage being applied to the reactor grid. When the grid voltage is increased, the frequency of the v.f.o. will be found to change. The change in frequency for a known change in grid voltage should be measured by the frequency meter.

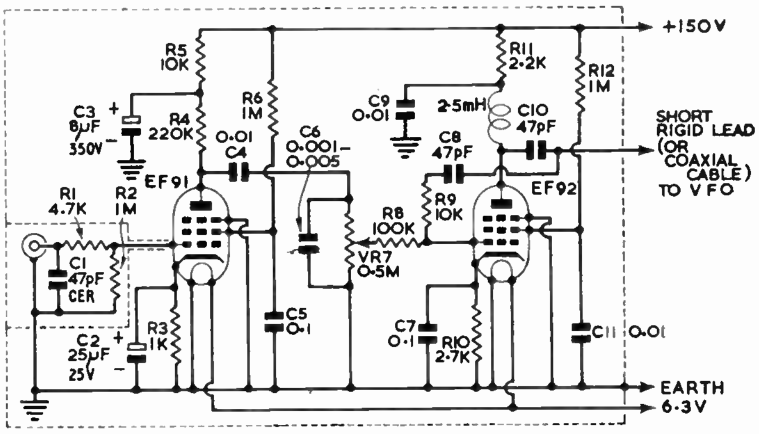


Fig. 9.74. Circuit diagram of a typical reactance valve modulator.

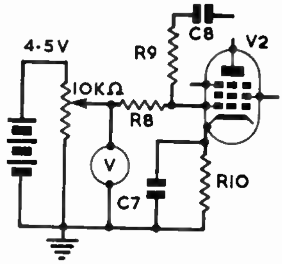


Fig. 9.75. Test circuit for determining the static characteristic of a variable-reactance modulator. V2 must be isolated from the speech amplifier.

This should be continued up to a maximum frequency deviation of about 4 kc/s or so. Next, the polarity of the battery from which the grid voltage is obtained should be reversed and the test repeated. Frequency changes in the opposite direction should now be obtained. The results should be plotted on a graph and the curve produced should be found to resemble that shown in **Fig. 9.76.** The characteristic should be reasonably linear over a total frequency swing of 5–6 kc/s.

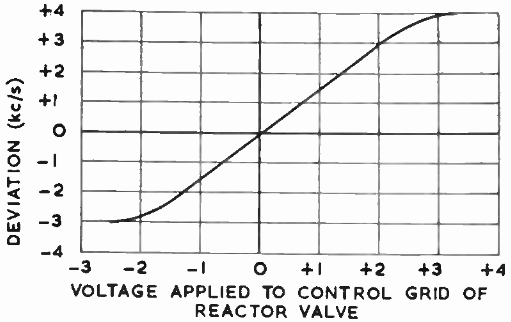


Fig. 9.76. Static characteristic of the reactor valve of the n.b.f.m. unit shown in Fig. 9.74. The frequency of the v.f.o. is approximately 3600 kc/s.

SINGLE SIDEBAND TRANSMISSION

SINGLE sideband suppressed carrier telephony transmission (commonly called s.s.b.) is a specialized form of amplitude modulated telephony and a brief examination of basic a.m. theory is therefore an essential preliminary to any description of the more advanced system.

A.M. Carrier and Sideband Relationships

The carrier of an a.m. transmission does not vary in amplitude; it is at all times of constant strength. The modulation introduced at the transmitter heterodynes the carrier and produces sum and difference frequencies. These are symmetrically disposed either side of the original carrier frequency and constitute two bands of side frequencies—those below the carrier form the lower sideband and those above the carrier form the higher sideband. The carrier by itself does not convey any intelligence; the intelligence is conveyed solely by the sideband frequencies.

Consider what happens when a carrier of 1000 kc/s is modulated 100 per cent by a pure audio tone of 2 kc/s. The energy propagated from the aerial would be three entirely separate and individual r.f. outputs on 998, 1000 and 1002

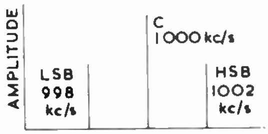


Fig. 10.1. Amplitude frequency relationship of a.m. signal modulated with single tone of 2 kc/s.

kc/s (**Fig. 10.1**). These three channels of r.f. energy would travel quite separately through the ionosphere and would eventually arrive at the receiving aerial and would then be accepted by the receiver, heterodyned by the local oscillator and converted to the final intermediate frequency of the receiver. If the i.f. passband was centred on 460 kc/s the i.f. amplifier would present three separate frequencies—462, 460 and 458 kc/s—to the detector. The combined effect of these three frequencies at the detector would in turn produce the modulation envelope. The way in which the modulation envelope is produced can be shown very simply by means of vector diagrams.

The carrier vector is actually rotating at 460,000 c/s with one sideband vector rotating at 458,000 c/s and the other sideband vector rotating at 462,000 c/s. Relative to the carrier, one sideband vector is rotating 2000 c/s slower (lagging) and the other sideband vector 2000 c/s faster (leading). If then the carrier is assumed to be stationary and is drawn as a vertical line whose length denotes the carrier voltage—let this be one unit in length—each sideband vector would be 0·5 unit in length (the power in one sideband is equal to 0·25 of the power in the carrier and as power is equal to E^2/R, the voltage is equal to 0·5 that of the carrier) and would be shown rotating round the carrier vector in opposite directions at the frequency of the modulation (**Fig.

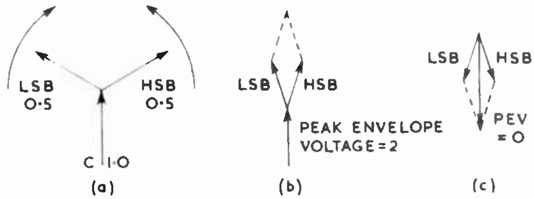


Fig. 10.2a, b and c. Vector presentation showing how the two sideband voltages combine with the carrier voltage to form the modulation envelope.

10.2 (a)). At some moment of time, the two sidebands will be in phase with each other and in phase with the carrier and the resultant vector length will be two units—the modulation crest. At 180° of rotation later they will again be in phase with each other but opposite in phase to the carrier and the resultant vector length will be zero—the modulation trough. (**Fig. 10.2(c)**). At all other degrees of angular rotation the resultant voltage will be some in-between value. The modulation envelope recovered at the detector is shown in **Fig. 10.3** and to the same scale, the audio output in **Fig. 10.4.**

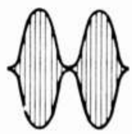


Fig. 10.3. Modulation envelope recovered at the detector.

Fig. 10.4. Resultant audio output voltage.

If, therefore, the resultant r.f. voltage output from a 100 per cent sine wave modulated transmitter is examined by means of an oscilloscope, the display will show the modulation envelope of the classical textbooks. The important point to understand is that the oscilloscope is not showing the true relationship of the carrier and the sidebands at all. It cannot separate them and show each individual component. The display seen is the resultant effect in terms of voltage—in fact, the oscilloscope is showing a continuous panorama of voltage vector diagrams.

When the single tone modulating signal is replaced by the output of a speech amplifier, the picture grows more complex. It may, however, still be analysed into the original carrier, unchanged in amplitude and frequency, about which are displaced symmetrically the upper and lower sidebands. The sidebands will, of course, undergo continuous change in amplitude and frequency as the voice changes in inflexion and intensity, but they may never exceed half the amplitude of the carrier if over-modulation is to be avoided Their maximum excursion in frequency is limited by the highest frequency component of the voice; for practical communications purposes this may be taken as 3 kc/s.

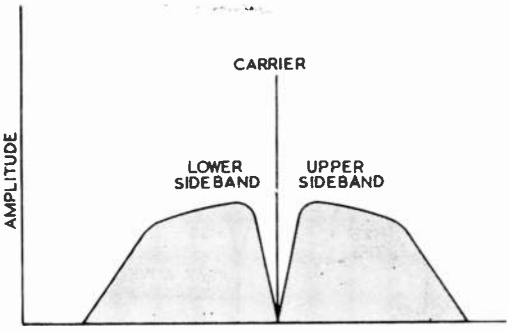


Fig. 10.5. Amplitude/frequency relationships of carrier and sidebands with 100 per cent speech modulation.

Fig. 10.5 illustrates in diagrammatic form the relationship of the carrier and its sidebands under speech modulation; it is this representation, rather than that of Fig. 10.2, which forms the foundation stone on which to build a solid understanding of s.s.b. There is, however, one noteworthy point which is more immediately evident from Fig. 10.2 than from the alternative representation. At the modulation crest the carrier and sideband voltages are in phase and add together; the voltages at the anode and screen of the modulated stage will also rise to twice the applied d.c. potentials, so the components in these circuits must be appropriately rated to avoid breakdown. As the transmitter peak envelope voltage (p.e.v.) at the crest of modulation is double that of the unmodulated carrier, and as output power is proportional to voltage squared, the peak envelope power (p.e.p.) will be increased four times. Assuming that the modulated stage is being operated under the usual amateur conditions of 150 watts d.c. input with an overall efficiency of 66 per cent, the unmodulated carrier output will be 100 watts, and the peak envelope power output 400 watts.

Single Sideband Transmission

If the transmitter of Fig. 10.1 were radiating a lower sideband s.s.b. signal, the carrier on 1000 kc/s and the higher sideband on 1002 kc/s would be suppressed at the transmitter, and the only output to the aerial would be continuous r.f. energy on 998 kc/s. This would be converted by the receiver to an i.f. of 462 kc/s and fed to the detector. The carrier of 460 kc/s would be generated by a local oscillator (b.f.o.) and also fed to the detector. A vector diagram would show the inserted carrier from the b.f.o. as a vertical line with the sideband vector rotating around it at the modulating frequency; the modulation envelope would be recovered at the detector and the resultant audio output would be 2 kc/s. It will be seen that the original 2 kc/s tone input has been recovered, yet the s.s.b. transmitter has radiated one continuous r.f. output on one frequency only, 998 kc/s.

If the tone modulation at the transmitter had been changed from 2 to 3 kc/s the sideband radiated would change from 998 to 997 kc/s. This would be converted by the receiver to an i.f. of 463 kc/s, combined in the detector with the local carrier on 460 kc/s and the resultant audio output would be 3 kc/s. Should the transmitter be modulated with both the 2 and 3 kc/s tone simultaneously two r.f. outputs would be radiated from the aerial on 997 and 998 kc/s. These would be converted by the receiver to two i.f. outputs of 463 and 462 kc/s. The resultant output recovered by the detector would be 3 and 2 kc/s—the original modulating frequencies.

When the tone modulation is replaced by the output of a speech amplifier the transmitted sideband becomes essentially a band of frequencies undergoing continuous changes in amplitude and frequency as the voice changes in inflection and intensity. Provided that the receiver i.f. passband is sufficiently wide, all frequencies within the received sideband will be amplified equally and fed into the detector. The locally inserted carrier merely serves as a datum line against which the sideband is demodulated and the resultant audio output is the original speech modulation.

COMMUNICATIONS EFFICIENCY

In an a.m. transmitter, the amplitude of each sideband is limited to half that of the carrier, so its maximum power will be one-quarter of the unmodulated output power. In the example already quoted of the 150 watt input amateur transmitter with an overall efficiency of 66 per cent, the unmodulated carrier output would be 100 watts, and the maximum power in one sideband would be 25 watts; both sidebands together would produce 50 watts. It is not, however, possible to specify the effectiveness of the transmission until something is known about the equipment on which it is being received. The sidebands necessarily occupy a band of frequencies 3 kc/s on either side of the carrier, so that if the receiver has a bandwidth of 6 kc/s they both add in phase and contribute to the total *talk power*. If a more selective receiver with a 3 kc/s bandwidth is used, only one sideband contributes to its output. Under the crowded conditions which obtain in the amateur bands, good selectivity is almost invariably a necessity, so that only 25 watts of the signal actually conveys intelligence to the listener. This is not a particularly profitable return for the generation of 100 watts of carrier power, not to mention the cost, weight and bulk of a high level modulator and associated power supply.

As the carrier remains constant in frequency and amplitude it conveys no intelligence from transmitter to receiver, and serves merely as a datum line against which the sidebands are demodulated. All the intelligence is conveyed in the sidebands, so that the carrier may be omitted and the signal demodulated perfectly clearly if the reference function is transferred to a local oscillator at the receiving end. The attractiveness of substituting for the 100 watt carrier the signal from a small oscillator valve in the receiver is obvious. The reduction of heterodyne interference within the amateur bands, which results from carrier suppression, may alone appear to be adequate justification for making the change, but there are other advantages.

The power amplifier which produces a 100 watt carrier must be capable of a peak output of 400 watts. With modifications the full capacity of the amplifier may be utilized to radiate intelligence conveying sideband energy. This does not necessarily mean that the peak envelope power output of the hypothetical amplifier will be 400 watts, because there are important differences between the operating conditions of anode-modulated class C amplifiers and the class AB or B linear amplifiers customarily employed in s.s.b. output stages. The question of linear amplification will be dealt with in detail later in this chapter, but as a rough guide it may be said that a 100 watt a.m. stage should be able to produce about 200 watts p.e.p. output in s.s.b. service. The whole of these 200 watts will contribute to the talk power at the receiving end, which compares very favourably with the 50 watts which produce speech output from a conventional a.m. transmission in a receiver of 6 kc/s bandwidth, and even more favourably if the receiver has a narrower bandwidth.

There is no difficulty at all in suppressing the carrier; a miniature double triode, or a pair of crystal diodes, together with a few resistors and capacitors are all that is necessary, leaving only the problem of how to get rid of one of the sidebands. It has already been said, however, that a receiver of 6 kc/s bandwidth is capable of combining the two sidebands so that both may contribute to the audio output. Furthermore, it matters little at the transmitting end whether the linear amplifier is producing single sideband or double sideband output, so long as the peak power remains substantially the same in both cases.

It is, however, desirable to eliminate one sideband in order to simplify the demodulation process in the receiver Even if the local oscillator at the receiver can be made stable enough to simulate the precise frequency of the suppressed carrier, that alone is not sufficient. It must also be identical in *phase*, otherwise the two sidebands will not combine to produce anything approaching the original modulating waveform. The best way to utilize fully a double sideband suppressed carrier (d.s.b.) signal is to tune it in on a wideband receiver to which is coupled an adaptor capable of synchronising the local oscillator to the required frequency and phase by means of information derived from the sidebands themselves. The main disadvantage of this system of detection lies in its susceptibility to lose synchronization because of interference from random noise or from strong adjacent-channel signals.

If, however, one of the sidebands is eliminated as well as the carrier, reception becomes fairly easy. The local oscillator may wander quite a few cycles away from the correct frequency before intelligibility suffers appreciably and does not have to be locked in phase. The best place to eliminate the unwanted sideband is clearly at the transmitter, because the full capacity of the p.a. stage may then be used to amplify the wanted sideband. In addition the transmission will occupy a bandwidth of only 3 kc/s. Alternatively, the conversion from d.s.b. to s.s.b. may be effected at the receiving end. This is relatively simple because increasing interference has already led many amateurs to increase the selectivity of their receivers to a point at which it is possible to reject one-half of a d.s.b. transmission and deal with the remainder as if it were a true s.s.b. signal. When received in this way d.s.b. is considerably more difficult to tune than s.s.b. and is only half as efficient.

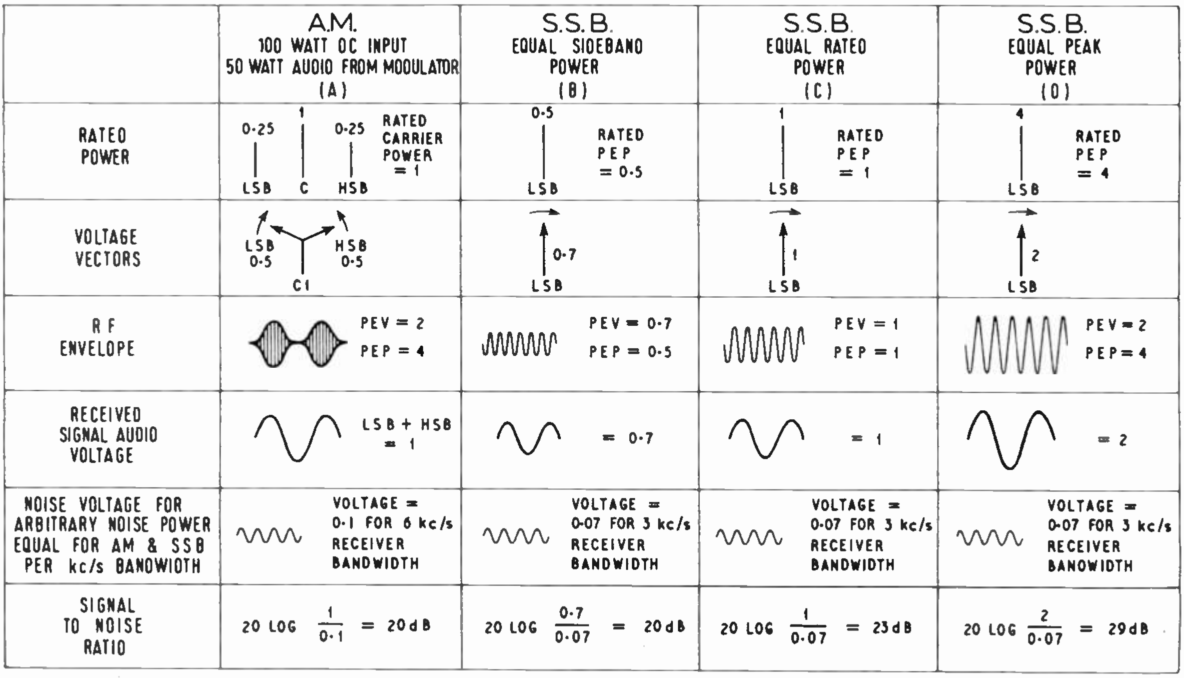


Fig. 10.6. Power relationships for a.m. and s.s.b. transmission. Single tone sine wave modulation.

Half of the power radiated by the d.s.b. transmitter is not used for reception and merely serves to increase the general level of interference. In this respect it is no worse than conventional a.m., although it has the advantage of having no resting carrier to cause a constant heterodyne whistle. When the fact that a d.s.b. transmitter is simpler to construct than an a.m. rig of comparable power output is considered, the reasons for its increasing popularity become evident, though it is to be hoped that users of the system will come to regard it as a stage on the road towards s.s.b., and not an end in itself.

Talk Power

Perhaps the clearest method—least open to misinterpretation—is to show the relative efficiency of the two systems in diagrammatical form where the powers and voltages concerned are to the same relative values. This method has been adopted in **Fig. 10.6.** The basis of comparison given in column A is an a.m. transmission of 100 watts d.c. power *input* rating, modulated by 50 watts of audio. At the crest of the modulation cycle the peak envelope power (p.e.p.) is four times the carrier power—400 watts. The term peak envelope power is defined as the r.m.s. power developed at the crest of the modulation envelope.

The carrier of one power unit in value requires a half-power unit of audio for 100 per cent modulation (this is the maximum power that can be used; any greater audio input would produce overmodulation and distortion) and this produces two sidebands with 0·25 unit of power in each. As voltage is proportional to the square root of the power, the carrier voltage is 1 and the voltage of each sideband is 0·5. The r.f. envelope developed by the voltage vectors is shown, and for 100 per cent modulation the peak envelope voltage (p.e.v.) is the sum of the carrier and the two sideband voltages, and this equals two units. This results in a p.e.p. of four units of power.

The r.f. signal is demodulated in the receiver and the diode detector develops an audio output voltage that is equivalent to the sum of the upper and lower sideband voltages. The noise power per kilocycle is an arbitrary value equal for a.m. and s.s.b. For a 20db signal-to-noise ratio, the noise voltage would be 0·1 volt for the 6 kc/s receiver bandwidth and the signal-to-noise ratio is then 20 log. the ratio audio voltage : noise voltage.

Column B shows the power and voltage relationships for an s.s.b. transmission of equal *sideband* power to the a.m. transmission. The audio output from the s.s.b. transmission (recovered by heterodyning the received signal with a locally inserted carrier) is 0·707 units in value. This represents a loss in detector output voltage of 3db due to the s.s.b. power being in one sideband. However, the reduction in receiver bandwidth gives a 3db advantage therefore the signal-to-noise ratio is the same for the two modes of transmission.

Column C shows the relationship for an s.s.b. transmission of equal *rated* power (100 watt a.m. transmitter and a 100 watt p.e.p. input s.s.b. transmission). It is seen that the audio voltage developed at the output of the diode detector is equal to the audio voltage of the a.m. transmission. The reduction of receiver bandwidth gives a 3db advantage and the signal-to-noise ratio is 23db.

Column D shows the relationship for an s.s.b. transmission of equal *p.e.p. input*. It is seen that the s.s.b. transmission gives a gain in detector output voltage of 6db. This, in addition to the 3db improvement in receiver noise output, gives a total signal-to-noise ratio of 29db, a system gain for s.s.b. of 9db.

It will be noted that the carrier envelope voltage of the a.m. transmission serves no useful purpose other than to beat against the sidebands and demodulate the signal at the detector. Furthermore, that for both modes of transmission the audio voltage recovered at the detector is directly proportional to the total *sideband* voltage, and that the two transmitters, a.m. and s.s.b. of equal *rated* power will produce an equal receiver audio voltage.

S.s.b. Advantages

There are no inherent difficulties in the construction of a s.s.b. transmitter. In the filter type, the chassis, the layout and even the majority of the components are virtually identical to a selectable sideband receiver. The usual 75 watt modulator with its bulky driver and output transformers and separate 150 watt power supply is no longer required. The s.s.b. transmitter can be built and used initially to gain operating experience with a low power output—say 50 watts peak. At a later date the power can be increased to the licensed maximum by building a suitable linear amplifier and h.t. supply and driving this with the existing transmitter.

It is quite true that s.s.b. working demands a higher standard of frequency stability and more care in netting, but this should not be looked upon as a disadvantage peculiar to sideband working. First, as receiver selectivity is improved there will be a demand for better frequency stability

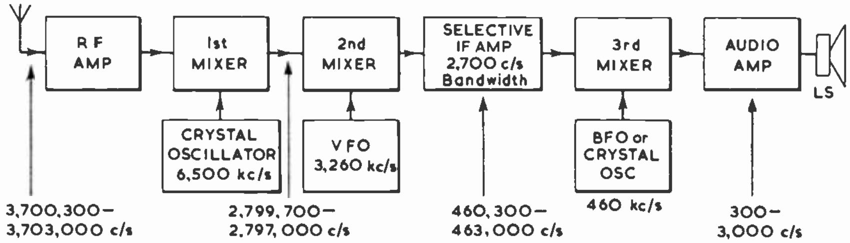


Fig. 10.7. Frequency translation process in a double superhet receiver.

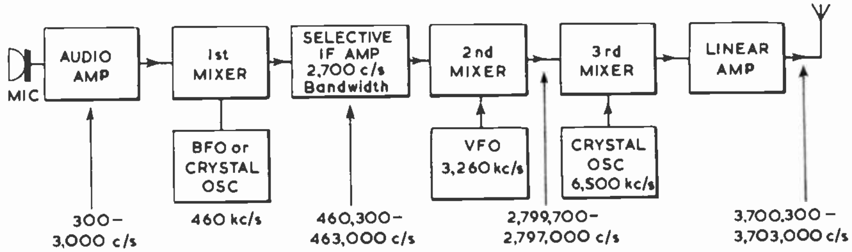


Fig. 10.8. Frequency translation process in an s.s.b. transmitter.

from existing a.m. transmitters. Second, many of the methods developed by amateurs for s.s.b. working are already being adopted to improve a.m. operation. These include exalted carrier detection, improved bandwidth control, better oscillator stability and press-to-talk or automatic voice control. Third, the power gain of 9db with s.s.b. operation (made up of 6db gain at the transmitter and 3db at the receiver) represents an equivalent power increase at the transmitter of eight times.

The advantages of the s.s.b. are, therefore:

(i) the r.f. spectrum required to transmit a given signal by means of s.s.b. is exactly that of the original signal, thus maximum use can be made of the available r.f. spectrum.

(ii) since only essential signals are transmitted by s.s.b. without a superfluous carrier or mirror image sideband, there is a considerable effective power gain.

(iii) most important of all, s.s.b. systems are effected far less adversely by the transmission disturbances inherent in ionospheric transmissions than a.m., f.m., or any of the double sideband systems.

The terms heterodyning and mixing have the same meaning in s.s.b. transmission as in receiver practice and they are both frequency translation processes. All superhet receivers translate the frequency down from the required amateur band to the i.f. channel (or channels) and then to the final audio channel. This is shown clearly by the block diagram of a typical double superhet receiver given in **Fig. 10.7**. If the third mixer is renamed product detector, the circuit arrangement is well-known and easily followed and understood. Suppose then that the translation process is reversed and the block diagram redrawn as in **Fig. 10.8**. The mixing processes are exactly the same as in Fig. 10.7, but the block diagram of Fig. 10.8 is that of a s.s.b. transmitter. In the receiver the translation process is *down* from the r.f. signal frequency to the audio frequency, and in the transmitter *up* from the audio frequency to the r.f. signal frequency. It is possible (in familiar terms) to follow mentally the progress of the signal intelligence from the microphone to the transmitting aerial, and the question, " How can you transmit voice frequencies without a carrier? " does not arise.

SUPPRESSING THE CARRIER

In a single sideband transmitter the initial carrier frequency is modulated by the output from the audio amplifier in a balanced modulator; balancing the modulator almost eliminates the carrier component in the modulator output circuit. The output from this stage is therefore a double sideband suppressed carrier signal. At the modulation crest, the two sideband voltages are in phase with the carrier voltage, i.e., the modulation process is exactly the same as that of the conventional a.m. transmitter.

It is most important in order to understand clearly what is happening in the balanced modulator of a s.s.b. transmitter to remember that when a carrier is " amplitude modulated " it does not in fact vary in amplitude at all. The output from the modulated stage, whether the balanced mixer in a s.s.b. transmitter or the p.a. in an a.m. transmitter, contains three components. These are the upper sideband, the carrier, and the lower sideband. The carrier wave Fc is heterodyned by the audio wave Fm and this produces sum and difference frequencies $Fc + Fm$ and $Fc - Fm$. The modulation process is thus seen to be a frequency translation process and the modulator is therefore a converter or mixer. When an audio frequency modulates a radio frequency, the process is generally called *modulation*, but when a radio frequency modulates another radio frequency it is called *heterodyning*. The processes are, however, identical, and the general terms modulator, converter or mixer mean the same thing.

The sidebands, $Fc + Fm$ and $Fc - Fm$, contain the voice intelligence, so the carrier frequency can be attenuated or balanced out in the modulator output without affecting the sidebands in any way. This can be done at high level in the p.a. stage of an a.m. transmitter* or at low level in the balanced modulator of an s.s.b. transmitter. In both cases, the output from the modulator is exactly the same—a double sideband suppressed carrier signal.

Balanced Modulators

The usual way to eliminate the carrier is by phasing in a balanced modulator. The possibilities for varying the circuit design are virtually limitless, so there is some excuse for assuming the balanced modulator to be complicated. In fact, it is a straightforward application of the bridge circuit and there is no reason why the simplest design should not perform as well as the more elaborate ones.

Balanced modulators fall into the following two subdivisions:

(i) The high impedance type which employs amplifier valves.

(ii) the pure bridge or ring circuit which uses semiconductor diodes and operates at a much lower impedance.

* This reference is to a d.s.b.s.c. transmitter using a push-pull p.a. stage.

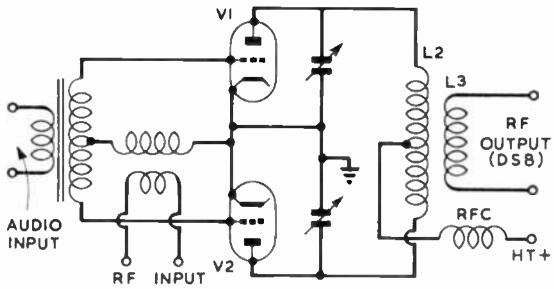


Fig. 10.9. Basic circuit for twin-triode balanced modulator.

The basic circuit of the high impedance type is shown in **Fig. 10.9.** An r.f. voltage applied to the grids of V1 and V2 will appear in amplified form at the anodes. This will cause r.f. currents to flow in opposite directions through the two halves of L2. If the circuit is so adjusted that the signal induced in one-half of L2 is balanced exactly by that induced in the other half, no energy whatsoever will be passed on to L3. Balance is governed primarily by the amplification factors of V1 and V2; ideally these valves should be identical, but this is difficult to achieve in practice and the best that can be done is to choose a pair whose characteristics match as closely as possible. The two units in a double-triode envelope often match fairly well, and if half-a-dozen valves are available at least one should be good enough for balanced modulator service. Almost any valve type can be made to give good results, but for ease of adjustment and long-term stability of balance, a low-mu triode such as the 12AU7, 12BH7 or GEC A2900 is recommended. The characteristics of high-mu types such as the 12AT7 are more prone to change as the valve ages. A valve which has seen enough service to permit its characteristics to stabilize is more likely to prove satisfactory than a brand new one; if a new valve must be used it should be aged artificially by operating it at normal heater voltage and a few milliamperes of anode current for 50-100 hours before it is tested for use in a balanced modulator. The importance of selecting the right valve cannot be overstressed. While it is possible to balance out large inequalities by external means, the circuit will drift out of balance quite rapidly, and will require continual readjustment.

Although a well-matched pair of valves will give about 25db carrier suppression on their own, this is rarely enough, and additional balancing is required. If the remainder of the circuit permits, the simplest way is to vary the operating point of one or other of the valves by altering the standing bias. An example of this method is shown in **Fig. 10.10.**

The basic arrangement may, of course, be varied widely to suit particular conditions, but there are certain pitfalls for the unwary. As the effective earth must be determined by the capacitors C1, C2, the tapping point on the coil at which the h.t. supply is fed to the valves must never be brought to r.f. earth potential by means of a bypass capacitor. An r.f. choke as shown in Fig. 10.10 will ensure that the tapping point will float in the required manner, and if the consequent voltage drop can be tolerated this component may be replaced by a resistor of 10K ohms or more. When the circuit is used at signal frequency the coil is usually tapped, but at intermediate frequencies the necessity arises to feed h.t. in parallel by means of two resistors in the manner shown in **Fig. 10.11.**

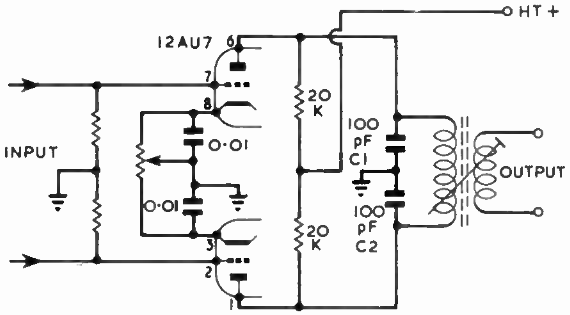


Fig. 10.11. Resistive method of supply anode voltage to twin-triode balanced modulator.

From time to time, circuits on the lines of **Fig. 10.12** have appeared in which the grids are excited in push-pull and the anodes strapped in parallel. Although there is no obvious reason why this arrangement should not be capable of as good performance as that of Fig. 10.10, it is often found more difficult to adjust and less stable in operation. It has the further disadvantage that (because of its similarity to the well-known push-push doubler) even-order harmonics are accentuated in its output and may give rise to spurious signals. Unless considerations external to the balanced modulator dictate the use of the push-pull configuration, it is best avoided.

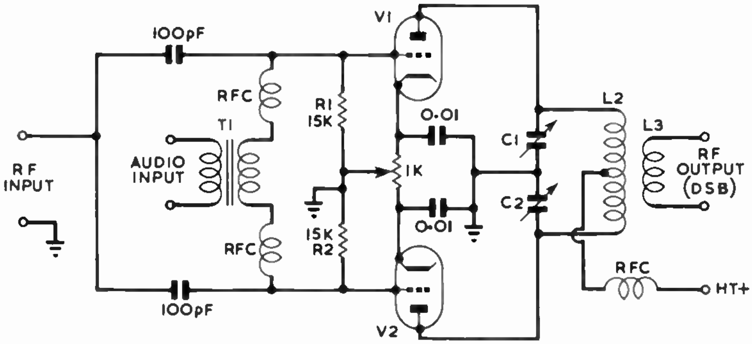


Fig. 10.10. Balancing potentiometer adjustment of grid bias.

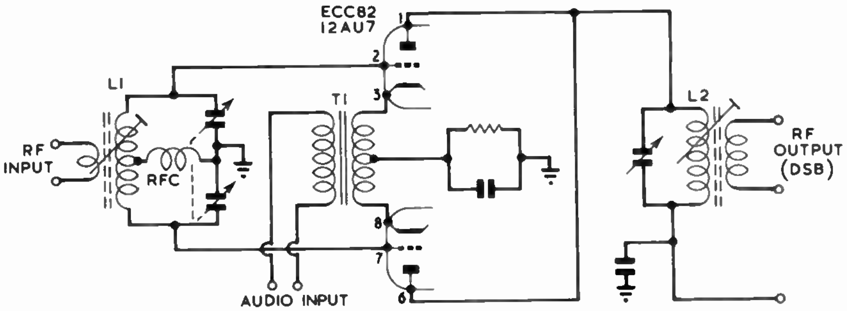


Fig. 10.12. Push-pull balanced modulator not recommended for general use (see text). Any step-down interstage transformer with centre-tapped secondary may be used at T_1.

Multi-grid valves may be substituted for triodes in the circuits already described. Their performance is not inherently better than triodes, but isolation offered by the separate grids makes them attractive for some applications. **Fig. 10.13** (a) and (b) show configurations which are suitable for pentodes and pentagrids respectively. The r.f. carrier is applied in parallel to the control grids of the pentodes and the oscillator injection grids of the converter valves. Modulating voltage is applied in push-pull to the screen grids of the pentodes, and to the signal injection grids of the pentagrids. Adjustment of r.f. balance is simple; the balancing potentiometer merely varies the operating points of the valves by altering the applied bias.

Circuits have been described in which balance is adjusted by the differential variation of the r.f. excitation applied to the grids. This may be done by replacing R1 and R2 by a potentiometer, the slider of which is returned to earth, and by substituting fixed resistors for the adjustable cathode network of Fig. 10.13. The general arrangement may be gathered by the skeleton diagram of **Fig. 10.14**. The disadvantage of this scheme is that both ends of the balancing potentiometer are at r.f. potential, so that the stray capacitance of this component is likely to introduce phase shifts which are unpredictable as they are undesirable. This method has nothing to commend it and should be avoided.

If a high degree of balance is to be maintained over a considerable period of time, it is essential to stabilize the anode supply to triode balanced modulators, and the screen supplies to multigrid valves. Gas filled voltage regulators are convenient for this purpose. No matter what precautions may be taken, however, the circuits described will ultimately drift because the characteristics of the valves will change as they age. Although this is likely to be a long-term trouble

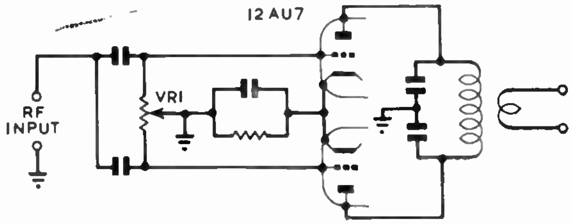


Fig. 10.14. Unreliable method of balancing (see text).

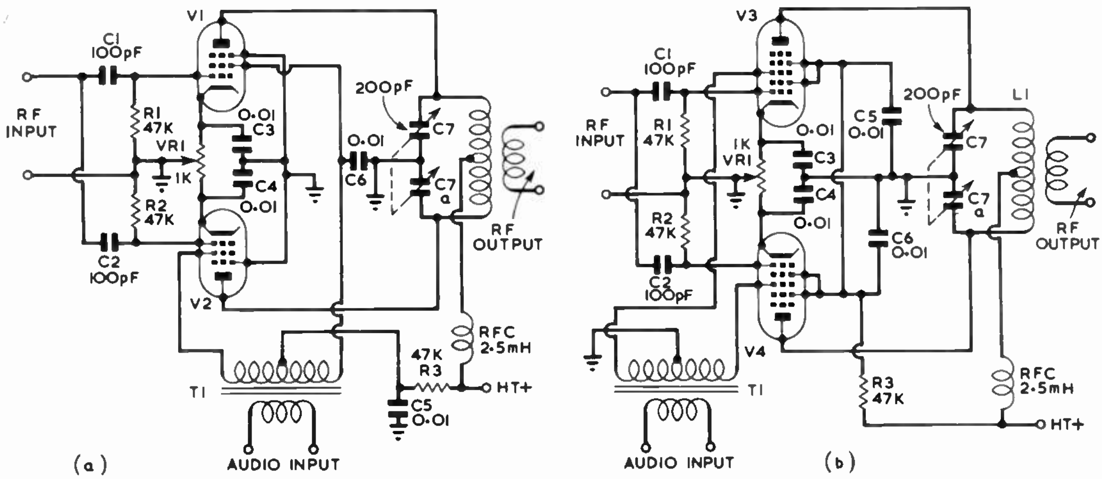


Fig. 10.13. Pentode (a) and pentagrid (b) balanced modulators. Almost any r.f. valve types can be used. V3 and V4 suppressor grids are normally at ground potential.

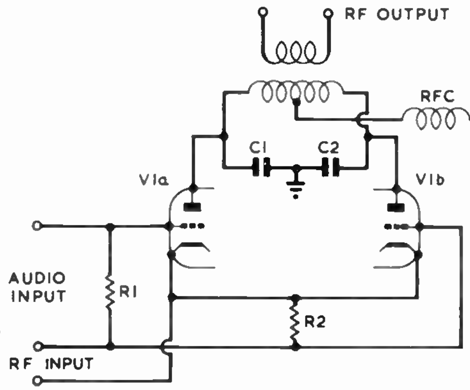


Fig. 10.15. Single-ended excitation of balanced modulator (see text).

which may be remedied by monitoring and re-adjustment, it can cause considerable inconvenience if the balancing controls are so positioned that the whole transmitter has to be taken out of its case to allow them to be reset.

It will be noted that the introduction of a.f. modulation in push-pull is recommended with all high impedance balanced modulators using conventional amplifier valves. Simplified circuits such as that of Fig. 10.15 have been designed in which modulation is applied to one of the valves only. The drawback of such a simplification is that the amount by which the carrier may be suppressed is only about half that obtainable with push-pull modulation; 20-25db represents creditable performance for the circuit of Fig. 10.15. For best results, it is essential to provide push-pull audio excitation.

Although there are many possible variations of the low impedance diode modulator, they all stem from the four diode ring modulator which has given the telephone companies such excellent service in landline s.s.b. work. The basic modulator shown in Fig. 10.16 is rarely used by amateurs but a number of derivatives have become popular. The diode balanced modulator requires impedance transformation in both the audio and r.f. circuits and may not be as convenient to incorporate in a transmitter as the high impedance type. Its long term stability is, however, claimed to outweigh all other considerations.

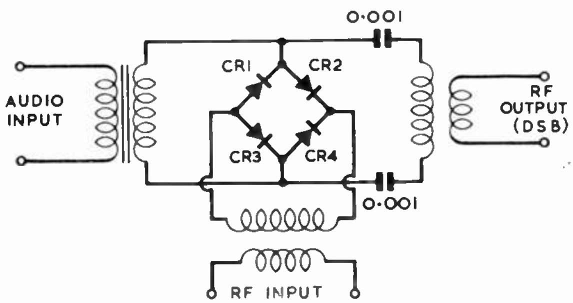


Fig. 10.16. Diode bridge modulator.

The shunt connected circuit shown in Fig. 10.17 is capable of excellent results, and is suitable for providing input to filters operating at intermediate or higher frequencies. The key to successful operation lies in the selection of a pair of diodes whose characteristics match closely and in the correct transformation of impedances at both audio and radio frequencies. Suitable matched diodes are available commercially. When installing them, an efficient heat shunt should be used to prevent the heat of the soldering iron causing the characteristics to change. In addition, the diodes should be located well clear of components which might generate enough heat to affect their performance. Many a mediocre balanced modulator owes its indifferent operation to the fact that it has been placed too close to a valve or a high wattage voltage-dropping resistor.

The selection of operating impedances is something of a compromise. The resistor R2 is included to prevent the carrier being shorted to earth through the audio transformer T1, and the input of the i.f., transformer must be modified as shown in Fig. 10.17 so that it will not act as a short circuit at audio frequencies.

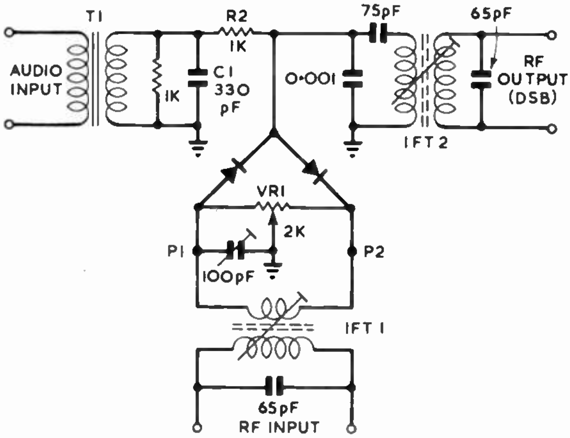


Fig. 10.17. Shunt-type diode balanced modulator.

Carrier input must be at low impedance; a standard i.f. transformer with the original secondary removed and replaced with a scramble winding of 50 to 100 turns tightly coupled against the primary will be found satisfactory. If the potentiometer VR1 does not alone give sufficient carrier suppression, it may be supplemented by a small variable capacitor of about 100 pF maximum value connected between earth and one or other of the points marked P1 and P2 in Fig. 10.17. If the balancing capacitor is needed, its correct placement and adjustment may quickly be determined by trial and error.

The application of modulation causes the bridge to become unbalanced at audio rate and double sideband energy to appear at the output. For minimum distortion, the carrier voltage should be about ten times as great as the peak audio voltage. Several volts of r.f. and a fraction of a volt of audio are the usual operating levels.

The series connected balanced modulator is shown in **Fig. 10.18**. Almost any audio transformer may be used, as long as it is bypassed at carrier frequency by C1, while a standard i.f. transformer will serve for IFT2. The same considerations as in the shunt-connected modulator apply to the carrier input transformer and to capacitative balancing. With either modulator the audio input may be provided by a cathode follower instead of a transformer and this arrangement is shown in **Fig. 10.19**.

Although the series and shunt balanced modulators are suitable for use at i.f. and r.f. to feed filter units, a pair of them may not be connected conveniently in parallel in the manner required for the generation of s.s.b. output by the phasing method. The modified ring circuit designed by D. E. Norgaard, W2KUJ, and reproduced in **Fig. 10.20** overcomes this difficulty. Almost all the commercial phasing-type transmitters made in the USA which employ diode balanced modulators make use of this arrangement. An ordinary loudspeaker transformer with a 15 ohm secondary will be suitable for T1, while L1 is a three or four turn link wound over the cold end of the carrier oscillator coil, and adjusted to give about 4 volts of r.f. across VR1. The push-pull tank circuit resonates at the desired output frequency. The 2·5 mH choke must not be omitted.

In describing balanced modulators, the convention of rating a circuit in terms of "so many decibels of carrier suppression" has been followed. This may be misleading

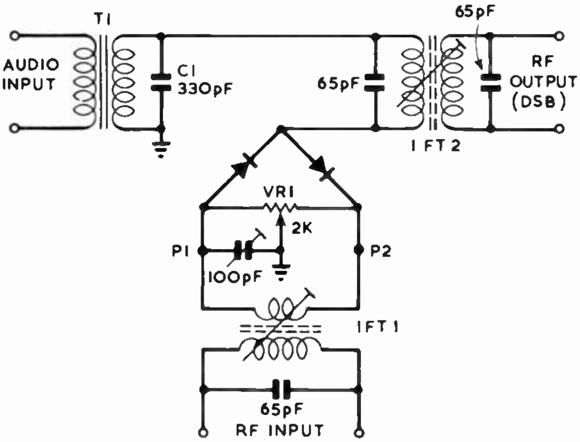


Fig. 10.18. Series-type balanced modulator.

to the uninitiated, because in fact the circuits work the other way round. To be strictly correct, the rating of a balanced modulator should be expressed as the number of decibels by which the output at maximum audio drive exceeds the minimum resting output in the absence of modulation. The point of making this distinction is to show that for the highest degree of carrier suppression any balanced modulator should be driven to its maximum undistorted output. The drive to the final stages of a transmitter should be controlled *after* the s.s.b. signal has been generated and not by reducing the audio input to the balanced modulator; if it is not, the carrier suppression of the system as a whole will suffer.

As the balance of both high and low impedance types is affected to a considerable extent by the amplitude of the

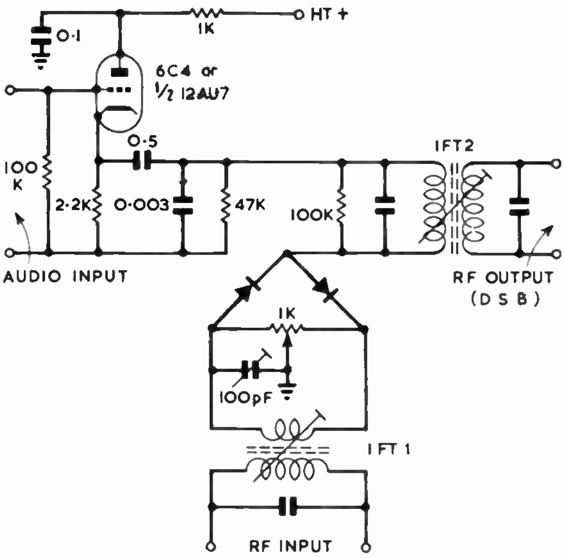


Fig. 10.19. Method of substituting an a.f. cathode follower in place of an audio transformer in a series-type diode balanced modulator.

applied r.f. carrier, excitation should be obtained from an oscillator of the highest stability. The voltage to the oscillator anode should, of course, be stabilized, as should the anode supply to the valves in the high impedance modulator.

SIDEBAND ATTENUATION

The double sideband signal generated by any of the balanced modulators described is of little value for communication purposes, so it has to be turned into s.s.b. by attenuating one of the sidebands. No matter what system may be used, the unwanted sideband is not eliminated completely; it is merely attenuated to the extent at which its nuisance effect becomes negligible. A filter attenuation of 30-35db has come to be regarded as the minimum acceptable standard. With care, suppression of 50db or more is attainable but it is debatable whether there is any practical advantage in striving after greater perfection.

The unwanted sideband may be attenuated either by phasing or filtering. The two methods are totally different in conception, and will be discussed in detail.

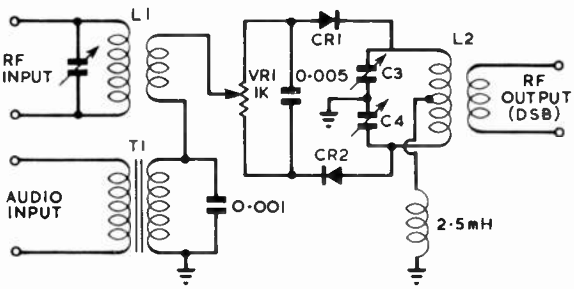


Fig. 10.20. Modified ring type balanced modulator. T1 is an audio transformer with low impedance secondary winding; a ratio between 1 : 8 and 1 : 12 is recommended. L1 is a three or four turn link wound over the cold end of the input tank inductor and adjusted to give 3 or 4 volts r.m.s. across its output. Any matched pair of diodes may be used for CR1 and CR2. The input and output r.f. circuits should resonate at the operating frequency.

The Phasing System

Although the filter method provides the classic way to generate a single sideband signal, the phasing system will be described first, because it was invented, pioneered and developed by amateurs. The system is based upon a mathematical theory which cannot be explained in simple language, but fortunately an understanding of a few basic principles is all that is required to enable the theory to be translated into practice.

A block diagram of a phasing type transmitter is shown in **Fig. 10.21**, from which will be seen that the output of an r.f. oscillator is fed into a network in which it is split into two separate components, equal in amplitude but differing in phase by 90°. Similarly, the output of an audio amplifier is split into two components of equal amplitude and 90° phase difference. One r.f. and one a.f. component are combined in each of two balanced modulators. The double sideband suppressed carrier energy from the two balanced modulators is fed into a common tank circuit. The relative phases of the sidebands produced by the two balanced modulators are such that one sideband is balanced out, while the other is reinforced. The resultant in the common tank circuit is an s.s.b. signal. The main advantages of a phasing exciter are that sideband suppression may be accomplished at the operating frequency, and that selection of the upper or lower sideband may be made by reversing the phase of the audio input to one of the balanced modulators. These facilities are denied to the user of the filter system.

If it were possible to arrange for absolute precision of phase shift in the r.f. and a.f. networks, and absolute equality in the amplitude of the outputs, the attenuation of the unwanted sideband would be infinite. In practice, perfection is impossible to achieve, and some degradation of performance is inevitable. Assuming that there is no error in the amplitude adjustment, a phase error of 1° in either the a.f. or the r.f. network will reduce the suppression to 40db, while an error of 2° will produce 35db and 3·5° will result in 30db suppression. If, on the other hand, phase adjustment is exact, a difference of amplitude between the two audio channels will similarly reduce the suppression. A difference between the two voltages of 1 per cent would give 45db attenuation, an error of 2 per cent would result in 40db, and 4 per cent 35db approximately. These figures are not given to discourage the intending constructor, but to stress the need for *high precision workmanship and adjustment* if a satisfactory phasing-type s.s.b. transmitter is to be produced.

There are many ways of obtaining r.f. phase shifts, but the two most satisfactory are those which use a pair of loosely coupled tuned circuits, and the so called low-Q network which uses a combination of resistance, inductance and capacitance. The former is shown diagrammatically in **Fig. 10.22** and is based on the property of coupled circuits which provides that if both are tuned to resonance the voltage induced in the secondary will lag behind that appearing across the driven primary by 90°. In practice it may prove difficult to tune both circuits to the precise point of resonance because of mutual interaction, but if this defect is encountered it may be easily overcome. The recommended method is to start by detuning the secondary completely, or to shunt it by a resistive load of about 200 ohms so that it cannot react upon the primary. The primary is tuned to resonance, using a valve voltmeter or other r.f. indicator, and is then detuned to the high frequency side until the voltage drops to 70 per cent of that indicated at resonance. The secondary is next brought to resonance (with its shunt removed) and finally detuned to the low frequency side until the indicated voltage drops to 70 per cent. The voltage across the two circuits will then differ in phase by precisely 90°. Because of losses, it is unlikely that the voltage appearing across the secondary will be exactly the same amplitude as that across the primary; it is therefore inadvisable to use capacitive coupling from the hot ends of the windings. Link coupling, as shown in Fig. 10.22, is preferable because it allows compensation to be made for inequalities in amplitude.

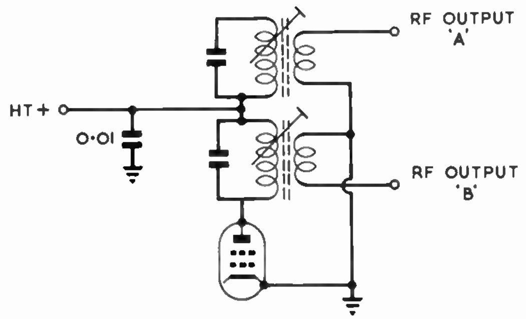


Fig. 10.22. Method of obtaining 90° phase shift by coupled tuned circuits.

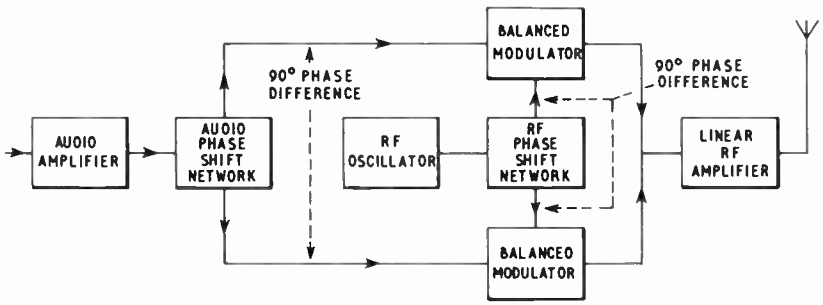


Fig. 10.21. Phasing type transmitter.

The early amateur phasing transmitters were designed for fundamental-frequency operation, driven directly from an existing v.f.o. tuning the 80m band, and used a low-Q phase shift network. The low-Q circuit has the ability to maintain the required 90° phase shift over a small frequency range, and this made the network suitable for use at the operating frequency in single band exciters designed to cover only a portion of the chosen band.

Today, with the great increase in s.s.b. activity, it is necessary to be able to change frequency over a range of 200 kc/s or more. Under these conditions the r.f. phase shift network would be quite incapable of maintaining the required accuracy of phase shift and the available sideband suppression would deteriorate to a point at which the exciter was virtually radiating a double sideband signal. It is true that the plain r.c. network will provide a phase shift that is independent of frequency. Unfortunately experience over the years indicates that this simple method is not capable of providing the required sideband suppression. Additionally, a change of frequency into any type of r.f. phase shift network will cause a change in output amplitude ratio, and this in turn will unbalance the modulator and severely degrade the carrier suppression

If amateurs are to enjoy the full benefits of s.s.b. working and reduce adjacent channel splatter and heterodyne interference to a low level on the amateur bands, a sideband suppression of 30-35db and a carrier suppression of 50db should be considered the minimum acceptable standard.

Any operating method that is fundamentally incapable of maintaining this standard should not be used on the amateur bands. For this reason, the fundamental type of phasing unit is not recommended. For acceptable results, the r.f. phase shift must be operated at a fixed frequency outside the amateur bands. The s.s.b. output from the balanced modulator is then heterodyned to the required bands by means of an external v.f.o. This method will be discussed in greater detail later in this chapter.

Audio Phase Shift Network

Although it is quite practicable to construct an audio phase shift network capable of covering from 50 to 5000 c/s, the results scarcely justify the high cost involved. Experiments have shown conclusively that frequencies below 300 and above 3000 c/s contribute little to the intelligence conveyed by the human voice, and that under conditions of high ambient noise there is a decided advantage in accentuating the higher frequency speech components and progressively attenuating those components which lie below 500 or 600 c/s. The naturalness of the voice under this kind of treatment must necessarily suffer, but to a lesser degree than might be expected. Telephone lines have a frequency response no better than that mentioned and their quality is adequate for normal aural communication.

A straightforward circuit giving exactly 90° phase shift over the total audio spectrum is impossible to realize, but by restricting the range to the minimum required for satisfactory communication, a tolerably good network may be constructed from simple fixed capacitors and resistors. The voltage at one of the output terminals leads the applied a.f. voltage by an amount which varies according to the frequency, while the output at the other terminal lags behind the applied voltage. The values of the components may be chosen so that the total phase shift between both output terminals approximates 90°.

Fig. 10.23 shows a network particularly suitable for amateur use. This is similar in configuration to that used by W2KUJ in his famous "S.S.B. Jr" which popularized single sideband in the USA, but the circuit values have been modified to enable more easily obtainable components to be used. The network operates over the range 300–3000 c/s, and if it is constructed of components of 1 per cent tolerance the maximum deviation from the ideal phase shift of 90° will be within 1·5°. Sideband suppression is therefore better than 35db at the worst frequency within the range. This network should be fed from a source impedance of 500 ohms, and the two a.f. input voltages should have a precise amplitude ratio (measured to earth) of 2 : 7. This is obtained by the position of the slider of the balancing potentiometer VR1; it therefore follows that this slider will be some way off the central position. The lower of the two voltages should occur at the input terminal common to the two 430 pF capacitors, C2 and C4. The output terminals should be connected to a load of infinite impedance such as the grids of a pair of class A amplifier valves. (This audio phase shift network may be obtained commercially as the Barker & Williamson Type 2Q4 Phase Shift Network, with all components sealed in a metal valve envelope designed to plug into a standard octal valveholder.)

As the audio phase shift network is the heart of a phasing exciter, no effort should be spared in constructing it as closely as possible to specification. Close tolerance silvered mica capacitors and high stability resistors should be used exclusively.

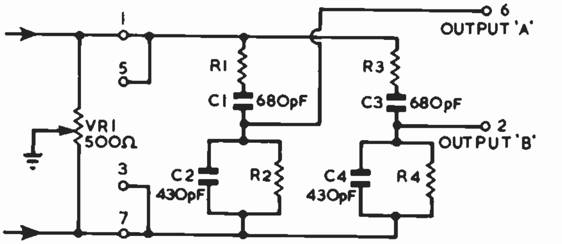


Fig. 10.23. "S.S.B. Jr." passive a.f. phase shift network. C1, C3 680 pF. C2, C4 430 pF. R1 487,500 ohms (470K + 15K + 2·2 K in series); R2 770,000 ohms (500 K + 270 K in series); R3 125,000 ohms (100 K + 15 K + 10 K in series); R4 198,000 ohms (150 K + 47 K in series); VR1 500 ohms. All capacitors are silver mica 1% tolerance. All resistors are high-stability ± 1%, of half watt rating. The input and output connection numbers relate to the octal base pin numbers of the B & W 2Q4 P.S. network.

If the speech amplifier used to drive the network described does not have its response restricted to the range over which the network is effective, objectionable distortion will result. An adequate degree of roll-off at low frequencies may be obtained by using low value coupling capacitors in the speech stages preceding the network. Insufficient high frequency attenuation will not necessarily make the signal as difficult to tune and read as poor low frequency cut-off, but it will cause splatter and interference to users of adjacent channels. This may be minimized by restricting the high frequency response of the audio amplifier in the manner shown in Chapter 9 (*Modulation*).

A stage of class A amplification must be interposed between each of the output channels of the phase shift network and its associated balanced modulator for three reasons: to provide amplification, to offer the correct terminating impedance, and to isolate the network from the somewhat variable load presented by the balanced modulator.

The necessity of building these two amplifiers as really high fidelity devices is frequently overlooked, but it is vitally important. Unless the frequency response is excellent, the amplifiers may introduce phase-shifts on their own account and nullify the good work done by the phase shift network.

If *RC* coupling is employed, large-value capacitors are essential, while audio transformers should be of proven high quality and wide frequency response. Shunt feed is recommended with transformer coupling to obviate the ill-effects of d.c. current flowing through the primary. Finally, the valves themselves should be operated on the strictly linear part of their characteristic curves and kept well within their ratings.

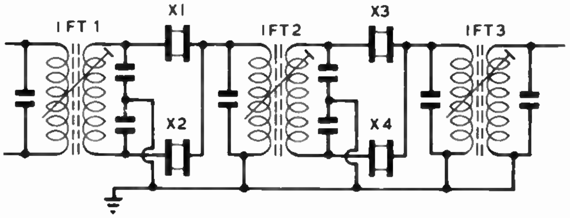


Fig. 10.25. Double section half-lattice filter. X1 and X3 same frequency. X2 and X4 same frequency. Recommended crystal spacing is given in Table 1.

The Filter System

Since the objective is to transmit only a single sideband, it is necessary to select the desired sideband and suppress the undesired sideband. This is possible because the modulating wave is restricted to a band of audio frequencies separated from the carrier by an appreciable amount. This relationship is shown in the diagram of Fig. 10.5. Removing the unwanted sideband by the use of selective filters has the advantage of simplicity and good stability. The unwanted sideband suppression is determined by the attenuation of the sideband selecting filter, and the stability of this suppression is determined by the stability of the elements used in constructing the filter. This stability can be quite high because it is possible to use materials that have a very low temperature coefficient of expansion. Two commonly used materials are quartz crystal plates and small metal plates.

The filter system, because of its proven long term stability, has become the most popular method used by amateurs. At present three types of selective sideband filters are in common use:

(i) the medium frequency crystal filter,

(ii) the high frequency crystal filter and,

(iii) the low frequency mechanical filter.

These three methods will be described in detail.

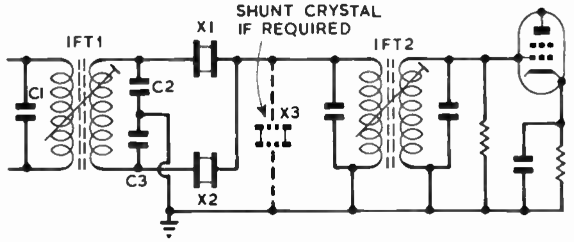


Fig. 10.24. Single section half-lattice filter. To obtain the centre tap IFT1 is modified by replacing the original resonating capacitor by two capacitors of double the value (C2 = C3 = 2C1).

M.F. Crystal Filters

The crystal sideband filter is attractive to the amateur because it can be home constructed and the response characteristics and shape factor are under the constructors control. Additionally, suitable crystals in the range 400–500 kc/s have been available on the surplus market at low cost in the FT241 series of 54th and 72nd harmonic crystals. All crystals whose marked frequency commences with the figure 2 are 54th harmonic types, and all crystals whose marked frequency commences with the figure 3 are 72nd harmonic types. In each case the fundamental frequency of the quartz plate is the frequency marked on the box in Mc/s, divided by the harmonic series (i.e. 54 or 72).*

Experience has shown that an audio frequency range of 300–3000 c/s is satisfactory for voice communication and gives acceptable speech quality. The filter passband is therefore required to be 2700 c/s, and in the perfect filter the " slope " of the sides of the filter response curve would be vertical so that the bandwidth 60db down would also be 2700 c/s. In practice this ideal is not practicable and the sides of the filter response slope outward so that the bandwidth 60db down is greater than the bandwidth at the 6db points. The ratio, 6db bandwidth to 60db bandwidth, is the *shape factor* of the filter. A shape factor of two is a good figure to aim for, and if the filter can be made with a steeper response than this, so much the better—the sideband suppression will be further improved.

The circuit of a single half-lattice filter is shown in **Fig. 10.24.** For improved sideband suppression half-lattice sections may be connected in cascade as shown in **Fig. 10.25.**

The optimum crystal spacing for an audio bandwidth of 300–3000 c/s is affected by the steepness of the passband *skirts*, (i.e. approximately 1·85 kc/s for a single half-lattice section, 2·2 kc/s for two sections, and 2·4 kc/s for three sections), and also by the inclusion or otherwise of a shunt crystal which may be used to steepen the response on the carrier side. A simple half-lattice filter will give an unwanted sideband suppression of approximately 30db, and if a better

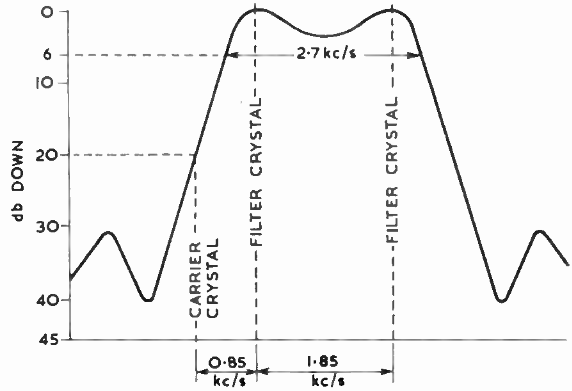


Fig. 10.26. Response curve of a single half-lattice filter with neutralizing. Note the side lobes just below the 30db level. The steepness of the response can be improved by a small neutralising capacitor of 2-3 pf (two lengths of p.v.c. insulated 22 s.w.g. connecting wire twisted together for half an inch) across the pins of the highest frequency crystal in each half-lattice arm. Too much capacity may bring up the side lobes to an unacceptable level.

* Tables showing the frequencies of all FT241 type crystals commonly used in s.s.b. filters are given in the *Radio Data Reference Book*, published by RSGB.

performance than this is required it will be necessary to use two or three filter sections. As a guide, the response of a single half-lattice filter is shown in **Fig. 10.26** and the response of the same filter with an additional shunt crystal is shown in **Fig. 10.27**. As a further guide, the expected performance with different filter configurations is given in **Table 10.1**.

All FT241 crystals should be checked for frequency and activity by placing the crystal in series with the output from a BC221 frequency meter and the input to a diode probe valve voltmeter used on the 10 volt range. The BC221 should be *slowly* tuned across the required frequency range and at some setting the valve voltmeter pointer will swing over. This is the series resonant frequency of the crystal and if the activity is satisfactory the valve voltmeter will read about half scale. Any crystal giving less than 3 or 4 volts should be discarded. If the filter comprises two or three sections, those crystals on either side of the passband centre should be matched in frequency within a few cycles of each other, either by edge grinding or by plating.

The only satisfactory method of edge grinding is to hold the crystal plate (without dismounting from its support wires) in a jig so that it cannot possibly move under any circumstances, and then apply the grinding medium to it in the form of a rigid flat surface such as a 3 in. × 1 in. × ½ in. Carborundum slip stone of No. 280 grit, obtainable from a local hardware dealer. The frequency may be raised a few hundred cycles by grinding the top edge.

Copper plating may be used to lower the frequency. A standard plating solution may be made by adding 15 gm. of copper sulphate, 5 c.c. of concentrated sulphuric acid and 5 c.c. of commercial alcohol to 100 c.c. of distilled water. The solution should be poured into a glass vessel in which is placed a piece of clean copper wire bent so that it clamps the edge of the vessel and extends about 1 in. into the liquid. The wire should be connected in series with a 330 ohm resistor to the positive terminal of a 1·5 volt dry cell. Both pins of the crystal are then connected together and wired to the negative side of the battery. When the crystal is dipped into the solution, a copper deposit will be formed which will lower its frequency. The exact variation may be determined by trial and measurement; if the process is accidentally carried too far, it may be reversed by changing the polarity of the dry cell and removing some of the copper. The process is not necessarily reversible with *all* crystals so caution is recommended.

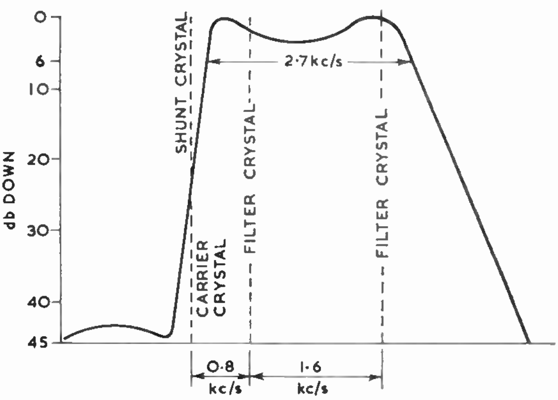


Fig. 10.27. Response curve of a single half lattice filter with shunt crystal. Note how the shunt crystal steepens the passband on the carrier side and improves the sideband suppression.

When the frequency of the crystal has been increased by grinding, or lowered by plating, the unit should be thoroughly washed under running warm water, or alternatively washed in pure carbon tetrachloride, before being returned to service.

The most suitable type of i.f. transformers to use for coupling are the medium Q, high L/C ratio types; that is a Q of around 60 and a resonating capacity value of not more than 100 pF. The Maxi-Q miniature i.f.t. Type IFT.11/465 are very satisfactory, as also the *standard* miniature i.f.t's. made by Electroniques (not the pot core types). After alignment by adjusting all transformer dust cores for maximum response at the centre passband frequency, the filter response should be plotted and the bandwidth measured at the 6 and 60db points to determine the shape factor that has been achieved. Normally the carrier crystal is placed at a frequency that is 20–25db down on the filter response curve.

It should be appreciated, however, that its positioning is a compromise and it should finally be adjusted to obtain the best balance of audio quality determined by on-the-air reports.

Before embarking on the construction of a sideband filter it is most important to realise that surplus FT241 crystals are

TABLE 10.1

	Single half-lattice filter with neutralizing	Single half-lattice with shunt crystal No neutralizing	Two half-lattice with 2 shunt crystals. No neutralizing	Three half-lattice Symmetrical receiver filter
Crystal spacing for 2·7 to 3·0 kc/s bandwidth	1·85 kc/s	1·6 kc/s	2·2 kc/s	2·2 to 2·4 kc/s
Carrier crystal spacing (Fig. 10.27)	0·85 kc/s	0·8 kc/s	0·25 to 0·4 kc/s	0·35 to 0·4 kc/s
Sideband Suppression	30db	35db	45 to 55db	60 to 72db
Type of response	Symmetrical	Asymmetrical	Asymmetrical	Symmetrical. Bandwidth at 6db points = 2·5 kc/s.

more than 20 years old, have deteriorated with time and may be several hundred cycles—in some cases more than 1 kc/s—off their original frequency. The Q of the crystal may be very low and in some cases the crystal may fail to oscillate at all.

Modern gold plated *AT* cut crystals are supplied in hermetically sealed B7G holders and have much better stability and output and may be obtained as current equipment ground and calibrated to within 0·01 per cent of specified frequency; alternatively the Quartz Crystal Co. Ltd. will supply a complete set including the carrier crystal, for either a single section or a double section filter. Newcomers to s.s.b. without previous experience of crystal manipulation and filter alignment are strongly advised that the initial additional cost of current production crystals is, in the long run, a worthwhile investment.

H.F. Crystal Filters

Many amateurs are natural experimenters and prefer, if at all possible, to make their own items of equipment. This applies particularly when there is a ready supply of low cost surplus " raw material." It is therefore not surprising that there has been increasing interest in the possibility of using the range of FT243 high frequency crystals in a bandpass filter with characteristics suitable for single sideband application. Experience has shown that it is in fact practicable to use these crystals and meet the requirements of the amateur s.s.b. operator in regard to a filter giving a bandwidth of approximately 2·7 kc/s and a shape factor of two, with simple circuitry, the application of a little common sense and initiative, and the use of test equipment likely to be used by an amateur.

The main advantage of the crystal filter is the exceptionally high Q and stability of the quartz plates and these characteristics make it possible to obtain the required selectivity at high intermediate frequencies, avoiding the need for the usual double conversion that is generally necessary to obtain the required degree of image rejection.

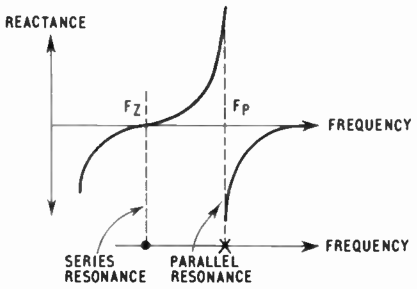


Fig. 10.28. Reactance characteristics of a quartz crystal. The horizontal arrows denote an increasing frequency.

It is well known that a quartz crystal has two resonant points very close together: these are the series resonance (the " zero " of impedance) and the parallel resonance (the " pole " of impedance) and this change of reactance or impedance is shown in Fig. 10.28.

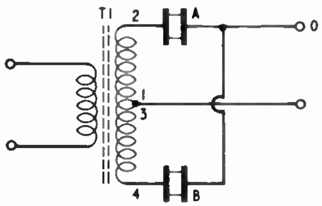


Fig. 10.29. Basic single section half-lattice crystal filter. For optimum results the two halves of T1 must have very tight coupling.

The circuit of a simple one-section half-lattice filter is given in **Fig. 10.29** and it will be seen that the two crystals and the two halves of the inductance T1 form the legs of a bridge. Provided that the voltage across the coil from 1 to 2 is exactly the same as the voltage from 3 to 4, and provided also that the impedances of the two crystals A and B are equal there will be no voltage at the common connection (point 0).

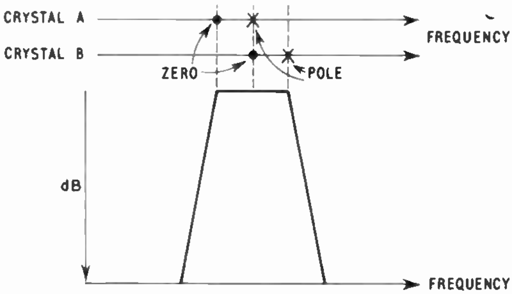


Fig. 10.30. Diagram showing the theoretical passband of a half lattice crystal filter.

For the requirement of a bandpass filter, crystals A and B are chosen to be different in frequency. Assuming the input frequency is at the zero (series resonance) frequency of crystal A the impedance balance of A and B is spoiled and there will be an output voltage between point 0 and the centre of the coil; this will also occur if the input frequency is at the pole (parallel resonance) frequency of crystal A. At the appropriate input frequencies the same thing will happen for crystal B, only the unbalance will be in the opposite direction. From this it follows that the filter passband will be as wide as the spacing of all the poles and zeros. If the series resonant frequency of crystal B is arranged to coincide with the parallel resonant frequency of crystal A this will theoretically give a perfectly flat passband of twice the pole zero spacing of each crystal; see **Fig. 10.30**. Fortunately for the constructor the surplus range of FT243 crystals have a measured range of pole zero frequencies suitable for the requirement. Two of these crystals spaced approximately 2 kc/s apart will give a satisfactory single sideband bandwidth. A further improvement in the steepness of the passband skirts and therefore better unwanted sideband rejection is obtained by cascading two half-lattice sections in a simple back-to-back circuit.

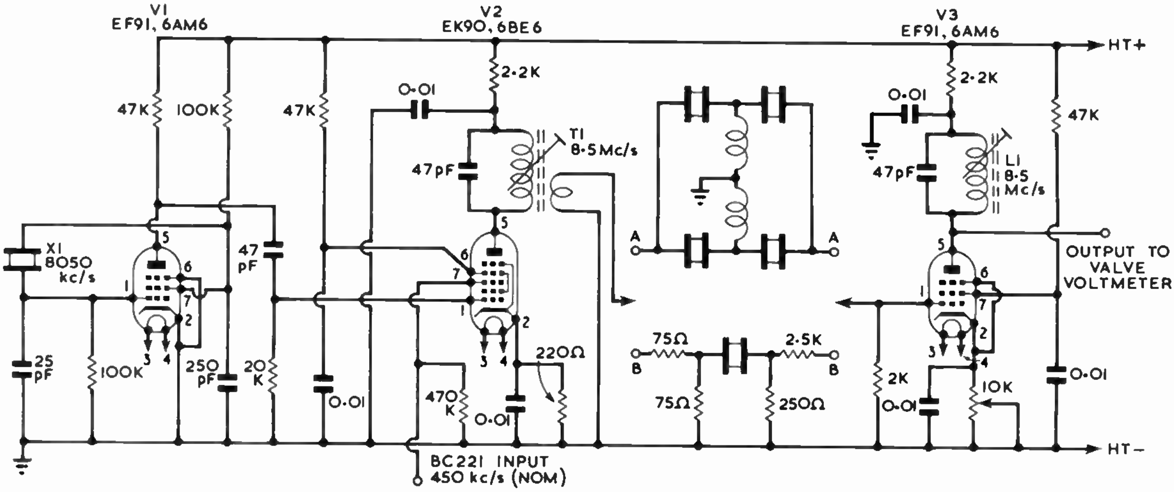


Fig. 10.31. This is the test circuit used by G2DAF for plotting filter pass bands and crystal frequency measurement. The BC221 frequency meter operates on its normal 430 to 470 kc/s range, and beats against the 8050 kc/s crystal controlled oscillator to give a precision output over the range 8480 to 8520 kc/s. For pole zero frequency measurement the arrowed connection points are taken to the B terminals, and for filter passband plotting they are taken to terminals A. The potentiometer in the cathode circuit of V3 enables amplifier gain to be set to give a convenient scale deflection on the valve voltmeter at the point of maximum filter response i.e. 0 db.

It is an impossible undertaking to attempt to measure the series and parallel resonances of high frequency crystals with the usual workshop signal generator—the tuning is far too coarse. The only satisfactory procedure is to construct a test rig so that the variable signal source is a frequency meter such as the BC221 used on its normal l.f. ranges. A suitable test rig is shown in **Fig. 10.31.** It will be seen that the BC221 beats against a crystal controlled oscillator to produce the required frequency at the converter valve anode. The crystal to be tested is plugged into the holder whose input and output has a resistive padding network to avoid measurement error due to stray circuit capacity (stray circuit capacity has no effect on the series resonance but will affect the parallel resonance). At the parallel resonance the crystal is offering a high impedance and the output voltage would be too small to be measured on a workshop valve voltmeter. Accordingly a class A amplifier is used, and the valve voltmeter probe is connected to the valve anode. This gives ample voltage to enable the pole and the zero frequencies to be accurately determined, and also enables the test rig to be used to plot the response of the complete filter, using the alternative connections shown.

Initially it is wise to purchase at least eight crystals—preferably 12—all of the same channel number. Using the test rig, measure the pole and zero of each crystal and mark the frequencies in pencil on the side of the crystal base (the crystal zero frequency coincides with a sharp rise in the valve voltmeter reading, and the pole frequency with a sharp dip). Now select four crystals with the same—or most nearly the same—pole zero spacing. Take two of these crystals and match them within 10 c/s of the same series resonant frequency by etching the lower up to the higher with ammonium bifluoride*. The two remaining crystals are also matched in the same way, but on a series resonant frequency that is approximately 2 kc/s higher. As 2 kc/s is a very small proportion of the crystal frequency the etching solution will work very fast and it is advisable to use one part of a normal saturated solution to one or two parts of water.

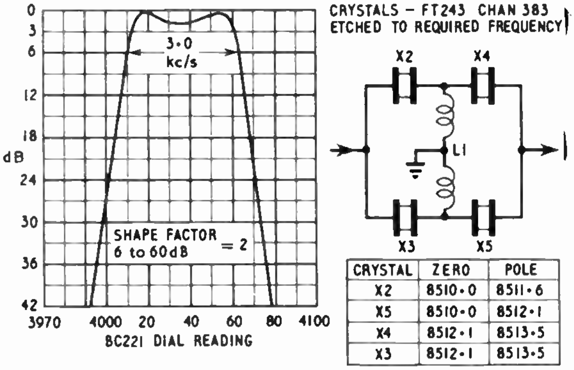
Finally, the filter coil is constructed by taking a length of 22 s.w.g. p.v.c.-insulated connecting wire, doubled back on itself to form two parallel wires. This is then wound on a ferrite ring core to form nine double turns (total 18 turns) and the inner of one winding connected to the outer of the other to form a bifilar winding with the junction the centre tap. The main requirement of the inductance is very tight coupling between each half, together with a perfect electrical balance. In practice the grade of ferrite material does not appear to be critical. Filters have been successfully built using the ring from a Mullard LA4 pot core assembly (35mm outside diameter) and also the ring from a mains suppressor filter choke (1⅜ in. outside diameter) obtained from surplus sources.

To give some guidance as to the characteristics that may be expected, the plot of the filter passband and the pole and zero frequencies of a typical experimental filter is given in **Fig. 10.32.**

The carrier crystal frequency is determined by plotting the filter passband and marking the 20db down points. One of the remaining FT243 crystals is etched so that its *parallel* resonance frequency is at this frequency. Finally when the transmitter is completed and tested on the air the carrier crystal can be pulled by means of a 50 pF trimmer across the oscillator grid circuit to obtain the best balance of voice quality.

FT243 crystals are mounted in ½ in. pin spacing holders, and are available over a wide range of frequencies in 25 kc/s

* This is highly corrosive and must be treated with extreme caution.



CRYSTAL	ZERO	POLE
X2	8510·0	8511·6
X5	8510·0	8512·1
X4	8512·1	8513·5
X3	8512·1	8513·5

Fig. 10.32. The high frequency filter passband and pole zero frequencies of the filter in use at G2DAF. It can be seen that crystal X5 is the odd one out, with a greater pole-zero spacing than the others. This causes the slight asymmetry of the curve.

steps from 5700 kc/s to 8650 kc/s, and in 33·333 kc/s steps from 5706·666 kc/s to 8340 kc/s. In general all crystals throughout the range are suitable for h.f. filter construction; there is however one notable point. These crystals were made by a large number of different manufacturers for the USA Services and the "cut" of the quartz plate is not always the same (i.e. certain crystals may have a pole zero spacing that is much greater than the average figure of about 2 kc/s; such crystals should be discarded as not suitable for bandpass filter use).

Commercially-made h.f. filters are now available from the following manufacturers: McCoy Electronics Co., Holly Springs, Pa., USA, and The Quartz Crystal Co. Ltd., New Malden, Surrey. The latter manufacturer will also supply current production h.f. filter crystals ground to specified frequency, together with matching carrier crystal.

Mechanical Filters

Three types of mechanical filters are available in the UK at competitive prices that put them within the price range of many sideband workers. These are the Collins F455 FA-21 filter with a 6db bandwidth of 2·1 kc/s and the Kokusai MF455-10K and MF455-15K with 6db bandwidths of 2·0 and 3·0 kc/s respectively. The mechanical filter is a mechanically resonant device which receives electrical energy, converts it into mechanical vibration, then converts the mechanical energy back into electrical energy at the output. The mechanical filter consists basically of four elements:

(i) an input transducer which converts the electrical input into mechanical oscillations,

(ii) metal discs which are mechanically resonant,

(iii) coupling rods which couple the metal discs, and

(iv) an output transducer which converts the mechanical oscillations back into electrical oscillations.

Fig. 10.33 shows the elements of the mechanical filter, and **Fig. 10.34** shows the electrical analogy of the mechanical filter. In the electrical analogy the series resonant circuits L1 C1 represent the metal discs, the coupling capacitors C2 represent the coupling rods, and the input and output resistances *R* represent the matching mechanical loads.

The transducer, which converts electrical energy into mechanical energy and vice versa, may be either a magnetostrictive device or an electrostrictive device. The magnetostrictive transducer is based on the principle that certain materials elongate or shorten when in the presence of a magnetic field. If an electrical signal is sent through a coil which contains the magnetostrictive material as the core, the electrical oscillation will be converted into mechanical oscillations. The mechanical oscillation can then be used to drive the mechanical elements of the filter. The electrostrictive transducer is based on the principle that certain materials, such as piezolectric crystals, will compress when subjected to an electric current. The transducer not only converts electrical energy into mechanical energy and vice versa; it also provides the correct termination for the mechanical network.

In practice, filters between 50 kc/s and 600 kc/s can be manufactured. Since each disc represents a series resonant circuit, increasing the number of discs will increase the skirt selectivity of the filter. Manufacturing technique at present limits the number of discs to eight or nine in a mechanical filter. A six-disc filter has a shape factor of approximately 2·2, a seven-disc filter a shape factor of approximately 1·85, a nine-disc filter a shape factor of approximately 1·5.

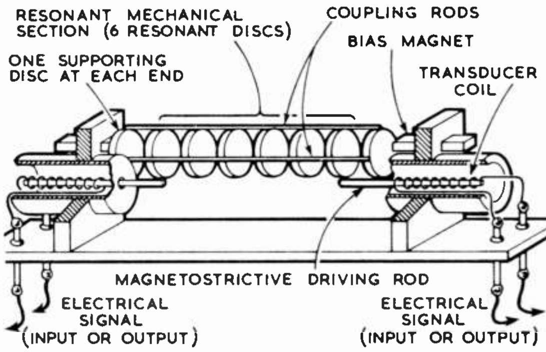


Fig. 10.33. Constructional details of a mechanical filter.

The Collins Mechanical Filter

Collins mechanical filters have been in use for a number of years and are available in the range from 60 kc/s to 500 kc/s with 6db bandwidths from 0·5 kc/s to 35·0 kc/s—in fact Bulletin 1007 obtainable from Collins Radio Co. of England Ltd. lists no fewer than 92 different types. The Type F455 FA-21 is designed specifically for the amateur market with a nominal 6db bandwidth of 2·1 kc/s and a 60db maximum bandwidth of 5·3 kc/s providing a shape factor of just over 2·5 : 1. It should be noted that although the shape factor is not so good as the more expensive 3·1 kc/s filter (shape factor 2·1) the slope is marginally better.

The F455 FA-21 filter is fitted in a rectangular case 2½ in. long, just over ½ in. wide, and ½ in. high and is intended for horizontal mounting. The centre frequency is 455 kc/s nominal; 6db bandwidth 2·1 kc/s nominal; 60db bandwidth 5·3 kc/s maximum; passband ripple 3db maximum; transfer impedance 5 K ohms; resonating capacity 130 pF ± 5 pF;

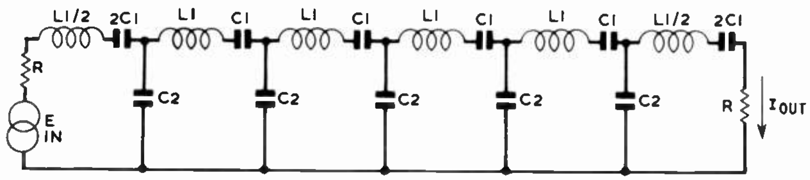


Fig. 10.34. Electrical analogy of a mechanical filter.

transmission loss 9·5db and spurious response attenuation (405–505 kc/s) 60db minimum. The resonating discs are driven by magnetostrictive transducers employing polarized biasing magnets; for this reason it is not permissible to have any d.c. at all flowing through the coil, and shunt feed to the valve anode is essential. The two transducer coils are identical, are isolated from the case, and are balanced to earth; either coil may therefore be arbitrarily designated as input or output and the filter can be coupled to the two anodes of a balanced modulator, or may be used to feed the two grids of a following balanced mixer.

Installation is simple, but it is important to provide screening between the input and output connections (a small cross screen underneath the chassis). The filter case must be earthed. Drive level should be kept below 2 volts r.m.s. For best performance the input and output transducer coils should be carefully resonated at the filter centre frequency. The two basic circuits (*a*) filter following a high impedance load such as the mixer valve in a receiver, and (*b*) following a low impedance load such as the modulator in a transmitter, are shown in **Fig. 10.35.**

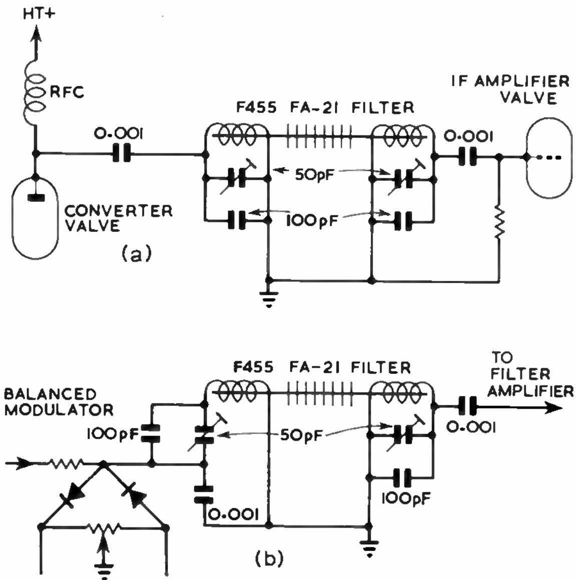


Fig. 10.35. Basic circuits for the Collins mechanical filter type F455 FA-21 showing at (a) a parallel resonated transducer coil following a high impedance load, and at (b) a series resonated transducer coil following a low impedance load. The resonating capacitors are adjusted for maximum signal through the filter at the centre passband frequency.

The positioning of the carrier crystal frequency is usually stated by the manufacturer. If it is not, the best procedure is to use a BC221 frequency meter as a variable carrier oscillator and get reports from other stations until the best balance of audio quality is obtained. The frequency is then obtained from the dial reading of the BC221 and a crystal

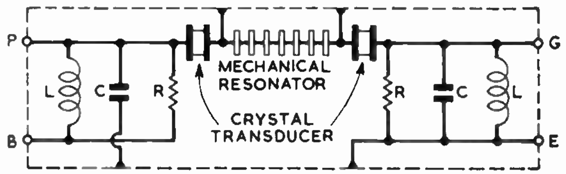


Fig. 10.36. Kokusai mechanical filter theoretical circuit diagram. With this filter it is permissible to have d.c. flowing through the transducer coil L, (i.e. series fed h.t. to a mixer anode). Accordingly the "B" terminal is isolated from the screening can and it is important that the filter is wired into the circuit the right way round.

ordered from the supplier, stipulating that the crystal is required for oscillation on the parallel resonance mode with 30 pF shunt capacity. Alternatively, an FT241 crystal can be edge ground or plated so that it oscillates in situ on the required frequency. The correct position will usually be found to be 20–30db down the skirt response; however the narrower the filter the more critical the positioning becomes. Usually practical experiment is preferable to an arbitrary selection.

The Kokusai Mechanical Filter

The Kokusai filter is available with a nominal centre frequency of 455 kc/s, with 6db bandwidths of either 2·0 kc/s or 3·0 kc/s. The case style is a spun aluminium cylindrical can approximately 2⅝ in. high and 1⅜ in. diameter with two threaded mounting studs and four insulated feed-through connecting tags. Mounting is intended to be vertical with the connecting tags going through the chassis.

This filter has the disc resonators excited by quartz crystal transducers; the circuit arrangement is shown in **Fig. 10.36.** The general specification of the two filters is as follows: MF-455-10K, 6db bandwidth 2·0 kc/s nominal; 60db bandwidth less than 7·0 kc/s. MF-455-15K, 6db bandwidth 3·0 kc/s nominal; 60db bandwidth less than 9·0 kc/s (in practice the filters stocked in the UK are generally better than the manufacturer's quoted figures, the 60db bandwidth rarely being greater than twice the 6db bandwidth, i.e. a shape factor of 2 or better); maximum deviation from centre frequency $\pm$ 0·8 kc/s; Passband ripple less than 4db; Temperature range 0°C to 70°C.

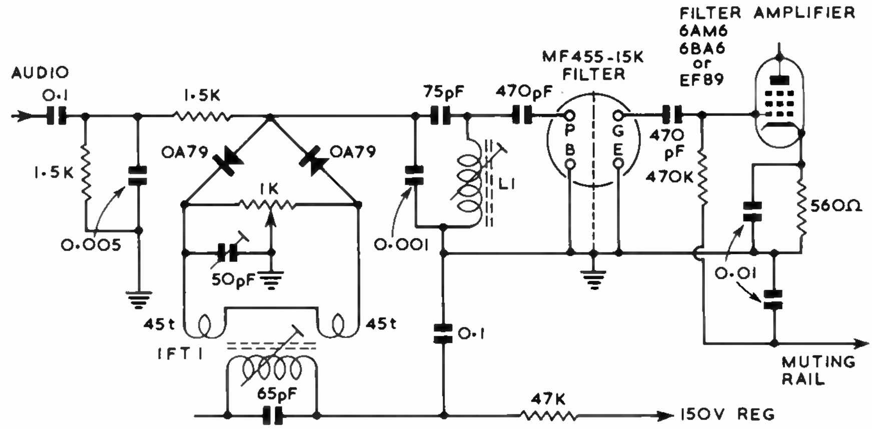


Fig. 10.37. Circuit details showing the method of using the Kokusai mechanical filter following a low impedance diode modulator. L1 is the primary winding of a standard Maxi-Q IFT.11-465 i.f. transformer with the original 65 pF resonating capacitor removed and replaced by the 75 pF and 0·001 uF capacitor shown. (The unwanted secondary pie is cut away with a sharp knife.) Feed a 1·5 kc/s tone into the microphone socket and resonate L1 for the maximum output at the anode of the filter amplifier valve.

The manufacturer does not give any values for input or output terminal impedance but does state " Both input and output should be connected to high impedances. Since mechanical filters tend to be rather sensitive to capacitances across both input and output terminals care should be exercised not to lower the impedance by stray capacity." It is clear from this that the Kokusai filter is designed to operate into relatively high impedance input and output loads. If the filter is connected directly across the output of a diode modulator presenting a low impedance load of a few hundred ohms, the transducer coil would be heavily damped, there would be loss of signal and a possible change in the filter passband characteristics. This difficulty can be overcome by using a series resonated tuned circuit, as an impedance step-up transformer, between the diode modulator and the filter transducer coil as shown in **Fig. 10.37**.

Because of its small physical size and ease of mounting, this filter is particularly suitable for incorporation in the older communication receivers such as the AR88, HRO and CR100, etc., and it is relatively easy to arrange to switch the filter in or out of circuit so that the normal selectivity can be used for a.m. and the filter for s.s.b. or c.w. A circuit suitable for most receivers is shown in **Fig. 10.38**.

The nominal centre passband frequency is 455·0 kc/s, but because of manufacturing tolerances the actual centre frequency may vary plus or minus 0·8 kc/s. For this reason, each filter is packed with a data sheet giving the filter serial number and the measured bandwidth at the 6db and 60db points for the filter concerned. Each bandwidth is given as plus or minus *X* kc/s relative to the design centre frequency of 455·0 kc/s—not relative to the actual filter centre frequency. To illustrate this more clearly, an example of the data sheet issued by the manufacturer is shown in **Fig. 10.39**. It will be noted that the total passband width at the 6db point is 3·2 kc/s but that this is quoted as two frequencies—each plus or minus relative to 455·0 kc/s and above this is a figure (in this case + 0·4 kc/s) indicating that the actual passband centre frequency is 455·0 kc/s + 0·4 kc/s = 455·4 kc/s. At the bottom of the table the bandwidth is given at the 60db points—again as two frequencies relative to 455·0 kc/s; these two figures added together give the total passband width at 60db down. The ratio of the total bandwidth at 6db to the total bandwidth at 60db is the shape factor of the filter. In this case, it is 5·9/3·2 = 1·9 approximately.

From the 6db and 60db frequencies given, the filter passband could be plotted on squared paper: however, the graph would not be quite correct because in practice the two skirts are not straight lines—they are in fact slightly concave. To avoid this error the manufacturer also gives the two passband frequencies at the 30db points. The procedure then is to get a sheet of one tenth of an inch squared paper and divide this vertically at 10db per inch starting from the top at 0db and horizontally at 1·0 kc/s per inch arranging that the 455·0 kc/s point is in the centre of the graph paper—this is then indicated clearly by a dotted centre-line. From the data sheet

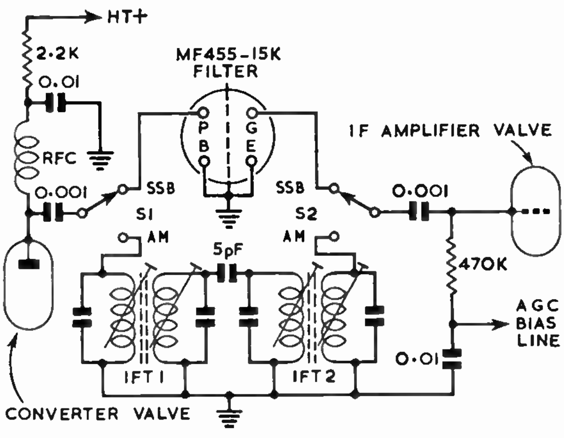


Fig. 10.38. Suggested circuit for the use of the Kokusai filter in an existing communication receiver to improve the s.s.b. selectivity. S1 and S2 can be switch banks fitted to the existing selectivity switch assembly, and should be of the type incorporating shorting plates. The shorting plate pole must be earthed. If the filter selectivity is not to be degraded, the two banks must be isolated from each other with an earthed cross screen. IFT1 and IFT2 are standard Maxi-Q IFT 11/465 i.f. transformers or similar.

KOKUSAI ELECTRIC Co., LTD.

MECHANICAL FILTER CHARACTERISTICS	Made by [illegible]	Checked by M. N.	Approvals [illegible]
TYPE MF455-15K	SERIES No. CCK 0665	Date AUG. 5, 1968 Temp. 26°C	
CENTRE FREQUENCY	⊕ −0.4 kc/s	GAIN PER STAGE	24.0 dB
BANDWIDTH AT 6dB ATTENUATION	+ 2.0 kc/s − 1.2 kc/s	INSULATION (More than DC 500V 200Mohm)	O.K.
30dB ATTENUATION (SSB)	457.42 kc/s 453.29 kc/s		
BANDWIDTH AT 60 dB ATTENUATION	+ 3.3 kc/s − 2.6 kc/s		

Fig. 10.39. Example of the data sheet packed with each Kokusai mechanical filter.

the 6db and 60db figures are plotted, the + (plus) figures to the right of the centre-line and the − (minus) figures to the left. By reference to the horizontal frequency scale, the two 30db figures are marked on the graph at the 30db down position. The 60db, 30db and 6db points on each side of the centre-line are then joined together with a shallow curve and this is the correct passband for the filter. A complete plot of the filter characteristics already given is shown in Fig. 10.40.

In the interests of acceptable voice quality the filter should pass frequencies down to 300 c/s, i.e. 300 c/s should be at the 6db point. From this it follows that the carrier frequency must be 300 c/s away from this position. A vertical line is therefore drawn on the graph 0·3 kc/s outside the 6db point and where this line cuts the passband curve is the correct position for the carrier crystal—on the graph shown, this is a frequency of 457·3 kc/s. If two carrier crystals are required (i.e. sideband switching in a receiver) the second frequency is plotted in exactly the same way but at the other side of the passband. If the carrier crystal on the lower side of the passband is in use, the filter will pass the higher sideband. If the carrier crystal on the higher side of the passband is in use, the filter will pass the lower sideband.

Ordering Crystals

When ordering crystals it should be remembered that every crystal can be made to oscillate on either its series or parallel resonant frequency and these are not the same (at frequencies around 455 kc/s the difference may be 200 c/s or more). It is not sufficient to quote only the required frequency; the manufacturer must know whether operation is required on the parallel or the series resonance, and this must be stated. As the parallel resonance is affected by the shunt circuit capacity, this capacity should also be stated. (Normally in the United Kingdom 30 pF is taken as a preferred standard value and the manufacturer will grind the crystal to oscillate at parallel resonance at the stated frequency with this capacity, unless otherwise instructed.)

For amateur use it is not necessary to go into involved calculations in an attempt to determine circuit capacity—the value of 30 pF can be quoted and is quite near enough in practice. The manufacturer also has to know (i) the type of holder, i.e. 10XJ (½ in. pin spacing) or B7G; (ii) frequency tolerance (this is normally plus or minus 100 p.p.m. or 0·01 per cent; (iii) operating temperature range (it would be sufficient here to quote " normal room temperature " or " amateur equipment ").

Carrier crystals are normally used in either Pierce, Colpitts or Miller oscillator circuits and in all these the crystal is excited at its fundamental parallel resonant frequency.

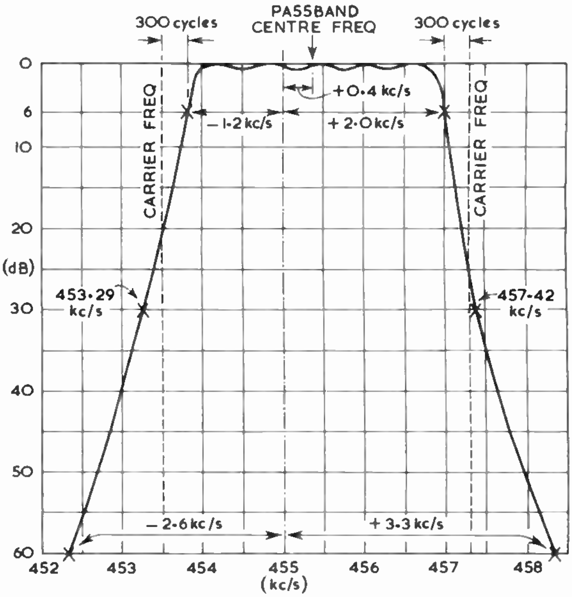


Fig. 10.40. Method of plotting the passband curve for a Kokusai mechanical filter from characteristics given on the data sheet supplied and from this determining the required carrier crystal frequencies.

In the case of either m.f. or h.f. sideband filters, the crystals operate as coupling elements in the series arms of the half-lattice sections; therefore the required crystal is specified for operation on the quoted series resonant frequency.

High frequency crystals will also be ordered for use in some other part of the equipment and in this case the manufacturer will have to know whether the mode of operation is fundamental or overtone. Overtone oscillators always operate at series resonance and as this is not affected by shunt circuit capacity it is unnecessary to state this. Often, final conversion crystals are operated on their overtone to avoid generating spurious signals—this applies particularly to a double conversion amateur band receiver. The mode of operation can be readily identified by reference to the type of oscillator circuit—this will be a Butler, Squier or Robert Dollar in which the crystal is used as a *series* coupling of low impedance to the required oscillatory r.f. current.

Remember that in an harmonic oscillator—as distinct from an overtone oscillator—the crystal oscillates at its *fundamental* parallel resonant frequency. The harmonic genera-

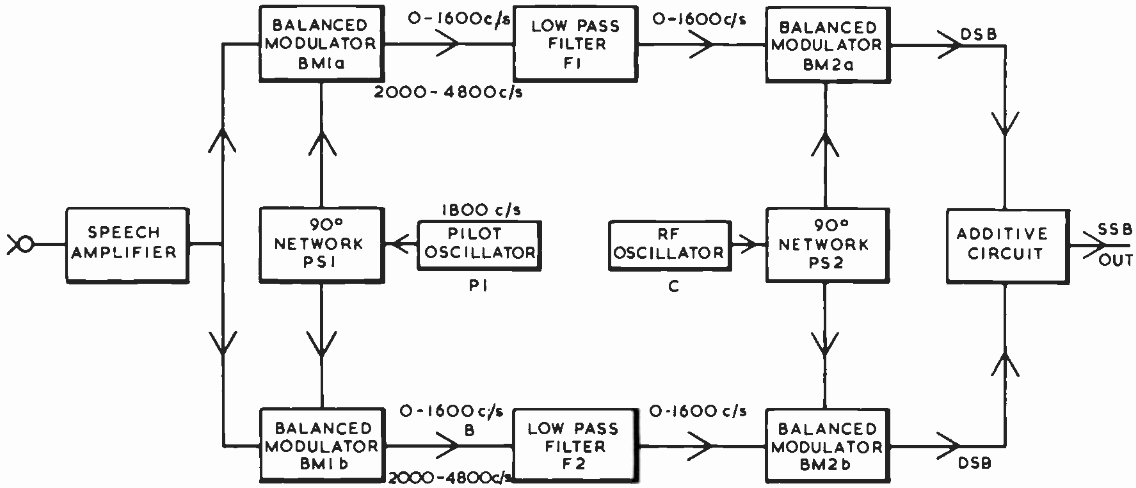


Fig. 10.41. Block diagram of the third method of generating a s.s.b. signal. For lower sideband transmission, the nominal output frequency will be fp + fc and for upper sideband transmission, the nominal output will be fc − fp. The nominal output frequency itself is not generated.

tion is obtained by the non-linearity of the associated valve, the valve output circuit being tuned to resonance at the required harmonic frequency. This type of oscillator normally uses the Colpitts, Miller or electron-coupled circuit.

The subject of crystals and crystal oscillator modes of operation is dealt with in detail in Chapter 6 (*H.F. Transmitters*).

The Third Method

In addition to the methods already described, details have been published of a third method of generating a s.s.b. signal. Because it is rather more complicated, insofar as it combines the principles of both phasing and filter-type equipment, the third method has not attracted much interest among amateurs. A brief description is however included because the system is said to preserve a higher degree of fidelity than a sharp-cutting filter, and to give better sideband suppression than a straightforward phasing exciter.

The third method combines the principles of both systems in such a way that a wideband audio phase-shift network is unneccessary, and although filters are used, their cut-off characteristics are not abrupt enough to detract from speech quality. The method is shown diagrammatically in **Fig. 10.41.** The output of a pilot audio frequency oscillator P is applied to a differential phase-shift network, preferably of low Q. The two components emerging from the network are adjusted to be approximately equal in amplitude, and to differ in phase by exactly 90°. Oscillator frequency is not particularly critical; the system can be made to work at any point from 1600 c/s to 4·5 kc/s, but as will be seen later there is an incidental dividend in remaining within the range 1600–1800 c/s. The quadrature signals from the first phase-shift network PS1 are applied to two identical balanced modulators BM1a and BM1b. Audio excitation from a conventional speech amplifier is fed in parallel to the other inputs of BM1a and BM1b. Because of the low frequencies involved, it is possible to achieve an exceptionally high order of pilot carrier attenuation in the balanced modulators, and the output of each will consist of two sidebands at audio frequency, $Fp + Fm$ and $Fp - Fm$. Amplitudes and waveforms will be identical, but the sidebands from one balanced modulator will be 90° out of phase with those from the second. The higher-frequency sidebands are eliminated from both channels by a pair of identical low-pass audio filters. If a pilot frequency of 1800 c/s is assumed and a minimum modulating frequency of 200 c/s, each filter will be required to pass all frequencies up to 1600 c/s, and to attenuate rapidly above 1600 c/s, reaching maximum rejection at 2000 c/s. It is comparatively easy to attain this performance with standard tolerance components, and high stability is not essential in either the pilot oscillator or the filter itself. The phase shift network PS1 is the only part of the circuit calling for precision adjustment, but a simple differential arrangement will take care of more drift than is likely to be met in any practical pilot carrier oscillator.

The quadrature audio signals issuing from F1 and F2 are applied to a further pair of balanced modulators, where they modulate quadrature r.f. carriers from PS1 exactly as in a conventional phasing transmitter. The r.f. excitation may be at the desired output frequency, or if multiband operation is contemplated, at a neutral frequency convenient for heterodyning into the required bands. When the double sideband signals from the two balanced modulators are combined in an additive tank circuit, the wanted components reinforce one another, and the unwanted sideband balances itself out to an extent which depends upon the accuracy of the two 90° networks PS1 and PS2. It is claimed that attenuation exceeding 40db is achievable.

The main advantage of using as low an audio frequency as possible for the pilot carrier lies in the resulting simplification of the low-pass filter design. In the numerical example quoted, the response of the filter is required to change from zero to full attenuation in the 400 cycles between 1600 and 2000 c/s; that is, in 25 per cent of the highest frequency to be passed. Had 3600 c/s been chosen as the pilot frequency, the slope of the filter response curve would have to be twice as steep in proportion if the same performance were to be maintained. This would naturally call for more sophisticated filter design. An incidental advantage of a low pilot frequency stems from the fact that the modulating spectrum becomes "folded" upon itself for all audio frequencies exceeding that of the pilot oscillator. In the earlier numerical example, a tone of 1000 c/s would beat with the pilot carrier to give a difference frequency of 800 c/s. If, however, the audio tone were 2600 c/s, the difference frequency would also be 800 c/s, but the phase would be altered by 180°. Taking the system as a whole, it follows that imperfect sideband suppression manifests itself as inverted speech superimposed upon the wanted sideband and not as adjacent channel interference. The protagonists of the third method claim that the listener's ear can discriminate more readily against this type of imperfection than against the type of interference caused by poor suppression in a conventional s.s.b. transmitter. The validity of this argument is open to question, because many amateurs now use highly selective receivers

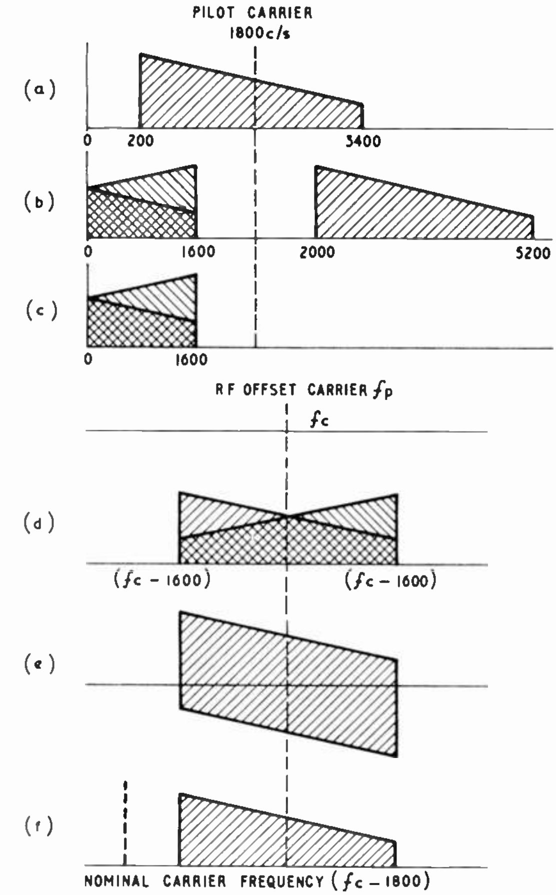


Fig. 10.42. Frequency relationship in a third method transmitter: (a) a.f. input spectrum to BM1a and BM1b; (b) a.f. output from BM1a and BM1b for a pilot carrier frequency of 1800 c/s. Note how the lower sideband is folded; (c) a.f. output from lowpass filters F1 and F2; (d) r.f. output from upper modulator chain to Fig. 10.40; (e) r.f. output from lower modulator chain in Fig. 10.40; (f) Single sideband from additive tank circuit in which (d) and (e) are combined. By reversing the phase shift of the output from either BM2a or BM2b, upper or lower sideband transmission may be selected at will.

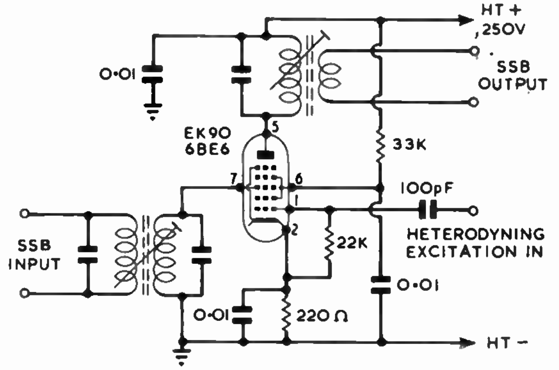


Fig. 10.43. Simplest form of frequency converter using a pentagrid valve. The tuned circuits should resonate at the appropriate input and output frequencies.

capable of rejecting image-type interference. Selectivity at the receiving end affords no improvement to a poor third method transmission. There is however some rough justice in the fact that a badly adjusted third method transmitter causes more irritation to its owner than to the occupants of adjacent channels, while the reverse is usually true of a poor transmitter of more conventional design.

A fuller explanation of audio spectrum folding would require mathematical treatment beyond the scope of this *Handbook*. The description in the preceding paragraph in conjunction with the diagrams of **Fig. 10.42** should, however, show in broad outline how the system operates. Readers who desire a deeper understanding of the third method are referred to the excellent article by D. K. Weaver in the *Proceedings of the I.R.E.* for December 1956.

A third method transmission is received in exactly the same way as s.s.b. generated by either of the more common systems. As will be seen from Fig. 10.42, the carrier insertion oscillator will have to be offset from the nominal frequency of transmission by an amount equal to the frequency of the pilot audio oscillator. The receiving operator will not, however, be conscious of this fact. Imperfect carrier suppression will show up, not as a steady heterodyne against which the carrier insertion oscillator may be zeroed, but as an audio tone at the frequency of the pilot carrier in the transmitter.

FREQUENCY CONVERSION

Multiband operation of an s.s.b. transmitter normally involves the generation of the signal at a fixed frequency, and the provision of one or more heterodyne converters to enable output to be obtained in the required amateur bands.

The pentagrid valve, of which the 6BE6 is representative, is probably the best known type of frequency converter. The circuit of **Fig. 10.43** is typical, and may be used either to raise the output from a filter in the i.f. range to the 80m band, or to change the transmitter frequency from one band to another. For linearity, the oscillator excitation to grid 1 should be close to 2·5 volts r.m.s. and the s.s.b. signal at grid 3 should not be allowed to exceed 0·1 volt r.m.s. at peak input. Should the s.s.b. output of the preceding stage exceed this value, a voltage divider should be interposed to reduce the excitation to the recommended level. The excitation could of course be reduced by turning down the audio gain but this would

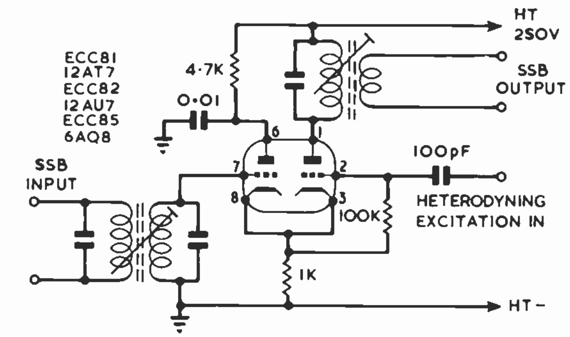


Fig. 10.44. "Single-ended" double triode frequency converter. The tuned circuits should resonate at the appropriate frequencies.

impair the ratio of sideband signal to resting carrier and nullify the good work done by the balanced modulator. It is a sound principle of s.s.b. transmission that the audio gain control should be used solely to secure optimum operating conditions for the balanced modulator, and that the excitation of the converter or amplifier stages should be adjusted *after* the single sideband has been generated.

Apart from the low level of output, the only serious shortcoming of the simple pentagrid converter lies in its inability to discriminate against the unwanted energy from the heterodyning oscillator if the s.s.b. signal is to be raised in frequency by a ratio of more than five or six to one. The following numerical example will make the reason clear. Assuming that the output of a 460 kc/s crystal filter is to be converted to 3760 kc/s, the heterodyning oscillator will operate at 3300 kc/s. A single tuned anode circuit of good quality (high Q) adjusted precisely to 3760 kc/s, will present roughly 25 times the impedance to the wanted s.s.b. signal as it will to the steady off-tune oscillator signal. The oscillator input voltage will, however, have been adjusted to at least 25 times the amplitude of the peak s.s.b. input, so despite the selective effect of the output tank circuit, the steady oscillator voltage across it will be just about the same as the peak s.s.b. voltage. Subsequent amplifier stages will, of course, afford further discrimination against the oscillator signal, but if the tuned circuits are wide-band or swamped in the interests of linearity, enough energy may leak through to produce an appreciable spurious signal outside the amateur bands. For conversion ratios exceeding three to one, it is desirable to employ a critically coupled pair of really high Q tuned windings in the anode circuit.

With appropriate changes to circuit values, any receiving type frequency converter may be substituted for the pentagrid in Fig. 10.43. If the oscillator is crystal controlled, a valve envelope may be saved by using the triode section of a triode-hexode as oscillator, but it may prove difficult to adjust the excitation to the precise value for maximum linearity. A self-excited oscillator should preferably be isolated from the frequency converter by a buffer stage, otherwise pulling and frequency modulation may be experienced.

Although multi-element valves are convenient and efficient, comparable results may be obtained with double triodes such as the 12AU7, 12AT7 and ECC85. A representative circuit is shown in **Fig. 10.44**. It may be operated at the same voltage levels as those recommended for the 6BE6 and has the same shortcomings.

When the conversion ratio exceeds three to one, many designers prefer to attenuate the oscillator voltage in the mixer output circuit by a system of phasing or balancing. The principle is identical to that of carrier elimination by balanced modulator. The classical balanced frequency converter circuit, widely used in s.s.b. equipment, is shown in **Fig. 10.45**. This is capable of attenuating the oscillator output by 20–25db. Excitation may be any value between 2·5 and 10 volts r.m.s., depending on the amplitude of s.s.b. signal the converter valve is required to handle.

Balancing is affected by adjustment to the cathode potentiometer with oscillator drive but no sideband input, until the heterodyning voltage in the output circuit is at its lowest value. This may be conveniently measured either with a pick-up loop to an absorption wavemeter or a valve voltmeter or oscilloscope probe. The capacitors C1 and C2 in the grid circuit should be silver mica of 1 per cent tolerance, each of equal value to provide an accurate centre-tap, and with an effective value that will correctly resonate the tuned circuit. This also applies to C3 and C4 in the anode circuit.

Where multiband operation is contemplated from the outset, it is sound practice to design the transmitter so that frequency conversion may be effected at low level by one or other of the arrangements described above. It sometimes happens, however, that the output of an existing exciter, capable of delivering several watts of power in one band only, has to be heterodyned to a different band. In this case, receiving-type valves are out of the question, and conversion has to be effected by small transmitting tetrodes or pentodes. The most satisfactory of the many possible circuit configurations is that shown in **Fig. 10.46.**

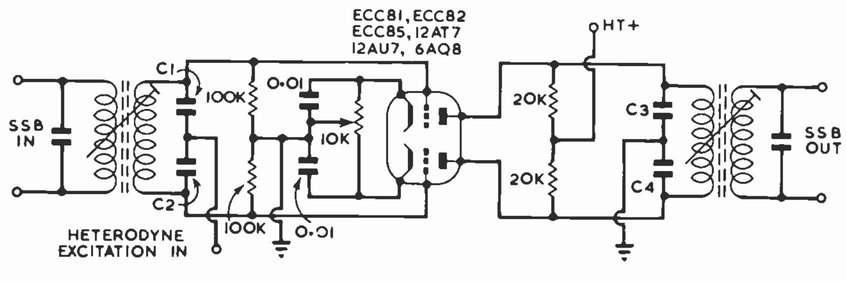


Fig. 10.45. Double-triode balanced frequency converter.

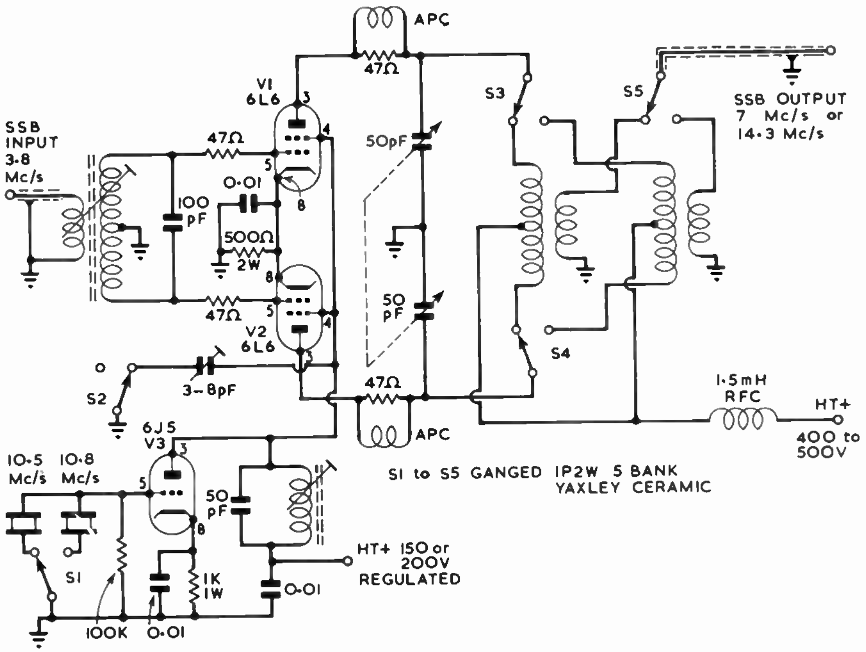


Fig. 10.46. Practical circuit suitable for high level mixing in the 7 and 14 M/cs band.

Choice of Mixing Frequencies

It is the usual practice in s.s.b. transmitting equipment to generate the initial single sideband signal at a fixed frequency and translate this signal to the required operating frequency by one or more frequency changing processes. The sideband signal *F2* is used to modulate a high frequency carrier *F1*, whose frequency is chosen so that either the upper or the lower sideband of the mixing process is on the desired operating frequency. As a result of this modulation process, the sideband signal will be translated to a new frequency that is either the sum of the carrier and sideband frequencies *F1* + *F2* or the difference between the carrier and the sideband frequency *F1* − *F2*. Of further importance is the fact that if the lower sideband of the mixing process is selected, an inversion of the sideband occurs—an upper sideband signal will be converted to a lower sideband signal.

The two essential components of the translation system are the modulator (commonly called mixer) and the carrier (commonly called oscillator, injection, or heterodyne signal). In order to provide a mixer output waveform that is in linear relationship to the s.s.b. input waveform, it is necessary to make the amplitude of the oscillator injection several times greater than the amplitude of the sideband input. A practical figure generally used is an oscillator injection voltage ten times (+ 20db) greater than the maximum signal input.

If two frequencies *F1* and *F2* are fed into a mixer it is important to remember that the mixer output will contain not only the wanted *F1* + *F2* or *F1* − *F2*, but additional frequencies composed of the following:

Fundamental	*Second Order*	*Third Order*	*Fourth Order*	*Fifth Order*
F1	*F1* ± *F2*	*2F1* ± *F2*	*3F1* ± *F2*	*4F1* ± *F2*
	2F1	*2F2* ± *F1*	*3F2* ± *F1*	*4F2* ± *F1*
F2	*2F2*	*3F1*	*2F1* ± *2F2*	*3F1* ± *2F2*
		3F2	*4F1*	*3F2* ± *2F1*
			4F2	*5F1*

Assuming the inputs were *F1* = 3000 kc/s and *F2* = 500 kc/s the mixer output would have these frequencies:

Fundamental	*Second Order*	*Third Order*	*Fourth Order*	*Fifth Order*
	3,500	6,500	9,500	12,500
3,000	2,500	5,500	8,500	11,500
	6,000	4,000	4,500	5,000
500	1,000	2,000	1,500	1,000
		9,000	7,000	10,000
		1,500	5,000	8,000
			12,000	7,500
			2,000	4,500
				15,000
				2,500

(Wanted output *Fo* = 3,500 kc/s)

From these figures it is clearly seen that a train of frequencies has been generated, with a separation equal to *F2* (the lowest input freq.).

The response or selectivity curve of an average tuned circuit of the kind likely to be used is shown in **Fig. 10.47.** The two frequencies most difficult to eliminate will be the two either side of the wanted output *Fo* of 3500 kc/s. These are the strong *F1* heterodyning input of 3000 kc/s and the third order product (*2F2* + *F1*) of 4000 kc/s. The curve shown is actually that of the 80m coil in position in a s.s.b. transmitter. The *Q* of the circuit can be determined from the formula $\frac{Fo}{Fh - Fl} = \frac{\text{centre frequency}}{\text{bandwidth at 3db points}}$

In this particular case this gives the value of $Q = 35$. The response has been plotted with the coil and the associated capacitor in place on the chassis, and is affected by the shunt loading of the valve, the shunt h.t. feed and the following grid input resistance. It is considered to be more sensible to evaluate the function of the tuned circuit under actual operating conditions, rather than to consider the unloaded

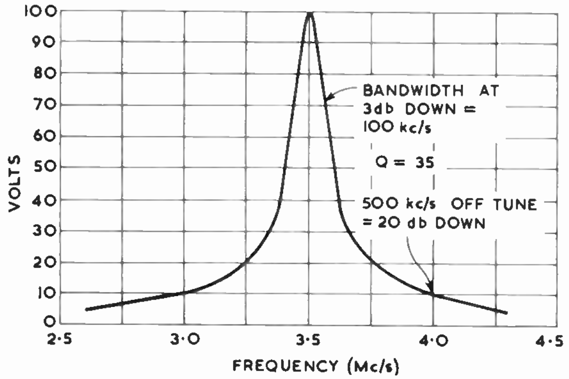


Fig. 10.47. Resonance curve for an 80 metre coil. Frequency at Fo = 3·5 mc/s.

Q measured on a Q meter. The response shown is believed to be a fair average of tests taken on representative coils and at different frequencies.

The two frequencies it is desired to eliminate are 500 kc/s removed. Plotting the 500 kc/s points on the response curve gives an attenuation of 20db and it is noted that the ratio *Fo : F2* is 7 : 1. A tuned circuit with equal Q value, resonant at 7 Mc/s would have the same response curve and is shown in Fig. 10.47 by doubling the frequency scale. However, 500 kc/s off tune is now a ratio of 14 : 1 and the attenuation has dropped to 14db. If the circuit were resonant at 1·75 Mc/s, 500 kc/s off tune would be a ratio of 3·5 : 1 and the attenuation would have increased to 25db.

It is therefore clear that the lower the ratio *Fo : F2* the better the attenuation from the tuned circuit. If two tuned circuits are used, coupled to give the highest possible selectivity, the total attenuation is twice that of a single tuned circuit and in this case would be $20 \times 2 = 40$db.

Looked at the other way round it can be seen that if an attenuation of 40db is required from two coupled tuned circuits in the mixer anode, the ratio of the lowest input frequency (*F2*) to the required output frequency (*Fo*) should not exceed a ratio of 7 : 1.

It is good design practice to aim for adequate suppression of all spurious conversion products and 60db should be looked upon as a desirable target. With the usual initial s.s.b. generation on 455 kc/s, a ratio of 7 : 1 sets the upper limit for *Fo* of $455 \times 7 = 3185$ kc/s. It is possible to go to a ratio of 8 : 1 (i.e. direct from 455 kc/s into the 80m band; or direct from 80m to 10m) but it is safer and results in a cleaner output to keep within the ratio of 7 : 1. In practice coupling is often tightened beyond the "critical coupling" point to prevent too much loss of gain, and this would reduce the available attenuation to a slightly lower value than 40db. This loss of attenuation could be offset by making the limit for the multiplying ratio 6 : 1. To sum up:

(i) In the interests of constructional simplicity it is desirable to keep the number of tuned circuits at the mixer output to two.

(ii) Assuming two coupled tuned circuits and normal Q values, the multiplying ratio of the frequency translation process should wherever possible not exceed 6 : 1.

(iii) A lower value than 6 : 1 will give greater attenuation of the unwanted products and therefore a cleaner output.

(iv) The lower multiplying ratio can be obtained either by increasing *F2* (the s.s.b. input), or reducing *Fo* (the wanted s.s.b. output); or a combination of both.

Spurious Product Amplitude

In order to show how undesired frequencies are generated in a mixer stage, it is necessary to consider the case where oscillator and signal voltages are applied to the same grid of the mixer valve. The wanted sum or difference frequencies can only be generated if the anode current, grid voltage characteristics have some non-linearity or curvature. The components of the anode current will be the d.c.; signal; oscillator; signal second, third, fourth, fifth . . . harmonics; oscillator second, third, fourth, fifth . . . harmonics; and the sum and differences of the signal and oscillator.

To obtain the desired sum or difference product it would be necessary to use a valve in which the characteristic curve had only second order curvature. Unfortunately, all practical valves have characteristic curves having higher order curvature and this contributes additional unwanted frequency components into the output current. Sometimes the frequency of these unwanted components is sufficiently removed from the desired output frequency and they are easily filtered out, but often these frequencies are very near to the desired signal frequency and they will fall within the passband of the selective filter used in the mixer output circuit. The amplitude or strength of these undesired mixer products varies with valve type, and even with valve to valve of the same type. It is also critically dependent on the amplitude of the signal input, the bias point, and the amplitude of oscillator injection. It is therefore not surprising that valve manufacturers have not been particularly successful in designing valves having the desired second order curvature to the exclusion of any higher order curvature. In practice, the circuit designer must

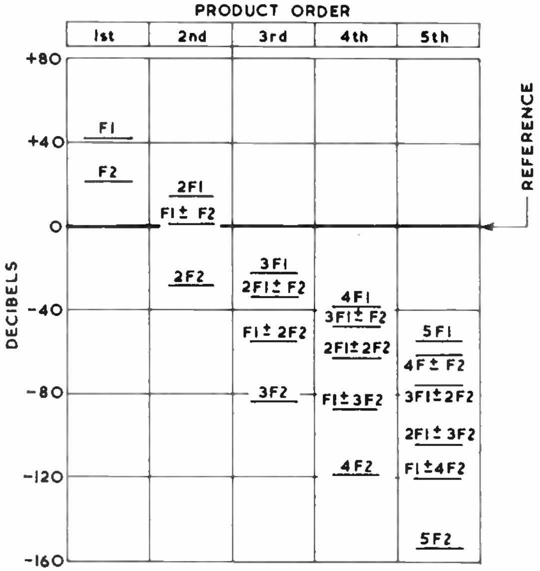
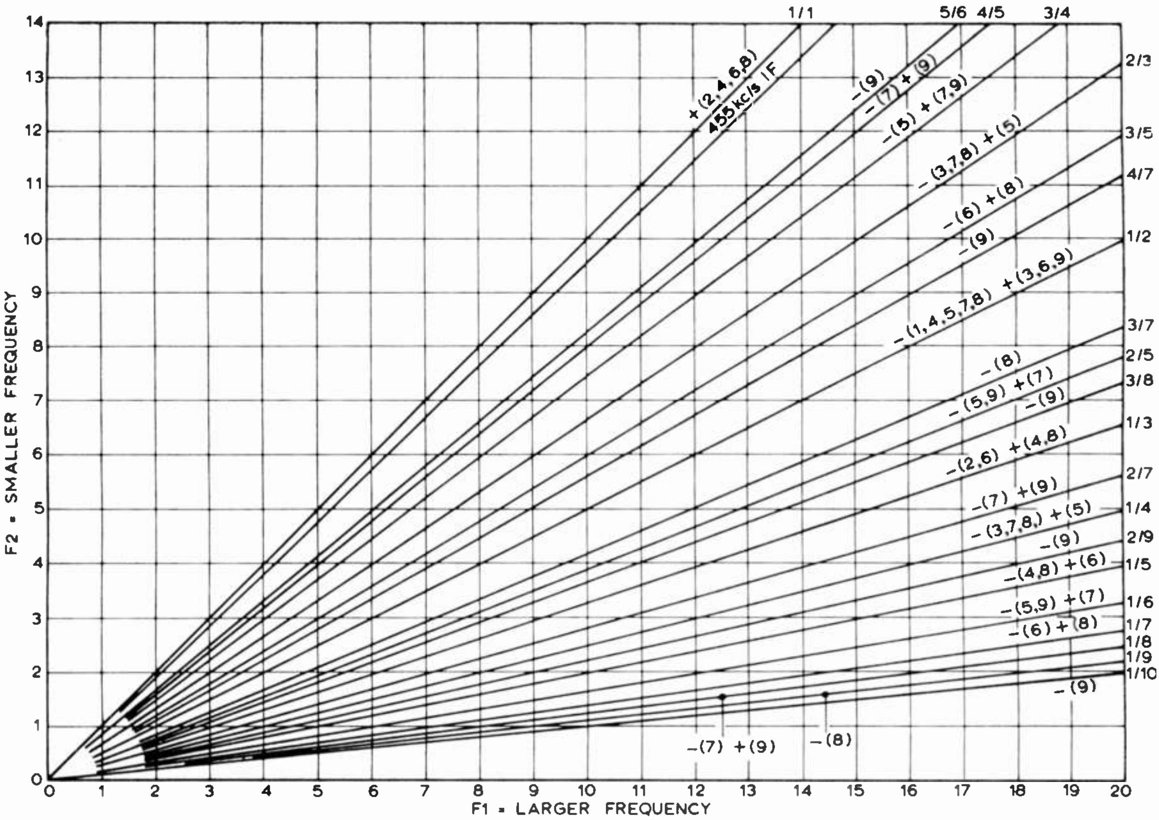


Fig. 10.48. Calculated frequency products contained in the anode current of a 12 AU7 triode mixer.



F2 ∼ F1

Order	1	2	3	4	5	6	7	8	9
1 / 1		•20 •02		•13 •31		•24 •42		•35 •53	
1 / 2	10		•12 •30	31	32	•33 •51	52	53	•54 •72
1 / 3		20		•22 •40		42 51		•53 •71	
1 / 4			30		•32 •50		52	71	
1 / 5				40		•42 •60		62	
1 / 6					50		•52 •70		72
1 / 7						60		•62 •80	
1 / 8							70		•72 •90
1 / 9								80	
1 / 10									90
2 / 3			21		•23 •41		43	53	

F2 ∼ F1

Order	1	2	3	4	5	6	7	8	9
2 / 5				41			•43 •61		63
2 / 7							61		•63 •81
2 / 9									81
3 / 4					32		•34 •52		54
3 / 5						42		•44 •62	
3 / 7								62	
3 / 8									72
4 / 5							43		•45 •63
4 / 7									63
5 / 6									54

• INDICATES SUM MIXING
OTHER DIFFERENCE MIXING

Fig. 10.49. Spurious response chart.

(Reproduced, by permission, from the 2nd Edition of "Fundamentals of Single Sideband" (published by Collins Radio Company, U.S.A.) where the subject is treated in greater detail.)

select his mixer valves by means of a series of experiments in which the amplitudes of these undesired mixer products are measured. A representative calculated determination of mixer product amplitude is shown in **Fig. 10.48.**

Consideration of this chart shows that there are several undesired products that are greater in amplitude than the desired signal and a considerable number that are weaker than the desired signal. Of further interest is the fact that as the product order increases, its amplitude decreases.

Avoiding Spurious Mixer Products

It is possible by an intelligent choice of the signal and oscillator frequencies to minimize the presence of undesired mixer products within the passband of the output circuits of the frequency translation system. The problem of frequency selection is relatively simple where the operating frequencies are fixed, but becomes increasingly complex where the operating frequency must be varied. In an attempt to simplify the problem, circuit designers make use of charts in which the frequency of the spurious mixer products is plotted with respect to the signal and oscillator frequencies. This type of chart is shown in **Fig. 10.49.**

As an example of the spurious product problem, consider the case of a s.s.b. transmitter for operation on the 80, 40 and 20m bands. The s.s.b. signal has been generated at 455 kc/s using a mechanical or crystal filter. Coverage of the lowest frequency band 3·5 to 4 Mc/s can be obtained by mixing the s.s.b. signal with the output from a v.f.o. tunable from 3·955 to 4·455 Mc/s. As the required output is the low sideband of the translation process (i.e. the heterodyning frequency is *higher* than the wanted 80m output) there will be a sideband inversion. In order to give a low sideband signal in the 80m band, this means that the mechanical filter must be arranged to provide a high sideband output. It has become an accepted convention adhered to by all s.s.b. stations, to transmit *low* sideband below 10 Mc/s and *high* sideband above 10 Mc/s as a result of a CCIR (International Radio Consultative Committee) recommendation.

The strong oscillator signal is, only 455 kc/s removed from the wanted output frequency and must be filtered out by the following tuned circuits or else balanced out through the use of a balanced mixer. As the mixer cannot be expected to retain an accurate balance over a 500 kc/s tuning range, it is necessary to resort to a combination of both methods to obtain suppression of this spurious frequency of at least 60db. At the higher operating frequencies required for the other amateur bands, it becomes increasingly difficult to suppress this product because the selectivity required in the tuned circuits is so high as to become impracticable. This difficulty can be overcome by using a second stage of frequency conversion.

The output from the first mixer at 3·5 to 4 Mc/s is mixed with the output from a crystal oscillator. This could be on 3·5 Mc/s, giving a tunable output of 7 to 7·5 Mc/s for the 40m band; however, the second harmonic of the crystal will give a crossover with the wanted 7 Mc/s output signal and in practice this would be avoided by using a lower frequency crystal—say 3·3 Mc/s. Even with this choice of frequency, the second harmonic of the crystal at 6·6 Mc/s is only 400 kc/s removed from the low-frequency end of the wanted output. The 7·4 Mc/s second harmonic of the input signal (3·7 to 3·8 Mc/s to give a band coverage of 7 to 7·1 Mc/s) is only 400 kc/s on the other side of the wanted output frequency. A very high order of selectivity would be required to reduce these spurious signals to a satisfactory level. In practice the designer would overcome this problem by placing the heterodyning frequency on the high side of the wanted output frequency by using a crystal of 11 Mc/s. The sideband input of 3·5 to 4 Mc/s modulating the 11 Mc/s carrier would produce a difference frequency output over the range 7·5 to 7 Mc/s. Unfortunately this would also produce a reverse dial calibration and would also invert the sideband; i.e. the transmitter would be radiating a high sideband signal. The sideband could be corrected by switching carrier crystals at the filter, but the reverse dial calibration would have to be accepted. It will be noted from the chart of Fig. 10.49 that there is a second order crossover at 3·67 Mc/s; this is shown in greater detail by the graph of **Fig. 10.50.** Because the 40m band only extends from 7 to 7·1 Mc/s the s.s.b. input (*F2*) will never be required below 3·9 Mc/s and the crossover may be safely ignored.

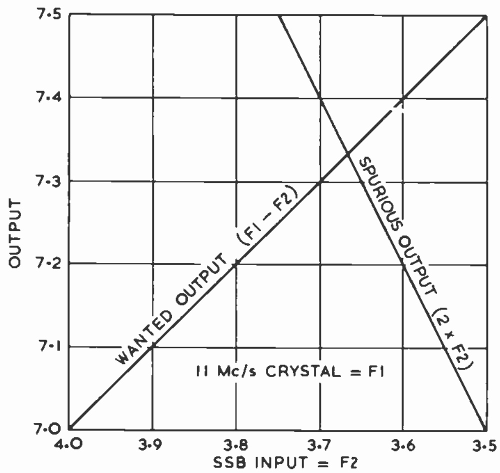


Fig. 10.50. Graph showing the second order crossover at 7.34 Mc/s when a basic 3·5 to 4·0 Mc/s s.s.b. signal is heterodyned by an 11 Mc/s crystal to give a difference frequency output in the 40 m band.

As the 80m output from the first mixing process is a low sideband signal, and it is the convention to transmit the high sideband on the three higher frequency amateur bands, the heterodyning frequency for 14 Mc/s will have to be higher than the required output frequency, i.e. the second translation process will invert the sideband, thus giving high sideband output from a low sideband modulating signal. A suitable conversion frequency for 20m would be 18 Mc/s. Inspection of the chart in Fig. 10.49 will show that there is a fourth order spurious when the signal input to the mixer (*F2*) is near to 3·5 Mc/s. The band, however, only extends from 14 to 14·350 Mc/s and this will be given by an input signal of 4 to 3.65 Mc/s, and the tuning range 3·65 down to 3·5 Mc/s will not need to be used.

There is an obvious temptation to include the 160m band. Where the basic exciter has a 500 kc/s coverage from 3·5 to 4 Mc/s many amateurs have attempted to get 160m s.s.b. output by heterodyning with some convenient crystal in the easily obtainable FT243 range of surplus crystals. Remembering the 160m band has a bandwidth of 200 kc/s from 1·8 to 2 Mc/s, consideration of the factors involved will show that the lowest frequency that can be used for heterodyning is 5·5 Mc/s and the highest is 5·8 Mc/s. Unfortunately the mixing process will produce the second harmonic of the 80m s.s.b. output and this will also be heterodyned by the crystal oscillator to produce a spurious distorted signal. (This is the third order difference product *2F2 – F1*.) When the v.f.o. is altered this spurious signal changes frequency at twice the rate of the wanted signal and in the opposite direction. **Table 10.2** has been compiled to show this clearly.

Many existing exciters are peaked to give maximum output at the h.f. end of the 80m band and the logical choice of heterodyning crystal would be 5·65 or 5·7 Mc/s. Under these conditions the spurious output is either on or very close to the wanted output. If any adjustment is made to the main tuning, either initially to find a clear channel or to avoid interference, the spurious output will move twice as fast and in the opposite direction. While it is appreciated that this output will not be as strong as the wanted one, it will most certainly be radiated by the aerial with sufficient strength to cause interference to stations in a wide area. It can at certain settings of the main tuning actually be on the wanted output frequency and put severe distortion on the transmitter signal.

TABLE 10.2

Basic 80m s.s.b. output	Second harmonic of basic 80m s.s.b. output	Heterodyning crystal frequency	160m output	
			Wanted	Spurious
3·5 Mc/s 3·6 Mc/s 3·7 Mc/s	7·0 Mc/s 7·2 Mc/s 7·4 Mc/s	5·5 Mc/s 5·5 Mc/s 5·5 Mc/s	2·0 Mc/s 1·9 Mc/s 1·8 Mc/s	1·5 Mc/s 1·7 Mc/s 1·9 Mc/s
3·6 Mc/s 3·7 Mc/s 3·8 Mc/s	7·2 Mc/s 7·4 Mc/s 7·6 Mc/s	5·6 Mc/s 5·6 Mc/s 5·6 Mc/s	2·0 Mc/s 1·9 Mc/s 1·8 Mc/s	1·6 Mc/s 1·8 Mc/s 2·0 Mc/s
3·7 Mc/s 3·8 Mc/s 3·9 Mc/s	7·4 Mc/s 7·6 Mc/s 7·8 Mc/s	5·7 Mc/s 5·7 Mc/s 5·7 Mc/s	2·0 Mc/s 1·9 Mc/s 1·8 Mc/s	1·7 Mc/s 1·9 Mc/s 2·1 Mc/s
3·8 Mc/s 3·9 Mc/s 4·0 Mc/s	7·6 Mc/s 7·8 Mc/s 8·0 Mc/s	5·8 Mc/s 5·8 Mc/s 5·8 Mc/s	2·0 Mc/s 1·9 Mc/s 1·8 Mc/s	1·8 Mc/s 2·0 Mc/s 2·2 Mc/s

It is clear from the examples shown that the design of a multi-band exciter is by no means a simple problem if spurious outputs are to be avoided—the correct choice of operating frequencies used in each of the frequency translation processes is of major importance. With a one or two band exciter the problem is relatively simple. As the requirement for multi-band operation increases—particularly if this is a six band coverage from 160 to 10m—the difficulties become more complex and can only be satisfactorily solved by an additional stage of frequency conversion together with the initial tunable s.s.b. output on a neutral frequency outside the wanted amateur bands. In general spurious product orders above five can be ignored because the level of a sixth, seventh, eighth or ninth order spurious signal should be sufficiently low to have no nuisance value, provided that the mixer valve is being operated well within its signal handling capabilities and with the optimum level of heterodyne injection voltage.

Heterodyning Oscillator Requirements

From any s.s.b. transmitter the frequency stability of the output signal is dependent on the frequency stability of the heterodyning inputs to the frequency changer valves. The total frequency error is the arithmetic of the errors in all these oscillators. For the purpose of amateur requirements the crystal oscillator can be considered stable; the stability of the transmitted signal will then be directly dependent on the stability of the variable LC oscillator, the v.f.o.

It is common practice on the s.s.b. bands for a number of stations to operate in a " net " (all stations accurately tuned to the same carrier frequency) and carry on a quick " backwards and forwards " normal type of conversation. Under these conditions it is most important that all stations remain accurately on frequency with negligible drift. With a transmitter that is moving off channel because of frequency drift in the v.f.o. speech can no longer be demodulated correctly at the receiver. Quite obviously v.f.o. stability is highly valued by s.s.b. workers. Ever since amateur sideband commenced, thousands of experimenters and enthusiasts have been searching for that very elusive object, the driftless v.f.o.

In regard to v.f.o. stability it should be clearly understood that, notwithstanding the many claims that have been made by the authors of various " ultimate " circuits published in the past, there is no such thing as a driftless LC oscillator. A quartz crystal has a high degree of frequency stability because quartz is a material with a low temperature coefficient. Replacing the crystal by building an equivalent series tuned circuit using L and C—as for instance in the so-called Clapp v.f.o.—does not give the same standard of stability as a crystal, since it is not possible to manufacture standard coils and variable capacitors with the temperature co-efficient of natural quartz.

Certain circuits such as the Tesla/Vackar and the Clapp use large values of fixed swamping capacity across the oscillator valve and claim that this reduces the effect of changing valve input capacity during initial warm-up. An LC oscillator that drifted 2 or 3 kc/s during the first 10 or 15 minutes and was thereafter rock stable, would, however, be very acceptable to almost everyone—no amateur would mind switching on and waiting 15 minutes before he intended to transmit. In practice the major annoyance is the long term drift taking place continually over the transmitting period of a couple of hours or so. This slow drift is caused by the changing temperature of the two components that make up the frequency determining resonant circuit, the L and the C. Aside from the temperature rise of the air in the cabinet and the air in the transmitting room, a considerable amount of heat from the valves—though the valveholders, the screening cans and skirts—warms up the chassis and this in turn warms up the v.f.o. tank coil and tuning capacitor.

Stability can be materially improved—not by a change to some " ultimate " circuit—but by re-building the v.f.o. so that the coil and tuning capacitor and associated components are built as a self-contained unit mounted on a sub-platform made of some poor heat conducting material such as polystyrene or Perspex. The " tuning unit " is then enclosed in a heat insulating " box " made from cellular polythene sheet ½ in. or so in thickness. The v.f.o. valve or valves are left in their original positions on the main chassis. As the object is to isolate the tank circuit elements from conducted heat, the variable capacitor shaft is coupled to the dial drive mechanism by a poor heat conducting flexible coupler such as the Eddystone Type 529 or Type 50.

That there is no simple solution to the problem of v.f.o. stability is born out by the fact that a number of the leading world communication equipment manufacturers have spent a considerable amount of time and effort in developing phase-locked or crystal synthesizer variable master oscillators and that these are now being supplied not only to the commercial market but to the amateur market as well.

The requirement for a home constructed amateur v.f.o. are:

(i) constant amplitude of output voltage over a 500 kc/s tuning range;
(ii) low harmonic output;
(iii) "simple" circuitry using standard components and valves;
(iv) after an initial warm-up period of 10 to 15 minutes frequency drift not to exceed 10 parts per million (10 cycles per Mc/s) per hour.

With reasonable care in construction and some experimental work to determine the required value of negative temperature co-efficient capacitor compensation, these parameters can be met with the standard Colpitts tunable oscillator shown in the circuit diagram of Fig. 10.100 and that of **Fig. 10.102.**

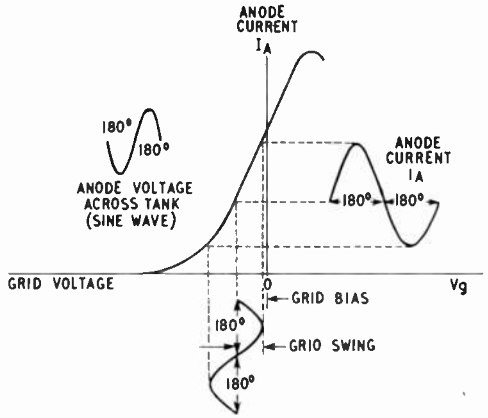


Fig. 10.51. Class A operation of a valve (dynamic characteristics).

THEORY OF LINEAR AMPLIFICATION

All the exciters described in this chapter are capable of driving a high power linear amplifier. Much effort will have gone into the construction and adjustment of an exciter capable of an acceptable degree of carrier and sideband attenuation, so care must be taken to ensure that subsequent amplifier stages do not degrade the overall performance of the transmitter by introducing avoidable distortion. This makes it essential to operate all s.s.b. amplifiers under strictly linear conditions.

R.F. Linear Power Amplifiers

The function of the power amplifier in a s.s.b. transmitter is to raise the power level of the input signal without change, so that the envelope of the output signal is a faithful replica of the envelope of the input signal.

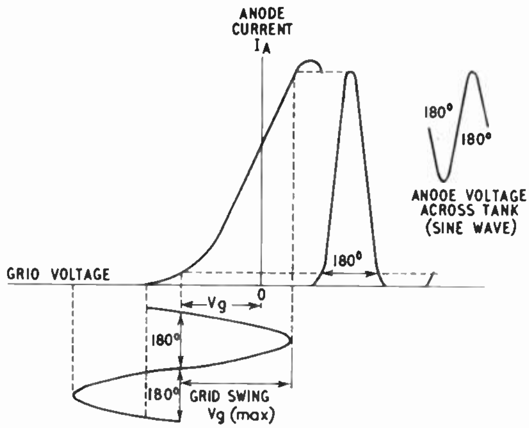


Fig. 10.52. Class B operation (dynamic characteristics).

Radio frequency amplifiers are classified A, B, and C according to the angle of anode current flow—the number of degrees the anode current flows during the 360° r.f. cycle. The class A amplifier has a continuous anode current flow and operates over a small portion of the anode current range of the valve. This amplifier is used for amplification of small signals where low distortion is required. Its efficiency in converting d.c. anode power input into r.f. power output is low, of the order of 30 per cent, but this is not of great importance where small signals are concerned (see **Fig. 10.51.**)

Class B amplifiers are biased to near anode cut-off so that anode current flows for approximately 180° of the r.f. cycle. Amplifiers operating with more than 180° of anode current flow, but less than 360°, are called class AB amplifiers. Both class AB and class B amplifiers are used in high power linear amplifiers stages to obtain higher efficiency and maximum output power with low distortion. The amplifier efficiency depends on the operating condition selected and

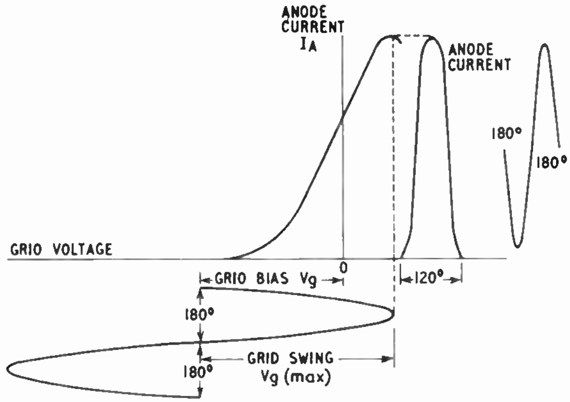


Fig. 10.53. Class C operation (dynamic characteristics).

the type of valve used and is usually of the order of 50 to 66 per cent (see **Fig. 10.52**).

Class C amplifiers are biased well beyond cut-off so that anode current flows for less than 180° of the r.f. cycle. The principal advantage of this mode of operation is high anode efficiency (of the order of 65–80 per cent); however, the class C amplifier is not suitable for s.s.b. use because the amplifier is not linear and it will not respond to low level input signals (see **Fig. 10.53**). The amplifier class can be followed by a number to indicate whether or not the valve is operated in the positive grid region over part of the r.f. cycle. Class AB1 indicates that the grid never goes positive and that no grid current is drawn; class AB2 indicates that the grid goes positive and that grid current is drawn. Class A amplifiers are nearly always operated without grid current and class C amplifiers are nearly always operated with grid current. It is therefore normal practice to refer to these as class A or class C without the necessity for further designation.

Choice of Valve

There is a wide choice of triode valves suitable for power amplifiers and these have the advantage of simplicity and low cost. Generally, they require a large amount of driving power and because of the considerable grid-to-anode capacity the valve must be carefully neutralized. The amplification

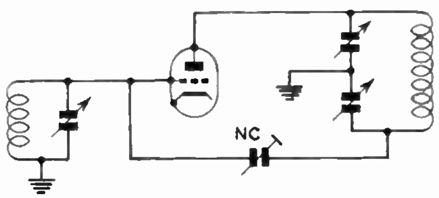


Fig. 10.54. Grid driven anode neutralized amplifier.

factor of triode valves suitable for amateur s.s.b. application is generally between five for low mu triodes and 75 or more for high mu triodes. Usually only the low and medium mu valves are suitable for the requirements of power amplifiers and it is therefore necessary to provide a large grid swing to obtain the power amplification available from the valve.

With the tetrode valve the screen grid is an electrostatic shield between the grid and the anode and this reduces the grid anode capacity to about one hundredth of that of the triode. It is still necessary to provide neutralization because of the higher amplification of the tetrode valve but the small value of feedback eases the neutralization problem. Because of the higher gain available, the valve requires a relatively low drive to obtain a high power output.

Pentode construction is also available in power amplifier valves and these give improved efficiency because the r.f. anode swing can be increased. The pentode, however, is more complex and expensive than the tetrode and in some cases requires additional supply voltage for the suppressor grid. These disadvantages have limited the development of pentode power valves and at present the pentode has little advantage over the well-designed tetrode.

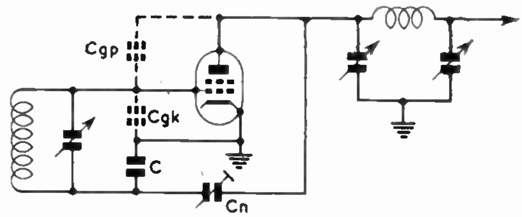


Fig. 10.55. Grid driven grid neutralized amplifier. For neutralization, $C_{gp}/C_{gk} = C_n/C$. C may be any convenient value between 250 and 500pF.

For linear power amplifier operation the following features are desirable:

(i) high gain;
(ii) good efficiency;
(iii) low grid to anode capacity;
(iv) linear characteristics at all frequencies within the desired operating range.

When a valve is operating in class A the degree of linearity is quite high but the efficiency is low—of the order of 25–30 per cent. By operating class AB1 the principal advantages of class A operation are retained while the efficiency is raised to between 50 and 65 per cent.

If operation is further advanced into the class AB2 region, the efficiency is improved only slightly, but any gain in this direction is more than offset by the more stringent requirements imposed upon the driver stage brought about by variation in driver loading as the amplifier is driven into the positive region. This can, and often does, result in a compression of the modulation peaks and non-linearity unless the driver is capable of providing the increased loading. Usually, in order to maintain driver regulation in class AB2 linear amplifier operation, a relatively large percentage of the driver output is dissipated in a swamping resistor so that when the grid is driven positive the relative increase in driver loading is small. After careful consideration of these factors one finds that class AB1 operation has much to offer.

Operating Methods

There are four basic methods of operating a linear amplifier as follows:

(i) Grid driven—anode neutralized;
(ii) Grid driven—grid neutralized;
(iii) Cathode driven—grounded grid;
(iv) Grid driven—passive grid.

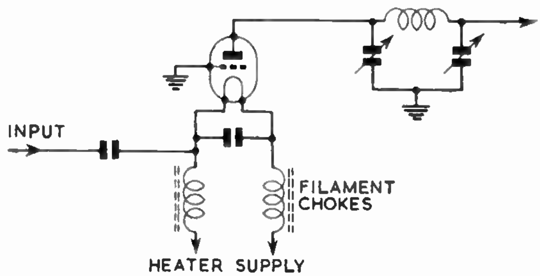


Fig. 10.56 Grounded grid (cathode driven) amplifier.

These basic circuits are shown in **Figs. 10.54, 10.55, 10.56** and **10.57** and in each circuit configuration the valve can be either a triode or a tetrode.

At this stage it might be advisable to consider what exactly is meant by the term *linear amplifier*. To the high fidelity music lover a linear amplifier is intended to give superb quality with the lowest level of distortion. To the single sideband operator a linear amplifier can be added to an existing exciter and will give a more powerful signal without impairing the speech quality. In fact the hi-fi enthusiast and the sideband operator when they talk about a linear amplifier are in fact talking about the same thing. In theory, a good hi-fi linear amplifier can be turned into a good low distortion sideband amplifier by replacing the audio input and output transformer by r.f. tank circuits. In addition, because the flywheel effect of the tuned circuits puts back the missing half cycle it is not even necessary to use two valves in push-pull. Finally, the operating parameters of anode, screen and

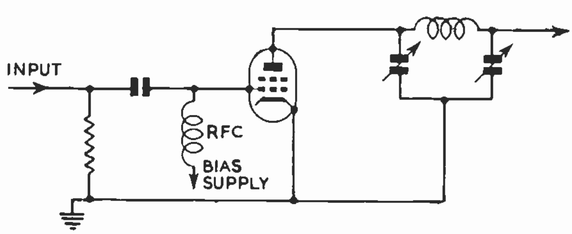


Fig. 10.57. Grid driven passive grid amplifier.

grid supply voltages, anode load resistance, grid driving voltage and power output, supplied by the valve manufacturer for audio operation, apply equally for single sideband r.f. service. An example of this is given in **Table 10.3** for the valve type 811A.

TABLE 10.3

811A Operating Data

	Class B Audio Service (Two valves)	Class B R.F. Service (S.S.B.) (One valve)	
		Grid Driven	Grounded Grid
Anode Voltage	1250	1250	1250
Grid Bias	0	0	0
Peak Grid Voltage	175	88	88
Zero Sig. Anode Current (mA)	54	27	27
Max. Sig. Anode Current (mA)	350	175	175
Load Resistance (ohms)	9200	4600	4600
Max. Sig. Grid Current (mA) *	26	13	13
Power Output (watts)	310†	155†	141‡

*Varies from valve to valve. †Computed power output. ‡Measured including circuit loss.

The operating parameters of a class B amplifier stage remain the same regardless of whether the valve functions in audio or r.f. service. Grounded grid operation is similar, except that the exciter must supply additional feed through power. Since class B audio service requires two valves, all currents and anode load resistance must be halved for single valve r.f. service. Class B audio data is readily available for most valves and can be used for r.f. service as shown above.

Is Linearity Necessary?

For sideband service the envelope of the signal in the anode circuit must be an exact replica of the envelope of the exciting signal. This implies that the power gain of the stage must be constant regardless of the signal level. This desirable basis of operation can only be obtained if the amplifier is operated in a linear manner. Any non-linearity creates distortion products that appear both in the signal passband and in those channels adjacent to it.

A practical place to examine a sideband signal for linearity and quality is in the adjacent suppressed sideband channel, not in the frequency band of the signal itself.

The excellence of a sideband signal is judged by the amount of (or lack of) sideband splatter in nearby channels. Theoretically, a sideband signal should be just as wide as the voice passband of the equipment—3 kc/s—and no wider. If the output signal of a linear amplifier stage is a replica of the existing signal there will be no distortion products—however, valves are not perfect and the transfer characteristics of even the best linear amplifiers exhibit non-linearity at the extremes of the anode current swing. So long as the signal input is a single tone (such as inserted carrier, or a single tone into the microphone socket) departure from linearity has no effect but, if the signal input contains two or more tones, the non-linearity of the power amplifier will cause " mixing " of the signal source and will produce new additional sum and difference frequencies that were not present in the original input signal. These new frequencies, generated in the power amplifier, are known as intermodulation distortion products.

The standard method of testing a linear amplifier to determine the level of distortion products is the two-tone test, in which two radio frequencies of equal amplitude are applied to the amplifier and the output signal is examined for spurious products. Those output signals falling in the harmonic region—" even order " products—are attenuated to a low level by the amplifier tank circuits. Unfortunately, the " odd order " products fall close to the fundamental output frequencies and cannot be removed by tuned circuits. These are the distortion products that put back the signal on the unwanted sideband and in the case of a poorly designed or incorrectly operated linear amplifier, cause objectionable splatter.

Fig. 10.58(a) shows the spectrum distribution of the products generated in a typical p.a. stage, while **Fig. 10.58(b)** shows on an expanded frequency scale those intermodulation products within the amplifier passband that cause s.s.b. distortion. In the example shown the two frequencies making up a typical two tone test are 3748 and 3750 kc/s. If the linear amplifier is perfect these will be the only frequencies appearing in the output. In practice, the amplifier is not perfect and there will be additional combinations of sum and difference frequencies generated by the non-linear transfer characteristics of the valve. These odd order products fall within the passband of the selective output circuits and will be radiated together with the wanted signal. The inside pair of intermodulation products are third order, the next fifth order, seventh order, and so on. It will be noted that those distortion products nearest to the original input frequencies, F1 and F2, have the greatest amplitude. These are the third order intermodulation products and it is the relative amplitude of the third order products in relation to the wanted signal that determines the excellence or otherwise of the transmitted sideband signal.

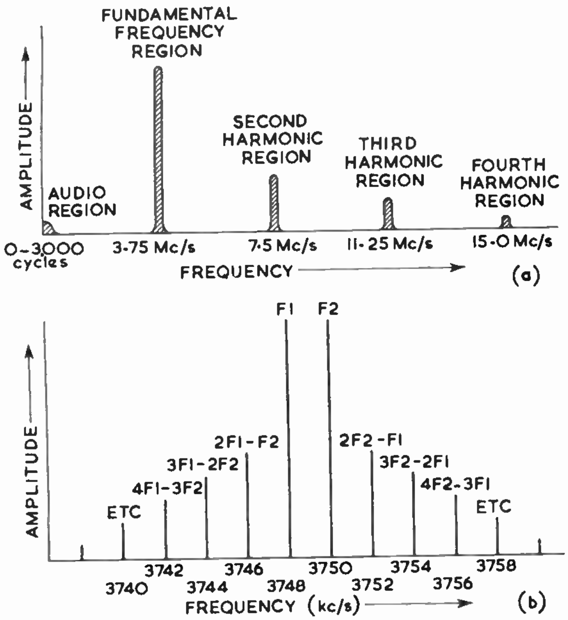


Fig. 10.58. Odd order intermodulation products causing s.s.b distortion. The frequencies shown assume a transmitter with a carrier on 3750 kc/s (radiating lower sideband) and modulated by a 2 kc/s tone input. The amplifier is driven at carrier frequency (F2) by unbalancing the modulator or using carrier insertion. The audio input of 2 kc/s produces the second frequency of 3748 kc/s (F1). Controls are adjusted for equal amplitude of F1 and F2.

Fortunately so far as amateur operation is concerned, it is not necessary to have or be able to use elaborate test equipment. Many s.s.b. operators have selectable sideband receivers with correctly calibrated S meters and they can check the relative signal level on the wanted and on the unwanted sideband, under tone input or voice conditions, and give a report of so many decibels down with a degree of accuracy that is reasonably high and quite adequate for amateur use.

The reduction of adjacent channel interference in the crowded amateur bands is ultimately of benefit to all. There is obviously little sense in aiming for a rejection of at least 35db in the filter and then allowing a badly operated linear amplifier to put the unwanted sideband back again in the form of objectionable intermodulation distortion products. The aim therefore is a distortion product level at least as good as the filter (35db down).

Reducing Distortion

Consideration has been given to the effect on the output signal of non-linearity in the power amplifier and how such non-linearity causes " mixing " and the generation of odd order intermodulation products. It will be noted that the frequency spacing of the distortion products is always equal to the frequency difference between the two original tones. A voice signal is made up of a multiplicity of tones—there will therefore under voice operating conditions be a multiplicity of intermodulation distortion products. These will be present in the transmitted signal and will be heard on the unwanted sideband as blurred and distorted speech that is completely unintelligible—in short as splatter.

When a linear amplifier is improperly adjusted or overdriven the spurious frequencies rise in amplitude and also extend far outside the original channel and will cause unintelligible splatter interference in adjacent channels. Splatter of this type is usually of far more importance than the effect on intelligibility or quality of the original signal. To minimize unnecessary interference the distortion products falling in adjacent channels should be reduced as far as it is possible to get them. Common courtesy on the crowded amateur bands dictates the use of transmitters with as little distortion as the state of the art reasonably permits.

There is clearly no point in going to the effort of constructing a filter that will give 45db sideband suppression and then putting the unwanted sideband back again in the form of distortion products generated in the amplifier stages. A filter with a high level of unwanted sideband suppression deserves a power amplifier with a low level of intermodulation product distortion—one is complementary to the other!

The first and most important means of reducing distortion in a single sideband linear power amplifier is to choose a valve with a good anode characteristic and choose the operating conditions for low odd order curvature. **Fig. 10.59** shows the anode characteristic and the operating point that will allow class AB operation with no odd order distortion products. From point *A* to *B* the curvature is second order or a simple ($I_a = kV_g^2$) curve. From point *B*, the curve continues at the same slope in a straight line to point *C*. The zero signal operating point *Q* is located midway horizontally between *A* and *B*. It is also located directly above the point of projected cut-off, point *P*, where an extension of *CB* crosses the zero anode current line.

Small signals whose peak-to-peak amplitude is less than the horizontal distance between *A* and *B* operate on a pure second-order curve, resulting in no single sideband distortion. When the input signal becomes greater than *AB* it enters a linear region on both peaks at the same time and since the slope of *BC* is correct there is no change in gain of the fundamental components and no single sideband distortion will result at large signals either. The anode current at point *Q* determines the static anode current I_a of the valve and, when multiplied by the d.c. anode voltage, determines the static anode dissipation.

Most valves have a characteristic similar to Fig. 10.59, although *AB* is not a pure square law and the region from *B* to *C* is rather limited and seldom straight. In practice, however, an anode current/grid voltage curve can be plotted from the desired load line on a set of constant current curves, or obtained from the valve manufacturer. By projecting the

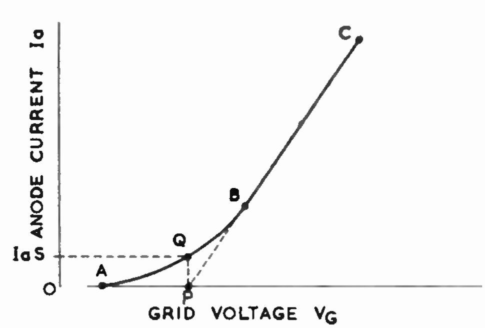


Fig. 10.59. Ideal valve characteristics for class AB operation.

most linear portion of this curve to intersect with the zero anode current line, the point of projected cut-off and therefore the grid bias and static anode current can be determined. This static anode current is the correct value for minimum distortion.

The screen voltage of a tetrode valve has a very pronounced effect on the optimum static anode current because the anode current of a valve varies approximately as the three-halves power of the screen voltage. For example, raising the screen voltage from 300 to 500 volts will double the anode current. The shape of the dynamic characteristic will stay nearly the same; however, the optimum static anode current for minimum distortion is now also doubled. In practice a limit is reached when the higher static anode current and therefore the higher static anode dissipation exceeds the rated anode dissipation for the particular valve in use. Should this condition arise it is necessary to make a choice between operating the valve at lower than optimum static anode current or alternatively reducing the screen voltage.

Although single sideband has now been in use for many years there is not yet any linear power amplifier arrangement that has shown itself to be superior either in performance or characteristics to an extent where it has become accepted as a standard particularly suitable for the requirements of s.s.b.

The type of linear amplifier that would be most suitable is dependent on circumstances that vary from one station to another. In fact, the choice must be made by the operator because it is inherently a personal one. It is possible, however, to give an indirect answer to the question by outlining a simple " design consideration " procedure as follows:

(i) Determine the h.t. supply voltage that is to be used. (This may be from an existing power pack, or may depend on the use of components that are already available, or it may be determined by considerations of safety, i.e., many amateurs fight shy of using voltages much above 750.)

(ii) Determine the driving power that is available from the exciter. (This can be obtained from the manufacturers' data for the driver valve in use, or can be measured into a dummy load, or approximated by lighting a suitable lamp load.)

(iii) By reference to **Table 10.4** chose a valve suitable for the h.t. supply available.

(iv) By reference to **Table 10.5**, decide on the basic method of operation suitable for the driving power that is available.

TABLE 10.4

500V	500-750V	750-1250V	1500-2000V	2000-3000V
2E26 829B	6146 807 1625	TT21 807 1625 805 811A 4-65A 4X150A EL38 QV08-100	813 4-125A 4X150A 304TL 4X250B	304TL 4-250A 4-400A PL-6569

The values given in Table 10.4 and 10.5 are approximations intended to serve as a guide. In Table 10.5 the " Driving Power " includes

TABLE 10.5

Basic Method of Operation

Driving Power (P.E.P.) output	Tuned Grid AB1	Tuned Grid AB2	Passive Grid	Cathode Driven (Grounded) Grid)
5 watts	2E26 6146 4-65A 4X150A 807 1625 QV08-100 TT21 EL38	829B		
10 watts	4-125A 813	807 1625 4-65A 813 4-125A		
25 watts	304TL	805 811A	813 4-65A 4-125A 4X150A	
50 watts		304TL	805 811A 4-250A 4-400A	813 4-125A 4X150A 805 (X2) 811A (X2)
100 watts			304TL PL6569	304TL 811A (X4) 805 (X4) PL6569

(*a*) The grid circuit and coupling losses,
(*b*) The input damping losses of the valve,
(*c*) The loss in the grid circuit swamping resistor,
(*d*) The grid current which may flow,
(*e*) In the grounded grid application, the percentage of driving power that appears in the output circuit as " feed through power." The driving power for the valve shown in the class AB1 column assumes a grid swamping resistor of 2000 ohms.

As the input impedance of a grounded grid amplifier is a function of the peak cathode current, the driving power required will be greater for four valves than for two. The number of valves in use is indicated by ($\times$4) or by ($\times$2). It is assumed in all cases that where two or more valves are used in the amplifier, they are operated in parallel.

Having decided on the type of valve that is most suitable for the h.t. supply and driving power available, the valve manufacturers' data can be consulted for the p.e.p. output rating. If this is less than the total output required it will be necessary to run two, three or four valves in parallel. To give two examples:

(i) The choice of valve type is 6146 with a 750 volts h.t. supply. The required maximum signal power output (p.e.p.) is 200 watts. This will require four valves in parallel.

(ii) The choice of valve type is 813 with a 2000 volt supply. The required output is 400 watts p.e.p. This can be obtained with two valves in class AB2.

Basic Circuit Considerations

As a guide to the choice of basic methods of operation given in Table 10.5, the main advantages and disadvantages can be summarized as follows:

Tuned Grid. Class AB1

Advantages

(*a*) Low driving power.

(*b*) As there is no grid current the load on the driver valve is constant.

(*c*) There is no problem of grid bias supply regulation.

(*d*) Good linearity and low distortion.

Disadvantages

(*a*) Requires tuned grid input circuit and associated switching or plug-in coils for multi-band operation.

(*b*) Amplifier must be neutralized.

(*c*) Lower efficiency than class AB2 operation.

Tuned Grid, Class AB2

Advantages

(*a*) Less driving power than passive grid or cathode driven operation.

(*b*) Higher efficiency than class AB1.

(*c*) Greater power output.

Disadvantages

(*a*) Requires tuned grid input circuit.

(*b*) Amplifier must be neutralized.

(*c*) Because of wide changes in input impedance due to grid current flow there is a varying load on the driver valve.

(*d*) Bias supply must be very "stiff" (have good regulation.)

(*e*) Varying load on driver valve may cause envelope distortion with possibility of increased harmonic output and difficulty with TVI.

Passive Grid

Advantages

(*a*) No tuned grid circuit.

(*b*) Due to relatively low value of passive grid resistor, high level of grid damping makes neutralizing unnecessary.

(*c*) Constant load on drive valve.

(*d*) Compact layout and simplicity of tuning.

(*e*) Clean signal with low distortion level.

(*f*) Simple circuitry and construction lending itself readily to compact layout without feedback troubles.

Disadvantages

(*a*) Requires higher driving power than tuned grid operation.

Cathode Driven

Advantages

(*a*) No tuned grid circuit.

(*b*) No neutralizing. (This may be necessary on 10m.)

(*c*) Good linearity due to inherent negative feedback.

(*d*) A small proportion of the driving power appears in the anode circuit as "feed-through power."

Disadvantages

(*a*) High driving power—greater than the other methods.

(*b*) Isolation of the heater circuit with ferrite chokes or special low capacity wound heater transformer.

(*c*) Wide variation in input impedance throughout the driving cycle causing peak limiting and distortion of the envelope *at the driver valve*.

(*d*) The necessity for a high-*C* tuned cathode circuit to stabilize the load impedance as seen by the driver valve and overcome the disadvantage of (*c*).

(*e*) In practice, with the type of valves commonly used by United Kingdom amateurs, the power output appears to be the same or slightly less than that for passive grid operation. The active anode current flows through the cathode circuit and produces across the cathode impedance a voltage which decreases the exciting voltage. This corresponds to negative feedback, and it is possible that the loss due to this feedback can be roughly equal to the feed-through power—the net advantage is then zero.

Beam power valves such as the 807, 1625 and 813 may be connected as high-mu triodes in the manner shown in **Fig. 10.60.** The triode-connected tetrode usually requires a much higher excitation voltage than a conventional triode, but is otherwise capable of similar performance. It is important, however, to remember that this high excitation voltage appears between the grid and cathode, and there is no grid-bias voltage to be overcome before grid current can flow. As a result of the large driving voltage and high current flow, triode-connected tetrodes often operate with control-grid dissipation powers in excess of the manufacturers' ratings. This may happen inadvertently if the grid circuit has only a single current meter because up to 75 per cent of the combined grid current may be to the control grid in such a circuit. Should the valve be cathode driven in a grounded grid circuit, with insufficient anode circuit loading, the power formerly fed through the amplifier into the output circuit becomes available to heat the control grid to even higher temperatures. Such action can destroy the valve in a short time.

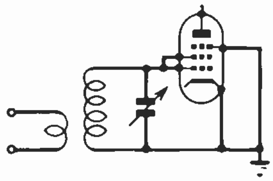


Fig. 10.60. High-mu triode connection of a tetrode or pentode amplifier valve.

While it is possible to protect the valve with a grid-current overload relay of low coil resistance and sufficient sensitivity, it is hardly likely that a suitable component would be available to the amateur constructor. Conventional tetrode connection (either grid driven or cathode driven) will avoid the risk of damage to the valve and is the preferred method.

Push-pull or Parallel Operation

In theory two valves operated in linear amplifier service, either in push-pull or in parallel, will give the same maximum signal power output. However, there may be certain *practical* considerations that make one method more desirable than the other.

A single sideband amplifier must operate at all times in the most linear manner if objectionable intermodulation distortion is to be avoided. This means that the valve (or valves) must operate into a precise value of anode load: this may vary in value from a few hundred ohms up to several thousand ohms depending on the valve type, operating potentials and required maximum signal power output. The anode load (R_L) is then required to be stepped down in impedance to the design value of the amplifier output, usually 75 ohms. From this it follows that the tank circuit is arranged as a step-down transformer.

As a precaution against the possibility of TVI it has become accepted practice to feed the transmitter output via a screened cable into a low-pass filter (capable of 40db or more attenuation to any r.f. in the television channels) and the filter either directly or via an aerial tuning unit into the aerial system. As it is reasonably cheap and easy to obtain, coaxial cable is generally used, and modern transmitters are therefore designed to have a tank circuit with an unbalanced output.

Of further advantage is some simple means—preferably from a panel control—of varying the loading so that the amplifier operation can be adjusted while the output waveform is monitored on an oscilloscope.

The old convention of plug-in tank coils is now considered to be completely out-of-date, and modern transmitters are designed for rapid changes from band to band by means of switching. It obviously would be quite a complex electrical and mechanical operation to arrange switching for five push-pull anode coils and five output link windings.

For these reasons the pi-network circuit has become deservedly popular; the anode tuning capacitor need not be a split stator type, inductance change is easily effected by suitable taps on a single coil, and the required switch is a simple single pole type. It is easy to see why in modern transmitters the pi-network is almost universally used. As both the input and the output of the tank circuit is single ended, two or more p.a. valves are always operated in parallel.

Finding the Value of the Anode Load

A large number of valve types are suitable for linear amplifier use but in many cases manufacturers' figures for single sideband service are not available. It is therefore necessary for the constructor to work out the valve operating conditions from first principles. This is particularly important in regard to the value of R_L and the values of L and C in the tank circuit. The amplifier can only give its rated power output without distortion if it is working into the correct anode load corresponding to the dynamic operating conditions and the load line that has been selected.

One "rule of thumb" formula that can be used to find the value of anode load (R_L) where the makers' figure for single tone anode current is known, is $V_a/2 \times I_a$, where V_a is the d.c. supply voltage and I_a is the maximum signal anode current in amps. This formula cannot be more than an approximation because it does not differentiate between the different classes of working. The correct load for class AB1 is not the same as the load for class AB2 or class B working.

The recommended procedure that will give an accurate answer for the value of R_L is the formula based on the operating conditions and taking into consideration the angle of anode current flow. This is $R_L = 2 \times (V_a - V_{a\,min}) / I_a \times K$, where V_a is the d.c. supply voltage, $V_{a\,min}$ the minimum value of anode voltage at the crest of the cycle when the anode current reaches its maximum value, I_a is the manufacturers' figure for maximum signal anode current and K is a constant whose value is dependent on the angle of anode current flow. For class B operation the angle of flow will be approximately 180° and $K = 3{\cdot}14$. For class AB operation the angle of current flow will be greater than 180° and K will have a value that is smaller than 3·14. For 200° the approximate value will be 2·88. The linear amplifier can be analysed and designed as a class C amplifier in which the angle of anode current flow is slightly greater than 180° and the value of $V_{a\,min}$ is slightly higher than for class C operation.

Zero bias triodes such as the 805, 811A and TZ40 are designed to have a low zero signal anode current and be operated in class B. Conversely, many high slope tetrode or pentode valves are capable of high power output without grid current and are normally driven in class AB1. Examples of the latter are 6146, 4-125A, 4X150 and 813.

A typical load line and the method of plotting is shown in

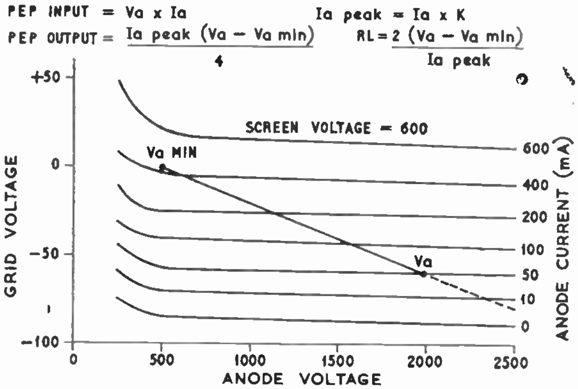


Fig. 10.61. Graphical determination of valve operating conditions. Typical constant current characteristics, showing how the load line is plotted. V_a is the zero signal point, showing that the resting anode current is 50 mA, the grid bias −60 volts, and the d.c. h.t. supply 2000 volts. $V_{a\,min}$ is the instantaneous value of anode voltage at a peak anode current of 450 mA (I_a Peak). This occurs at the crest of the grid driving cycle when the r.f. voltage reaches a peak value of +60 volts; the effective bias value is then zero as shown. (I_a is the d.c. anode current meter reading at maximum signal.)

Fig. 10.61. The instantaneous anode voltage should not be allowed to drop much below the applied screen voltage in order to avoid excessive screen current and the risk of exceeding the makers' figure for the maximum permissible screen dissipation. Normally, $V_{a\,min}$ is given a value slightly less than the recommended screen voltage for the valve. To take the popular 813 as an example, operated in class AB1, with 2000 volts h.t. and 750 volts on the screen, the maximum signal anode current is 150 mA. Using the formula,

$$R_L = \frac{2 \times (V_a - V_{a\,min})}{I_a \times K}$$

and substituting the values, this gives $\dfrac{2\,(2000 - 600)}{0{\cdot}15 \times 2{\cdot}88} = 2800/0{\cdot}43 = 6500$ ohms.

Where the Maximum Signal Anode Current is not Known

A number of valve types—designed for some other application—are available as low cost initial equipment and are therefore attactive to the amateur for linear amplifier operation. In many cases, however, operating data for r.f. or audio use is unobtainable. In these cases it is necessary to design from first principles. The procedure can perhaps most clearly be shown by taking a specific example. An excellent valve available at low cost is the EL38, a high slope pentode developed as a television receiver line time base output valve capable of operation at high anode voltage and high peak anode current—characteristics particularly suitable for s.s.b. amplifier use.

It will be assumed that the valve is to be operated in class AB1 with an expected efficiency of 60 per cent. Inspection of the manufacturers' data for limiting values gives V_a 800 volts, $V_{a\ peak}$ 8 kV and V_{g2} 400 volts. (Suitable values of anode and screen supplies for amateur service would be 1000 and 300 volts.) Since the 40 per cent power loss must equal the maximum rated anode dissipation of 25 watts, the total power input (100 per cent) must be $25 \times 100/40 = 62{\cdot}5$ watts.

Dividing the maximum power input by the anode voltage gives a maximum signal anode current I_a of 0·0625 amp., and $I_{a\ peak} = 0{\cdot}0625 \times 2{\cdot}88$ (K is 2·88 for class AB operation) $= 0{\cdot}18$ amp.

Assuming that the instantaneous anode voltage is allowed to swing down just below the value of the screen voltage $(V_a - V_{a\ min}) = 800$ volts and $R_L = 1600/0{\cdot}18 = 8888$ ohms.

The p.e.p. input $= V_a \times I_a = 1000 \times 0{\cdot}0625 = 62{\cdot}5$ watts and the p.e.p. output $= I_{a\ peak} \times (V_a - V_{a\ min})/4 = 0{\cdot}18 \times 800/4 = 36{\cdot}0$ watts. To check the figures the output power is subtracted from the input power, $62{\cdot}5 - 36 = 26{\cdot}5$. This is slightly more than the rated anode dissipation but is quite satisfactory for amateur sideband use. (If a greater power output than this is required it would be satisfactory to increase the h.t. supply to say 1250 volts.)

Four EL38 valves in parallel with a 1000 volts h.t. supply would make an excellent linear amplifier running at 250 watts input and a p.e.p. output of 150 watts. The correct value of R_L would be one quarter the value of one valve—$8888/4 = 2222$ ohms.

The power output obtainable from a linear amplifier at a given anode voltage is determined by the peak anode current $I_{a\ peak}$; this in turn is determined by the minimum anode voltage $V_{a\ min}$. A large $I_{a\ peak}$ is required in order to obtain large output, and a small $V_{a\ min}$ in order to have a good efficiency. With $I_{a\ peak}$ and $V_{a\ min}$ determined by considerations of power output and anode efficiency, respectively, it is required that R_L have the value given by the formula

$$R_L = \frac{2 \times (V_a - V_{a\ min})}{I_{a\ peak}}.$$

If R_L is smaller than it should be, then $V_{a\ min}$ is thereby increased and both efficiency and output power suffer.

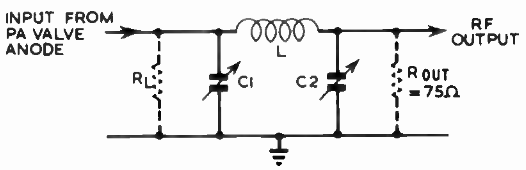


Fig. 10.62. Basic pi-network tank circuit. R_L required anode load; C1, anode tuning capacitor; L, inductance; R_{out}, load into which the transmitter is required to work (normally 75 ohms); C2, aerial loading capacitor.

If R_L is made too large, $V_{a\ min}$ becomes very small and with triodes this causes $I_{a\ peak}$, and therefore the output power, to decrease; with pentode and tetrode valves a virtual cathode forms if $V_{a\ min}$ is too small causing secondary emission from the anode to the screen, excessive screen current and a flattening of the anode current peaks. In the examples given $V_{a\ min}$ is determined by the operating voltage on the screen. In the case of triode valves $V_{a\ min}$ is limited to a value that is more positive than the peak r.f. grid voltage.

A summary of the operating conditions for four valves in parallel would be as follows:

Operating Conditions for Four EL38

D.C. anode voltage	1000 volts
D.C. screen voltage	300 volts
Zero signal d.c. anode current . .	80 mA
Max. signal d.c. anode current . .	250 mA
Effective load resistance	2220 ohms
D.C. grid voltage	value required to give 80 mA standing anode current
Maximum signal power input . .	250 watts
Maximum signal power output . .	150 watts (approximately)

The amplifier would be loaded and the drive adjusted with single tone input (audio tone or inserted carrier) to the maximum signal anode current of 250 mA. On speech the p.a. anode current meter would not be allowed to swing beyond half this value to prevent overdriving and distortion. In general the linearity of a tetrode or pentode amplifier is improved by running the zero signal anode current as high as possible without exceeding the rated anode dissipation—80 watts is a good compromise value.

The Pi Network Tank Circuit

The Q of the anode circuit, of which the tank is a part, must be sufficient to keep the r.f. anode voltage close to a sine wave shape. Because of the greater angle of anode current flow the requirements for linear amplifier operation are less stringent than for class C operation. However, if the anode circuit Q is insufficient the r.f. waveform may be distorted resulting in low anode efficiency and also poor attenuation of the harmonics of the output signal. Too high a value of Q results in large circulating r.f. currents and power loss. A compromise value giving a good balance between the conflicting requirements and fully sufficient for sideband working is a Q of 12.

This is the recommended type of output circuit, and once the correct value of R_L has been calculated as shown, the required C and L values can be read off directly from the tables given in the excellent and highly recommended article *Simplified Design Procedures for Pi-Network Tank Circuits* by G. C. Fox, A.M.I.E.E. (G3AEX) in the June 1961 issue of the BULLETIN and reproduced in the RSGB *Radio Data Reference Book*.

Simplified Calculation Procedure

An alternative to the use of pi-network abacs is a simplified design procedure first suggested by G. R. B. Thornley (G2DAF). The basic pi-network tank circuit is shown in Fig. 10.62.

As the pi tank circuit is an impedance matching device the first requirement is to find the reactance values for C1, L and C2 in relation to the tank Q.

$$\text{The tank circuit impedance ratio} = \frac{Rl}{R_{out}} \qquad \text{(i)}$$

$$\text{The tank circuit reactance ratio} = \frac{Cl}{C2} = \sqrt{\frac{Rl}{R_{out}}} \qquad \text{(ii)}$$

$$XCl = Rl \text{ divided by } Q = \frac{Rl}{Q} \qquad \text{(iii)}$$

$$XC2 = \frac{XCl}{\sqrt{\frac{Rl}{R_{out}}}} \quad \text{(ii)} \qquad \text{(iv)}$$

$$XL = XCl \textit{ plus } XC2 \qquad \text{(v)}$$

Considering the case of a pi-network tank designed for an R_L of 3000 ohms and a Q of 12, the numerical values would be as follows:

$$\frac{R_L}{R_{out}} = \frac{3000}{75} = 40 \ldots\ldots\ldots\ldots \text{(i)}$$

$$\sqrt{\frac{R_L}{R_{out}}} = \sqrt{40} = 6{\cdot}3 \text{ (approx.)} \ldots\ldots\ldots \text{(ii)}$$

$$XC_1 = \frac{R_L}{Q} = \frac{3000}{12} = 250 \text{ ohms} \ldots\ldots\ldots \text{(iv)}$$

$$XC_2 = \frac{XC_1}{\sqrt{\frac{R_L}{R_{out}}}} = \frac{250}{6{\cdot}3} = 40 \text{ ohms} \ldots\ldots\ldots \text{(iv)}$$

$$X_L = XC_1 + XC_2 = 250 + 40 = 290 \text{ ohms} \ldots \text{(v)}$$

These values are a simple approximation, but are quite near enough for amateur purposes. From a reactance chart the values for 80m are:

$C_1 = 180$ pF; $C2 = 1100$ pF; $L = 14\,\mu$H.

The values for the other bands scale down in the same ratio as the band wavelength as follows:

Band	*Cl*	*L*	*C2*
80m	180 pF	14 μH	1110 pF
40m	90 pF	7 μH	550 pF
20m	45 pF	3.5 μH	275 pF
15m	33 pF	2.5 μH	205 pF
10m	22 pF	1.7 μH	137 pF

If more detailed pi tank circuit formulae are required, the subject is covered more fully in Chapter 6 *H.F. Transmitters.*

Pi Coil Winding Data

While it is possible to wind a pi-tank coil and get it to resonate correctly on each band—with the calculated values of C1 and C2—by a process of trial and error, this is a time consuming method. A much more satisfactory procedure is a method of relating the inductance values in μH to actual turns of wire on a coil former of suitable diameter. Amateurs are indebted to R. G. Wheatland, G3SZW for providing the practical formulae* given in Table 10.6.

TABLE 10.6

Former Diameter	Winding Length	Number of Turns	Wire Gauge and Spacing
(A) 1½ in.	2 in.	$8{\cdot}6\sqrt{\frac{L}{d}}$	18 s.w.g. spaced 16 turns per inch.
(B) 1¾ in.	3½ in.	$10\sqrt{\frac{L}{d}}$	16 s.w.g. spaced 12 turns per inch.
(C) 2½ in.	3¾ in.	$9{\cdot}5\sqrt{\frac{L}{d}}$	14 s.w.g. spaced 8 turns per inch. (Eddystone former.)
L = Inductance in μH. d = Former diameter in inches.			
Former " A " is suitable for two 6146 valves or similar. Former " B " is suitable for two TT21 valves or similar. Former " C " is suitable for two QY3-125 valves or similar.			

With the transmitter output connected to a 75 ohm non-inductive dummy load, check that each band resonates with the correct values of C1 and C2; it may be necessary to adjust the tapping points to achieve this.

THE ADVANTAGES OF PASSIVE GRID OPERATION

The term " passive grid " is used to define a method of operating a linear power amplifier in which the grid input circuit is " passive " in regard to frequency, i.e. the normal coil and tuning capacitor are omitted and replaced with a non-inductive resistor. This method has a number of advantages making it particularly suitable for amateur operation.

Variation of Input Damping

It is of interest to consider the basic problems that are common to the normal grid driven method. The first is the variation of input damping. Throughout the drive cycle, in a radio frequency amplifier biased to class AB1, the input damping changes and these changes are reflected back to the driver valve, as shown in **Fig. 10.63.**

A method of overcoming this distortion is to make the driver impedance as low as possible. This can be accomplished by connecting a swamping resistor across the grid circuit. The value of this resistor depends on the valve being used, the operating conditions, and possibility of grid current, but is usually of the order of 2000 ohms. The characteristics of certain valve types are such that it is advantageous to drive the valve slightly into grid current. In this instance a swamping resistor will assist in masking the effects of input impedance change when grid current starts to flow.

Amplifier Stability

The second problem is that of amplifier stability. It is vitally important that there is no positive feedback in a linear power amplifier and most tetrode and pentode valves used for amateur sideband use require neutralizing. Because of the high power gain it is more difficult to neutralize correctly a class AB amplifier than a class C one. The variation in grid cathode capacity can also cause an amplifier that is perfectly stable and correctly neutralized under static conditions to become imperfectly neutralized and become unstable under voice peak conditions.

Heavy grid damping will prevent any feedback due to the grid-anode capacity of the valve being of sufficient amplitude to cause instability, without any necessity to incorporate neutralizing circuits. In addition the low grid resistor will provide a constant load on the driver valve and will effectively damp out the effect of variation of input damping during the driving cycle caused by input capacity change, and the further effect caused by the onset of grid current flow.

* " L. C. Calculations " by R. G. Wheatland, G3SZW, RSGB Bulletin August 1965.

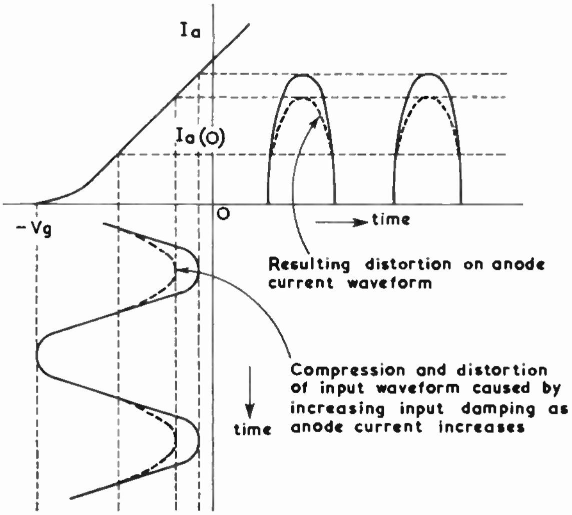


Fig. 10.63. Linear amplification characteristics showing how variation of input damping causes distortion of the input waveform.

The Passive Grid Linear Amplifier

While it is possible to stabilize a triode amplifier by making the passive grid resistor of low enough value it is likely that in practice the required value would be so low, an excessive amount of driving power would be necessary in order to develop the necessary grid driving voltage across it. The passive grid method is therefore particularly recommended for tetrode and pentode amplifier valves.

The lowest value of grid resistor that will normally be used is 75 ohms—this would be a good match to an exciter with a pi-output network usually working directly into a 75 ohm aerial load. The highest possible value of grid resistor will be that at which the amplifier becomes unstable due to positive feedback; this will vary from valve to valve but is estimated to be around 1000 ohms.

As a low value will give a greater measure of stability and a more constant load on the driver valve, its value in ohms will be determined by the required maximum signal grid driving voltage and the available power output from the driver valve. This is determined from the formula $R = V^2/P$, where V is the maker's value for the peak r.f. grid driving voltage and P is the p.e.p. output rating of the driver valve in use. It should be noted that as p.e.p. is an r.m.s. value. the value of V used in the formula should also be an r.m.s. value. However, the use of the peak voltage figure normally given in valve data ensures that in practice the required drive voltage is developed using only half of the available driver output. This gives a very desirable two to one margin of safety and ensures that the driver stage can never be overrun at any time, and that the level of intermodulation distortion products from this stage is always less than the distortion product level from the linear power amplifier.

Problems of Ratings

It is customary to rate a.m. transmitters on the power developed in the carrier and most published data on valves for a.m. telephony gives operating conditions which take into account the peaks which occur when 100 per cent modulation is applied. With single sideband operation it is not possible to have a similar form of rating since at zero modulation there is no output from the valve (this assumes complete suppression of the carrier).

Looking at the A.M. Modulation Envelope

It will be assumed that a transmitter set up to the maximum licensed input of 150 watts is anode modulated with 75 watts of audio (the usual amateur conditions). Further, that the overall efficiency of transfer of r.f. output into the load is 66·6 per cent. The load is a non-inductive resistor of 100 ohms. If the vertical deflection plates of a cathode ray oscilloscope are connected across the dummy load the vertical

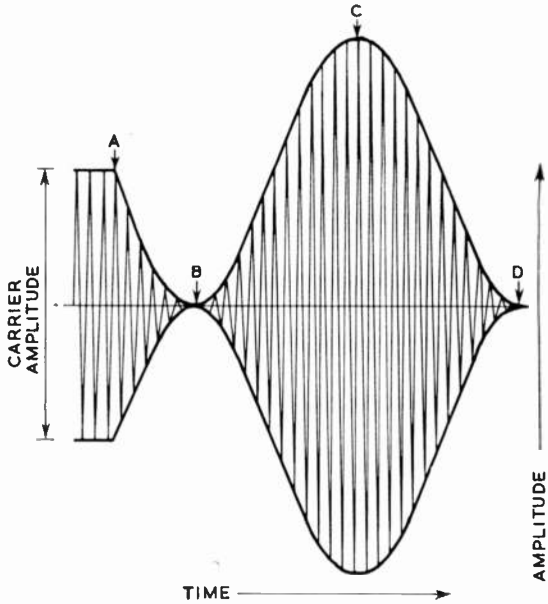


Fig. 10.64. Graph showing how the cycles of r.f. produced by the resultant voltages of the carrier and the two sidebands are continually changing in amplitude from zero at the modulation trough at point *B* to twice the carrier value at the modulation crest at *C*.

deflection will be a measure of the r.f. voltage appearing across the load. In the unmodulated condition the r.f. output power is 100 watts, the voltage across the load is 100 volts and the current flowing through the load is 1 amp. The modulator is now driven to the 75 watt output condition by applying a 1 kc/s sine wave to the microphone input socket. This will fully modulate the r.f. carrier and the cathode ray tube trace will double in amplitude. If the oscilloscope horizontal time base is switched on and the speed adjusted to some multiple of the 1 kc/s modulating frequency, the r.f. modulation envelope will be displayed.

The carrier and the sidebands are quite separate from each other and can be received independently. As all the component frequencies of the transmitted wave are r.f. (two sidebands and a carrier) on different frequencies. they get " in and out of step with each other " and the resultant is the modulation envelope.

The diagram in Fig. 10.64 is a graph of voltage plotted against time—it would be equally valid if it were a graph of current through the 100 ohm load plotted against time. It is convenient to think in terms of voltage because the oscilloscope is the only instrument that will show the transmitter output as a visual presentation, and an oscilloscope is a voltage operated device. Considering the single audio cycle of the modulation envelope shown in Fig. 10.64, at the left hand side of the diagram at point A the carrier voltage is 100 volts and the current through the load is 1 amp. One-four-thousandths of a second later at point B, the r.f. cycles are zero—the voltage across the load and the current through it are also zero. During the next two-thousandths of a second the r.f. cycles increase in amplitude until they reach the crest of the envelope at point C. During this period of time the individual r.f. cycles occurring have twice the original carrier amplitude and the voltage across the load has doubled, therefore as $P = E^2/R$ the power output of each cycle of r.f. energy at the modulation crest is four times the unmodulated carrier power—this is the peak envelope of the transmission, the p.e.p. output, and this is 400 watts. It is also important to appreciate that the p.e.p. output is a real power r.m.s. value. The voltage across the load is 200 volts r.m.s., and the current through it is 2 amps r.m.s. *The product of these two is 400 watts of effective power.* One-two-thousandths of a second later the cycles of r.f. have reached point D in Fig. 10.64, and the transmitter output is again zero.

The method of finding the effective (r.m.s.) value is to take the peak r.f. values at many points over a complete cycle of audio alternation, square these values and then find their average, finally taking the square root of the value thus obtained. The part of the diagram between points B and D in Fig. 10.64 is one complete cycle of the modulating frequency and is already divided into 25 cycles of r.f. It is then convenient to measure the individual lengths of these, square all the values obtained, add them together and divide by 25 to find the average and then calculate the square root. This value will be found to be exactly 1·225 times the carrier amplitude, i.e. the effective (r.m.s.) value of the individual cycles of r.f. energy occurring within each cycle at the modulating frequency is 1·225 times the unmodulated carrier value, therefore an r.f. ammeter indicating 1 amp of current through the 100 ohm load will be expected to indicate under 100 per cent modulation conditions, 1·225 amps, and this is exactly what it does do.

The output power is given by the formula $P = I^2R$, and substituting the values this becomes 1·225 squared + 100 = 1·5 × 100 = 150 watts.

The most satisfactory method of giving the operating conditions for valves designed especially for linear amplifier sideband service is that of p.e.p. input and output ratings.

Output Power of a S.S.B. Transmitter using a Two Tone Test Input

50 ohm dummy load (R)			75 ohm dummy load (R)		
Current (I) (amps)	Mean Power output (watts)	P.E.P. output (watts)	Current (I) (amps)	Mean Power ouput (watts)	P.E.P. output (watts)
0·5	12·5	25	0·5	19	38
1·0	50·0	100	1·0	75	150
1·5	112·5	225	1·5	168·75	337·5
2·0	200	400	1·63	200	400

POWER MEASUREMENTS

Extract from the UK licence conditions governing s.s.b. operation

Suppressed or reduced carrier single sideband systems

The radio frequency output peak envelope power under linear operation from an A3A or A3J transmitter must not exceed that from an A3 transmitter working at an overall efficiency of 66 per cent when supplied with the appropriate maximum permitted d.c. input. The output power shall be measured, using an oscilloscope, by the following process:

(i) Adjust the A3 transmitter output stage for class C working and apply a pure sinusoidal tone to the transmitter. With the d.c. input power limited to the maximum value appropriate to the frequency band concerned note the peak-to-peak deflection on the cathode-ray oscilloscope.

(ii) Adjust the transmitter for single sideband linear operation and replace the tone by speech; the maximum deflection on the cathode-ray oscilloscope, showing the r.f. output caused by the peaks of speech, should not be greater than twice the previously measured deflection obtained with tone input.

As an alternative the following method may be used:

Suppressed or reduced carrier single sideband operation

The radio frequency output peak envelope power must not exceed that from an A3 transmitter working at an overall efficiency of 66 per cent when supplied with the appropriate maximum permitted d.c. input power. The output power shall be measured, using a resistive dummy load, r.f. ammeter or voltmeter and oscilloscope, by the following method:

(i) Apply two non-harmonically related sinusoidal tones* of equal amplitude to the s.s.b. transmitter, with the carrier fully suppressed, and adjust the input power to give a mean radio frequency output power under linear operation of 200 watts (see Note 1) when measured into a resistive load by means of an r.f. meter (see Note 2). Under this condition note the peak-to-peak deflection on the cathode-ray oscilloscope (see Note 3).

(ii) Replace the tone by speech; the maximum vertical deflection on the cathode-ray oscilloscope shall not be greater than the previously recorded deflection obtained with the two-tone input.

Note (1) 200 watts mean radio frequency output power in the case of those bands limited to a maximum d.c. input power of 150 watts; 66⅔ and 13⅓ watts for those bands limited to a maximum d.c. input power of 50 watts and 10 watts respectively.

Note (2) In the case of v.h.f. and u.h.f. measurements the r.f. meter may be replaced by a crystal rectifier and calibrated meter; for s.h.f. measurements a bolometer may be used.

Note (3) In the case of v.h.f., u.h.f. and s.h.f. measurements, this use of an oscilloscope may not be practical. In this case the test may be limited to a measurement of the mean radio frequency output power as outlined in part (i) of the procedure.

* A two-tone oscillator is described in Chapter 19—Measurements.

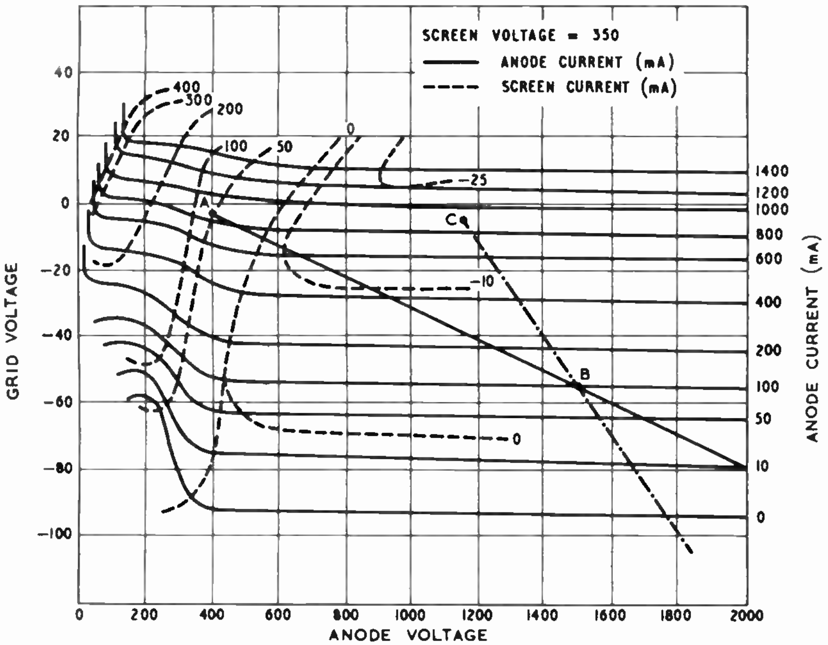


Fig. 10.65. Typical constant current characteristics of the Eimac 4CX250B Valve.

The mean r.f. output power of 200 watts is equivalent to 400 watts p.e.p. output when using a two tone test signal and can be measured with an r.f. ammeter in series with a resistive dummy load of known value. The power dissipated in the load is $I^2 R$; i.e. the current squared multiplied by the value (in ohms) of the load, therefore the current in a 75 ohm load when running at maximum allowable output will be 1·63 amps. For a load of 50 ohms the current will be 2 amps; other current readings can readily be converted to mean output power by the formula given above.

R.F. AMPLIFIER ADJUSTMENT AND LOADING

The tetrode amplifier has the advantage of high power efficiency, together with good linearity and low harmonic output. It has become deservedly popular over the years as an s.s.b. power amplifier. For these reasons the following discussion in regard to r.f. amplifier adjustment and loading will be based on class AB1 tetrode operation.

Fig. 10.65 shows a set of constant current characteristics for a typical tetrode valve. These curves show the dynamic characteristics of the valve—that is, the instantaneous values of anode and screen current for any given grid and anode voltage conditions.

Typical operating conditions for the 4CX250B valve are: V_a 1500 volts; V_{g2} 350 volts; V_{g1} −55 volts; I_a (*zero signal*) 100 mA. These values show that with an h.t. supply to the anode of 1500 volts and a screen supply of 350 volts, the valve will require −55 volts grid bias and will draw a standing anode current of 100 mA. From these figures it is possible to mark the zero-signal operating point of the valve, corresponding to $V_a = 1500$ and $V_{g1} = -55$ volts; this is shown in Fig. 10.65 by the point *B*. This is the point at which the valve rests with zero-signal r.f. grid drive.

At resonance the pi-tank circuit is a pure resistive load R_L and this load can be represented by a straight line passing through point *B* and having a slope determined by the ratio, voltage-swing/peak-current.

As it is desired to operate in class AB1 and not draw grid current, the control grid must not be driven positive and the peak r.f. drive must be restricted to a value that is slightly less than the bias voltage—it will be assumed that the peak r.f. grid drive is held at 100 volts peak to peak (e.g. + 50 volts on the positive half of the driving cycle, and −50 volts on the negative half of the driving cycle). When the peak grid drive is applied, the first positive half cycle will carry the valve operating point along the load line from *B* to *A* and back to *B* again. During this half cycle the grid-voltage swing from −55 volts up to −5 volts and back again to −55 volts, has caused the valve anode current to swing from 100 mA up to 1000 mA and back to 100 mA again. At the same time the anode voltage swings from 1500 volts down to 400 volts. During the negative half cycle the control grid voltage will swing from −55 volts down to −105 volts and the valve operating point will move down the load line from *B* to a point opposite −105 on the grid voltage scale and back to *B* again. The negative going grid voltage therefore swings the anode current down to cut-off for a small portion of the cycle, and the anode voltage continues rising up to 2600 volts and back down again due to the fly-wheel action of the anode tank circuit. This half of the load line need not be plotted and is not important because the valve does not " work " during the negative half cycle.

The load line represents in graphical form the dynamic characteristics of the valve—that is the instantaneous anode current and voltage at any moment of time—and it represents in graphical form the load (R_L) into which the valve is working. Its value in ohms is therefore the *total* voltage swing of the valve anode (1100 × 2) divided by the maximum instantaneous anode current (1000 mA). Expressed as a formula this is; $R_L = 2(V_a - V_{a\,min})$ divided by $I_{a\,peak}$. From this it is seen that the *slope* of the load line is determined by the ratio, voltage-swing/peak-current. A higher value of peak current would move point *A* higher up the graph and the slope *A* – *B* would be steeper. A lower value of $I_{a\,peak}$ would bring point *A* farther down the graph and the slope *A* – *B* would be nearer to the horizontal.

Driving and Tuning

From inspection of Fig. 10.65 it is now possible to make the following deductions:

(i) *Reducing drive will reduce the length of the load line.* (If the r.f. grid drive is reduced to half, the grid voltage swings to only half the original peak-to-peak amplitude; the operating point *B* remains the same but the load line is reduced to half its original length.)

(ii) *Detuning the tank circuit will tilt the load line in a clockwise direction.* The load line will therefore have *minimum* slope at resonance. Note, however, that the moving point will still intercept essentially the same anode current values. It is clear from this that the anode current in a tetrode is not a good indicator of resonance (very little dip).

It will be seen from Fig. 10.65 that the constant screen current lines are concentrated in the upper left hand side of the graph and are tilted upwards at a steep angle. Also of note is the fact that the screen current consists of zero or even negative values in the off-resonance position (*C* – *B*), but at resonance is always positive. From this it is now possible to make the third deduction:

(iii) *A peak in screen current indicates tank circuit resonance.* As the load line is confined vertically by the constant peak-to-peak amplitude of the grid driving voltage (two imaginary horizontal lines, one at –5 volts and one at –105 volts) during the rotation of the load line while tuning, its length increases as resonance is approached and reaches a maximum at resonance. As point *A* penetrates the heavy screen-current region the d.c. screen current rises and the screen current meter indicates a sharp peak at resonance.

Loading

Once the tank circuit is tuned to resonance it presents a pure resistive load to the valve. However the value of this load is affected by the coupling to the external load (the aerial); increased coupling lowers the value of R_L and the load line assumes a steeper angle. At this steeper angle the load line will not intercept the heavy screen current region to the same extent as it did before and the screen current will reduce. From this it is now possible to make the fourth deduction:

(iv) *Screen current will fall as loading is increased.* During the r.f. driving cycle the valve operating point is continually moving through many different instantaneous values of screen and anode current. The average of all these values is what the d.c. meter in the circuit reads. For a linear amplifier valve operated in class AB1 the d.c. meter reading is approximately one-third of the peak value of current at the top of the load line.

Tuning-up Procedure

It is now clear that the screen current meter is by far the best indicator both of resonance and loading with a tetrode amplifier and should always be used in preference to the anode current meter. Of great importance is the realization that a linear amplifier loaded for maximum r.f. output indicated on an aerial ammeter or forward power meter is *not* sufficiently loaded to prevent flat topping on speech peaks. Loading must always be set to obtain a predetermined value of screen current under single-tone driving conditions to obtain as nearly as possible a given set of data sheet conditions as given by the valve manufacturer. If data sheet conditions are not available it is essential to examine the modulated envelope on an oscilloscope.

Due to the inertia of the movement the anode current meter cannot follow at syllabic rate. It is therefore normal on voice peaks for the anode current meter to read no more than half the true maximum value. This means that an amplifier should never be talked up to more than half of the maximum signal single-tone anode current reading.

Tuning-up should be undertaken with the transmitter output connected to a non-inductive dummy load and after matching the manufacturer's data sheet conditions the output should be transferred to the aerial feeder. A change in meter readings indicates a small amount of mismatch (standing wave on the line) and it should be possible to correct this with a further small adjustment to anode tuning and loading capacitors. The amplifier is now ready for speech operation, and after removing the carrier or the inserted tone and re-connecting the microphone, talk into the microphone in a normal operating voice level and adjust the exciter r.f. drive control for the highest level that is possible without drawing grid current on voice peaks or flat topping (check this with the oscilloscope).

The Two-tone Envelope

A single tone input to an s.s.b. transmitter drives the linear amplifier at one frequency. The amplifier output is a pure c.w. signal exactly the same as the output of a telegraph transmitter under key-down conditions. As such it is possible to ascertain by meter readings the performance of the amplifier at maximum signal (p.e.p.) conditions. Information on the linearity of the amplifier is, however, lacking.

In order to study linearity thoroughly by observation of the amplifier output, some means must be provided which will vary the output level from zero to maximum signal with a regular pattern (Fig. 10.66) that is easily interpreted. A simple means of obtaining an output signal is to use two audio tones of equal amplitude to modulate the sideband transmitter. The resultant (or beat between the two r.f. signals) produces a regular pattern which, when observed on

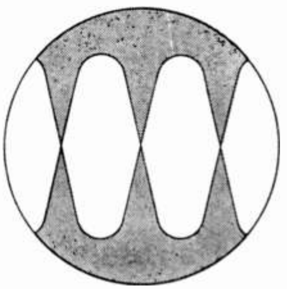


Fig. 10.66. Oscilloscope pattern of a two-tone test signal used for aligning linear amplifiers.

an oscilloscope, has the appearance of a carrier 100 per cent amplitude modulated by a series of half sine waves, as shown in Fig. 10.67.

Because the pattern is produced by adding two pure sine waves, it is known as the *two tone test* signal. As it is merely a double sideband suppressed carrier signal in another guise, it may be generated in a phasing exciter by disabling one of the two balanced modulators and applying a single audio tone to the input of the speech amplifier. In the filter exciter, the same result may be produced by applying single tone

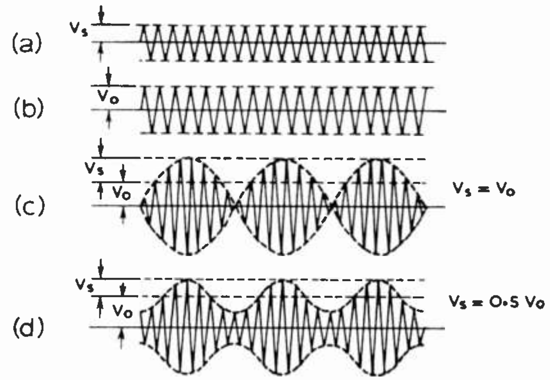


Fig. 10.67. Typical heterodyne waves, showing how the combining of two waves of slightly different frequencies results in a wave which pulsates in amplitude at the difference frequency of the component waves, and how the wave shape of the envelope of the resultant wave depends upon the relative amplitudes of the two components. (c) Shows the familiar half-sine-wave two-tone test pattern.

modulation and reinserting carrier of exactly the same amplitude as the s.s.b. signal passed through the filter. It is not necessary to go to the trouble of building an audio oscillator to give output at two separate frequencies.

To the sideband operator the two-tone envelope is of special importance because it is from the envelope that the power output from an s.s.b. system is usually determined. An s.s.b. transmitter is rated in p.e.p. output with the power measured with a two equal-tone test signal. With such a signal the actual watts dissipated in the load are one-half the p.e.p.

The generation of this two-tone envelope can be shown clearly with vectors representing the two audio frequencies as shown in Fig. 10.68. When the two vectors are opposite in phase the envelope voltage is zero. When the two vectors are exactly in phase, the envelope value is maximum. This generates the half sine-wave shape of the two-tone s.s.b. envelope which has a repetition frequency equal to the difference between the two audio tones.

When the half sine-wave signal is fed into a load, an r.m.s. calibrated cathode ray oscilloscope across the load indicates the r.m.s. value of the peak envelope voltage. This c.r.o. deflection is equal to the in-phase sum of $V1 + V2$, where $V1$ and $V2$ are the r.m.s. voltages of the two tones. Since $V1 = V2$, the p.e.p. $= (2V1)^2/R$ or $(2V2)^2/R$. The mean power dissipated in the load must equal the sum of the power represented by each tone, $V1^2/R + V2^2/R = 2V1^2/R$ or $2V2^2/R$. Therefore, with a two equal-tone s.s.b. test signal,

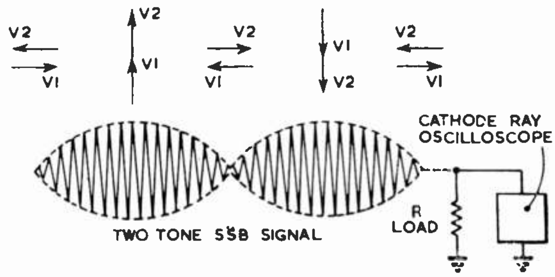


Fig. 10.68. Power measurement from two-tone s.s.b. test signal.

the mean power dissipated in the load is equal to 0·5 of the p.e.p., and the power in each tone is equal to 0·25 of the p.e.p. The peak envelope power can be determined from the relationship, p.e.p. $= V^2(c.r.o.)/R$. The mean power can be determined from the relationship, $P_{mean} = \frac{1}{2}V^2(c.r.o.)/R$. Similar measurements can be made using an r.f. ammeter in series with the load instead of the c.r.o. across the load. The analysis can be carried further to show that with a three equal-tone s.s.b. test signal, the power in each tone is one-ninth of the p.e.p., and the mean power dissipated in the load is one-third the p.e.p.; with a four equal-tone test signal the power in each tone is one-sixteenth of the p.e.p., and the mean power is one-quarter the p.e.p. and so on.

Practical Adjustment Procedure

When a linear amplifier has been built and given a preliminary check for accuracy of wiring, the following procedure is recommended:

(i) Apply reduced potentials to anode (and screen if applicable) and adjust the bias so that the valve is running at the maximum permissible anode current—this will be just within the maximum rated anode dissipation. Check that there is no trace of self-oscillation at the required operating frequencies, or of parasitic oscillation at v.h.f. Correct neutralization is infinitely more important in a class AB amplifier than in a class C stage used for telegraphy or A3 telephony. This does not apply to a passive grid amplifier where the heavy grid damping makes neutralization unnecessary.

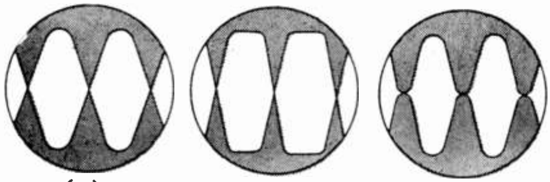
(ii) Increase applied potentials to the values recommended by the valve manufacturer and adjust the grid bias voltage so that the valve is taking the recommended zero signal anode current. If this is not known, adjust the bias so that the standing anode current is just within the rated maximum anode dissipation. Make sure that the neutralization is effective and the amplifier perfectly stable, no matter how the grid and anode tuning controls may be set.

(iii) Couple an oscilloscope to the output tank inductor by means of a link and coaxial cable. Connect a non-inductive resistive load of 75 or 80 ohms capable of dissipating the full mean power output to the output tank and fully mesh the loading capacitor: this corresponds to minimum loading.

(iv) Switch the exciter to the lowest band and with a 1 kc/s or 1·5 kc/s audio input from an audio signal generator into the microphone socket, adjust the audio gain and carrier injection controls so that the exciter drives the linear amplifier with two frequencies of equal amplitude (two-tone output). If there is no provision for carrier insertion, it will be necessary to use a two-tone oscillator—see Chapter 19.

(v) Adjust the exciter r.f. drive control, the amplifier anode tuning and loading together until an undistorted pattern of maximum amplitude is obtained at the maximum d.c. input permitted by the valve manufacturers' ratings for two-tone input conditions. If two-tone test data is not available, a value of 0·7 times the permitted d.c. maximum signal current for single-tone input may be employed. The correct tuning point for anode resonance is indicated by a sharp increase in screen current and a dip in anode current—this position of the anode tuning control should coincide with maximum r.f. output. If it does not, the maximum r.f. output occurs at some other position of the anode tuning control this indicates that the valve is not correctly neutralised and the n.c. requires slight readjustment.

(vi) If the amplifier is capable of delivering more than 400 watts p.e.p. output (200 watts mean) to the load, the face of the oscilloscope should be calibrated at the 200 watts mean r.f. output level by inserting an r.f. ammeter in series with the resistive dummy load as previously described in the section " Output Power of a S.S.B. Transmitter using a Two Tone Test Input".



(a) (b) (c)

Fig. 10.69. Oscilloscope patterns of the output of a linear amplifier with two-tone test input; (a) amplifier correctly adjusted; (b) peaks flattened because of insufficient anode loading or overdrive; (c) distortion at cross-over points because of incorrect bias voltage.

(vii) Examine the " cross-overs " on the oscilloscope critically. They should be sharp as in **Fig. 10.69(a)**. If, however, they appear compressed as in **Fig. 10.69(c)** the bias should be adjusted until they are correct. With suitable valves this type of distortion should not give trouble, but if it persists the bias supply should be suspected. A flattening of the envelope peaks as in **Fig. 10.69(b)** indicates insufficient loading (loading is increased by *reducing* the capacity of the output pi-network capacitor) or overdriving the amplifier beyond its maximum signal capabilities. This condition produces severe splatter and distortion on the transmitted signal and must be avoided.

Whenever possible the manufacturers' figures for anode and screen current for maximum signal two-tone input conditions should be used and the drive and loading adjusted to obtain these readings on the amplifier panel meters. (This procedure is described in detail under the heading, " R.F. Amplifier Adjustment and Loading.")

(viii) The exciter carrier injection control should be returned to its zero setting and the audio input from the audio signal generator removed. Insert the microphone and drive the linear amplifier under normal speech conditions to the maximum deflection on the oscilloscope previously observed under two-tone input conditions; while doing this the readings to which the anode and screen current meters kick should be carefully noted. Thereafter, the gain of the exciter should be controlled so that the recorded values are not exceeded, otherwise intermodulation distortion and splatter will result. As the tuning of a correctly loaded amplifier is rather flat, the anode current dip is not so pronounced as with a class C stage. The screen current on the other hand, will rise to an unmistakable peak as the anode circuit is tuned to resonance. It should be remembered that the anode and screen meters cannot follow the syllabic rate because of the inertia of the meter movement—an amplifier loaded to 200 mA under single-tone conditions will show 140 mA under two-tone input conditions and approximately 100 mA under speech conditions, *for the same peak envelope power output*. Never talk the amplifier up to a peak current swing (as seen on the anode current meter) that is more than half of the loaded current (single-tone) value.

Summary of Basic Formulae

Symbols

The following symbols will be used in all formulae given below.

VOLTAGE

V_a .. .. .. D.C. anode voltage

$V_{a\ min}$ .. .. Instantaneous anode voltage at the peak of the r.f. grid driving cycle.

V_{g2} .. .. .. D.C. screen voltage.

V_{g1} .. .. .. D.C. grid voltage.

CURRENT

I_a .. .. .. D.C. anode current (as read on anode current meter).

$I_{a\ peak}$ .. .. .. Instantaneous anode current at the peak of the r.f. grid driving cycle.

$I_{a\ zero\ sig}$ .. .. Static anode current (resting anode current under zero drive condition).

I_{g2} .. .. .. D.C. screen current.

I_{g1} .. .. .. D.C. grid current.

RESISTANCE

R_L .. .. .. External anode load resistance.

R_{out} .. .. External aerial load resistance (normally 75 ohms).

POWER

P.e.p. input .. .. Peak Envelope Power input. This is the r.m.s. power input at the crest of the modulation cycle.

P.e.p. output .. Peak Envelope Power output. This is the r.m.s. r.f. power output at the crest of the modulation cycle.

$P_{out\ mean}$ .. .. Average, or mean r.f. power output under modulating conditions.

P_a .. .. .. Anode dissipation.

P_{g2} .. .. .. Screen dissipation.

P_{g1} .. .. .. Grid dissipation.

MISCELLANEOUS

K .. .. .. Constant whose value depends on the angle of anode current flow. (For 180° K = 3·14; for 200° K = 2·88). A useful figure for quick calculation is a value of K = 3·0.

Eff .. .. .. Anode efficiency.

Basic Formulae

P.e.p. input $= V_a \times I_a$

$I_{a\,peak} = I_a \times K$

P.e.p. output $= I_{a\,peak} \dfrac{(V_a - V_{a\,min})}{4}$

$R_L = 2\dfrac{(V_a - V_{a\,min})}{I_{a\,peak}}$

Single-tone Input Test Signal

D.c. anode current $I_a = \dfrac{I_{a\,peak}}{K}$

Anode Input Watts and p.e.p. $= \dfrac{I_{a\,peak} \times V_a}{K}$

Average Output Watts and p.e.p. $= I_{a\,peak} \dfrac{(V_a - V_{a\,min})}{4}$

Anode Efficiency (Eff) % $= \dfrac{p.e.p.\ output}{p.e.p.\ input} \times 100$

$= K \dfrac{(V_a - V_{a\,min})}{4 \times V_a}$

Two-tone Input Test Signal

D.c. anode current $I_a = \dfrac{2 \times I_{a\,peak}}{K^2}$

Anode Input Watts $= \dfrac{2 \times I_{a\,peak} \times V_a}{K^2} = I_a \times V_a$

Average Output Watts ($P_{out\,mean}$) $= I_{a\,peak} \dfrac{(V_a - V_{a\,min})}{8}$

Peak Output Watts (p.e.p. output) $= I_{a\,peak} \dfrac{(V_a - V_{a\,min})}{4}$

Anode Efficiency (Eff) % $= \left(\dfrac{K}{4}\right)^2 \times \dfrac{(V_a - V_{a\,min})}{V_a}$

$= \dfrac{P_{out\,mean}}{I_a \times V_a} \times 100$

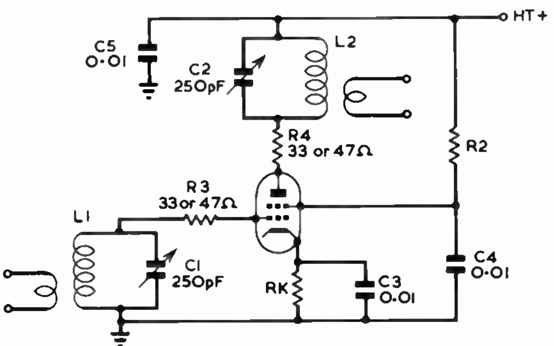


Fig. 10.70. Typical class A r.f. linear amplifier. For operating conditions, see Table 10.7.

LINEAR AMPLIFIERS

There are many different linear amplifier circuits, all capable of giving satisfactory service. The final choice is purely a personal one and will depend on the individual requirements in regard to such things as maximum power output, availability of valves and components and preference in regard to the use of high voltage (2000 to 3000) or low voltage (700 to 1250) from the main power supply. The circuits in this section have been chosen in such a way as to cover the broad field from 100 watts p.e.p. output up to the UK licensed maximum of 400 watts p.e.p. output, and at the same time to show as many different basic methods of operation as possible.

The Class A Linear Amplifier

The class A linear, of which a representative diagram is given in **Fig. 10.70**, will be seen to bear a strong family resemblance to the r.f. and i.f. amplifiers in receivers. Care must, however, be taken not to presume too far upon this relationship, otherwise trouble in the form of unexpected distortion may result. A true class A stage must be so biased and driven that the valve operates at all times on the linear part of its characteristic curve. The r.f./i.f. amplifier is essentially a small-signal amplifier; that is, the signal applied to its grid is only a fraction of a volt, so the valve will operate linearly no matter how it is biased. Furthermore, the stage is not required to produce any power and works into the constant load provided by the grid of the following stage. In s.s.b. service few class A stages work under such favourable conditions. They are usually called upon to produce a reasonably large voltage swing, and perhaps several watts of output power as well. If they are to drive a class AB2 stage, the loading will require some kind of stabilization to minimize fluctuations. There is, therefore, limited scope for small receiving valves in s.s.b. linears. It pays to follow the lead of the designer of high fidelity audio equipment who customarily builds an amplifier of far higher power handling capability than is really needed. By underrunning valves, the problem of distortion is avoided.

Cathode bias will necessarily result in negative voltage feedback which will reduce the power and amplification obtainable from the stage. If it is essential to obtain maximum performance, fixed negative bias should be applied to the grid.

The efficiency of a practical class A stage varies from zero under no-drive conditions to about 30 per cent at full input which means that the maximum output is limited to something less than one-third of the rated anode dissipation of the valve used. Provided that the internal insulation is built to stand it, operation at high voltage and low current will give greater efficiency and better linearity than low voltage and high current. It is impossible to list recommended operating conditions for all suitable valves under all circumstances, but **Table 10.7** gives a representative set of typical working conditions for the valves already mentioned. Suitable conditions for valves not included in the table may be deduced from the audio frequency power amplifier ratings given in manufacturers' characteristic sheets.

TABLE 10.7

Typical operating conditions of a representative selection of r.f. tetrodes and pentodes in class A linear amplifier service.

	6AU6	EF89	6CH6	6CL6	EL84	5763	6146	807
V_a	250 volts	250 volts	250 volts	300 volts	250 volts	300 volts	500 volts	500 volts
I_a	10 mA	9 mA	40 mA	30 mA	48 mA	40 mA	40 mA	50 mA
V_{g2}	150 volts	100 volts	250 volts	150 volts	250 volts	225 volts	150 volts	200 volts
I_{g2}	4·5 mA	3 mA	6 mA	7 mA	5·5 mA	2·4 mA	2 mA	1·6 mA
V_{g1}	−1 volt	−2 volts	−4·5 volts	−3 volts	−7·5 volts	−7·5 volts	−22 volts	−15 volts
R_k	68 ohms	180 ohms	100 ohms	82 ohms	135 ohms	175 ohms	470 ohms	280 ohms

The fixed bias voltage and the cathode resistor in the table above are alternatives.

The construction of a class A stage follows common practice; input and output circuits should be well shielded from one another to prevent self-oscillation, and high frequency parasitics should be eliminated by the generous use of stopper resistors. If oscillation at operating frequency should prove troublesome, the capacitance bridge neutralizing arrangement will provide a sure remedy.

Class AB Amplifiers—Triodes

Although it would be technically possible to achieve sufficient output for amateur purposes by using class A amplifiers exclusively, the result would certainly be uneconomical. Greater efficiency is essential in any practical transmitter, and this may be obtained by operating the high level stages in class AB or class B. For the purposes of this chapter, the precise shade of difference between a class AB amplifier and a true class B stage is unimportant and will be ignored.

When examining the ratings of power valves, it is important to note that the efficiency and typical operating conditions occasionally quoted for r.f. amplifier service in valve data sheets refer to the amplification of carrier type a.m. waveforms and are not directly applicable to s.s.b. use. The linear amplifier is a variable efficiency device: its efficiency in the absence of drive is zero, rising in a regular manner to between 50 and 75 per cent at maximum drive. The greatest efficiency obtainable depends on a number of interdependent variables, as well as upon the valve itself, but 66 per cent may be taken as rough generalization. The figures given in manufacturers' literature relate to d.c. measurements under conditions of no modulation; on modulation peaks, input and efficiency double, and in troughs both drop to zero. In s.s.b. service there is no carrier to provide a reference point against which to strike an average, so peak input and peak power handling capacity are the standards of measurement normally used. Conveniently enough, these standards are used by manufacturers themselves when rating valves for audio frequency linear amplifier service. When available, these audio ratings provide a more convenient guide as to what may be expected in an r.f. linear than any other published data.

In theory, any r.f. valve may be used as a class AB or class B linear, but there are a number of external factors which narrow the field considerably. Except in the special case of the zero-bias triode, fixed negative bias is essential, and if the stage is driven into grid current the voltage must have excellent regulation so that it does not vary over any part of the input cycle. Dry cells are inexpensive and may be used if the amount of grid current flowing through them is strictly limited. At high values of reverse current any battery will quickly develop enough internal resistance to cause the instantaneous bias voltage to fluctuate with input. The life of the cells under such conditions would be uneconomically short. Battery bias is therefore restricted to tetrode and pentode stages.

Most important of all is the effect of the amplifier on the preceding drive stage. With any valve which draws grid current over only a part of the input cycle, the onset of grid current will cause the load presented to the driving stage to fall from something approaching infinity to a fairly low value. Wide fluctuations in loading would have a disastrous effect on the linearity of the driver so some way has to be found to minimize the variation. The most practical method is to load the driver with a resistor chosen to dissipate at least ten times the peak load presented by the driven stage. In this way the load on the driver increases only 10 per cent at the onset of grid current, and distortion is kept within bounds. Resistive loading gives rise to no difficulty with tetrodes or pentodes, which require a watt or less of driving power, but rules out low-mu triodes, which would need far too large a driver stage to make them either practical or economical.

One possible way to use low-mu triodes is to bias them to cut-off and to adjust the drive so that they will not operate beyond the negative grid-voltage region; that is, in class AB1. The peak efficiency in this mode of operation is only 40–50 per cent, because the peak current which the valve can draw at zero grid voltage is relatively low. A high voltage bias supply has also to be provided. Zero bias triodes such as the 805, 811 and TZ40 are a much more favourable proposition. As grid current flows throughout the whole of the driving cycle, the load presented to the driver *during the time at which it is delivering power* is virtually constant.

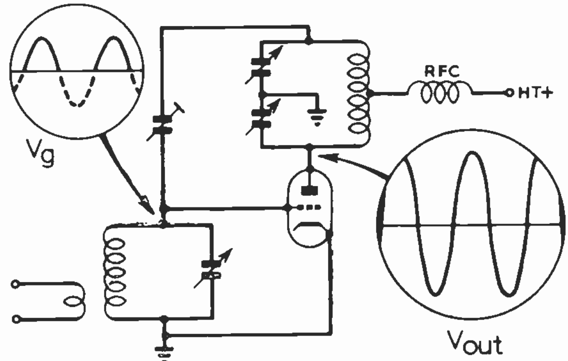


Fig. 10.71. Waveforms at grid and anode of a class B zero bias linear amplifier.

Fig. 10.71 shows a zero bias triode driven by a pure sine wave. Only that portion of the driving waveform which is depicted by the solid line can, however, result in r.f. output; when the driving signal enters the region represented by the dotted line the valve is cut off, and the positive excursion of the output waveform is provided by the flywheel effect inherent in the anode tank circuit. As the valve itself is cut off, it cannot produce intermodulation distortion, and it isolates its output circuit from any distortion which may occur in the driver stage because of the change in loading. Some distortion close to the cross-over point is, of course, inevitable, but practice has proved that this has negligible effect on the signal. It is the distortion at maximum input which causes trouble.

Triode valves have the disadvantage in r.f. amplifier service of requiring neutralization which is more difficult to accomplish in a class AB amplifier than it is in a class C amplifier because of the higher power gain. In order to ensure that neutralization is effective and the amplifier stable on all bands, both the theoretical circuit and the physical layout must be symmetrical. For this reason the valves should preferably be operated in push pull as shown in **Fig. 10.72** with a balanced tank circuit instead of the more usual parallel operation with a pi output network.

The 805 was originally designed as a high quality class B audio amplifier, and it is difficult to find a better valve for linear service at peak outputs up to 150 watts. A pair in push-pull (or parallel) will deliver 300 watts and require only 6 watts of driving power, which can be supplied fairly comfortably by a single 807 in class A. The circuit diagram is self-explanatory, but the factors governing the design of the various tank circuits will be examined in detail so that those wishing to adapt the design for use with different valve types may be able to do so without difficulty. The makers' recommended operating conditions for the two in push-pull are given in **Table 10.8**, from which will be seen that the optimum load for highest undistorted output is 6700 ohms. An output circuit Q of about 15 is desirable which necessitates that the reactance of both the tank capacitor and the inductor at resonance shall be $\frac{6700}{15}$ or 430 ohms approximately (see page 10.36). The required tank capacitance may easily be calculated from the formula $X_C = \frac{1}{2\pi fC}$. This is, however, the total value of C_3 and C_4 in series, so the value of each half of the split stator capacitor requires to be double that obtained by calculations. Suitable values for all amateur bands from 3·5 to 28 Mc/s are given in **Table 10.9**, together with coil winding instructions based on the Eddystone $2\frac{1}{2}$ in. diameter ceramic formers.

TABLE 10.8

Recommended operating conditions for two type 805 triodes in class B push-pull zero bias.

D.c. anode volts	1250 volts
Grid excitation—peak-to-peak	235 volts
Anode current—zero signal	148 mA
Anode current—maximum signal	400 mA
Grid driving power—maximum signal	6 watts
Anode input—maximum signal	500 watts
Power output—maximum signal	300 watts
Anode-to-anode load	6700 ohms

TABLE 10.9

Anode and grid circuit values for push-pull zero-bias class B amplifier using type 805 triodes.

Band	Anode		Grid	
	C	L	C	L
3·5 Mc/s	200 + 200pF	20 turns CT	300 + 300pF	32 turns CT
7 Mc/s	100 + 100pF	14 turns CT	150 + 150pF	20 turns CT
14 Mc/s	55 + 55pF	8 turns CT	75 + 75pF	12 turns CT
21 Mc/s	35 + 35pF	8 turns CT double spaced	50 + 50pF	12 turns CT double spaced
28 Mc/s	25 + 25pF	6 turns CT double spaced	40 + 40pF	10 turns CT double spaced

Anode inductors are wound on Eddystone or similar $2\frac{1}{2}$ in. diam. ceramic formers. The 21 and 28 Mc/s coils should be double spaced, and because stray inductance has an appreciable effect at these frequencies, pruning may be necessary in individual cases.
Grid inductors are wound on $\frac{3}{4}$ in. or 1 in. diameter formers, threaded 20 or 21 turns per inch.
Capacitor settings are " in use " values and include valve capacitances and circuit strays.

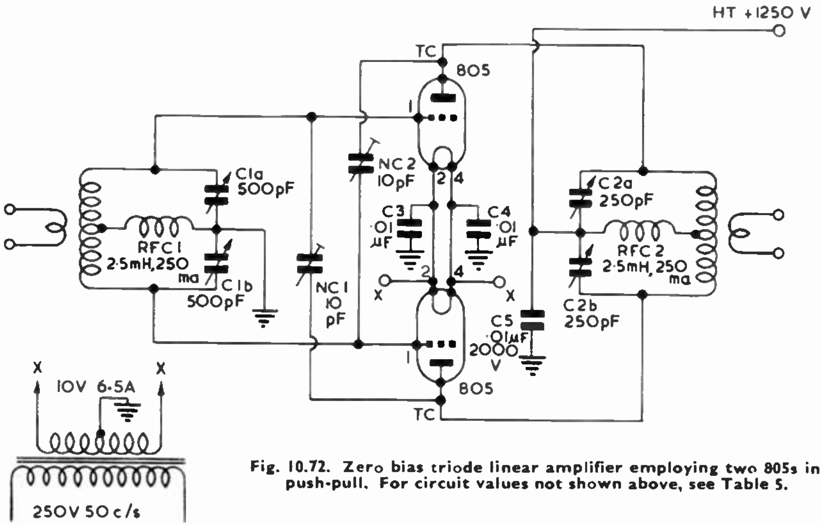


Fig. 10.72. Zero bias triode linear amplifier employing two 805s in push-pull. For circuit values not shown above, see Table 5.

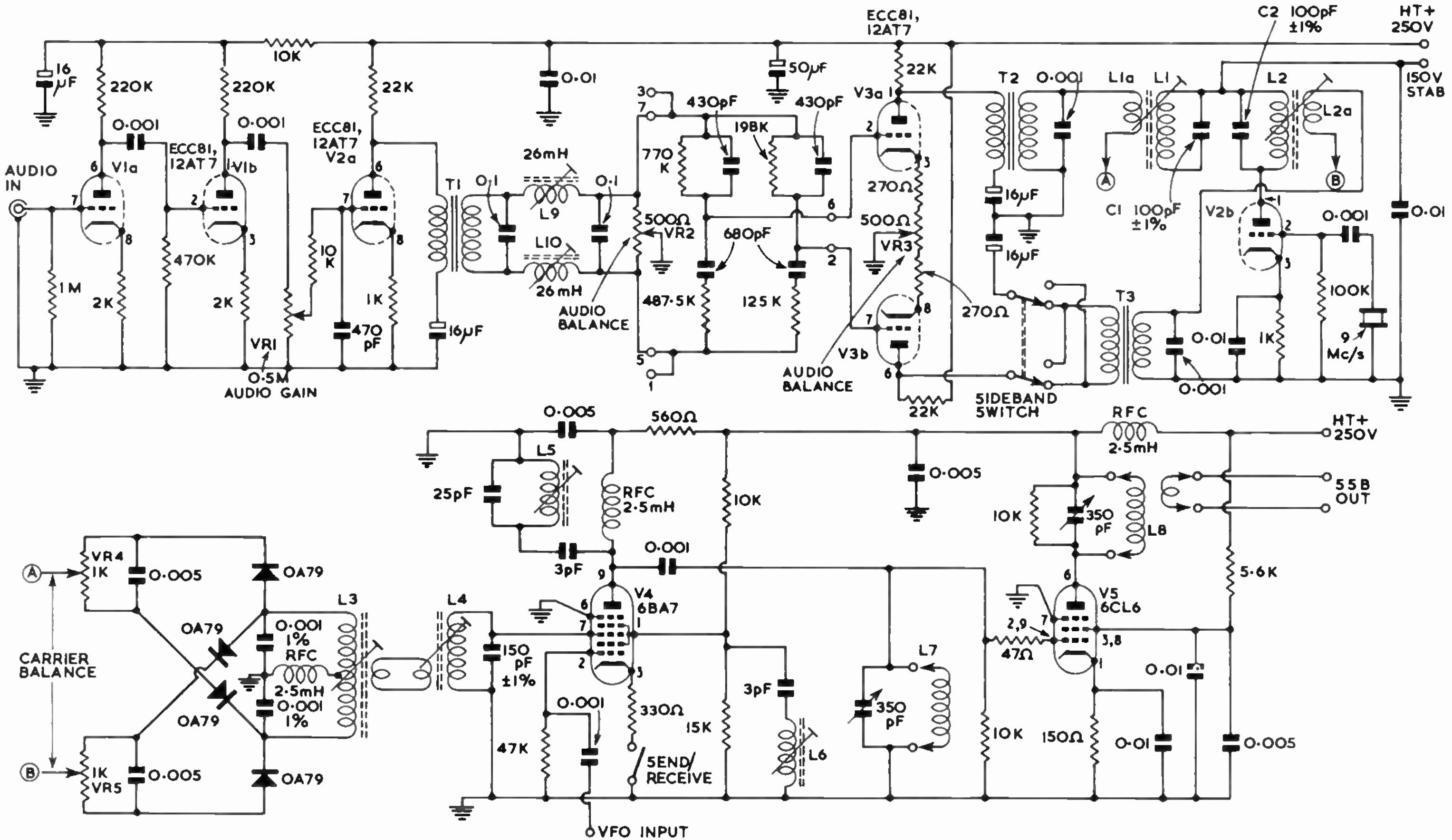


Fig. 10.73. Circuit diagram of a 9 Mc/s phasing exciter. L1, L2, 25 turns 20 s.w.g. enam. Closewound on ⅜ in. diam. slug tuned former. L1a and L2b are 4 turn links wound over the "cold" ends. The spacing between L1 and L2 which should be approximately 1½ in. may be adjusted to give optimum phase shift. L3 8 turns centre tapped ⅜ in. diam. and ¾ in. long, with two turn link at centre. L4, 25 turns 20 s.w.g. enamel close wound on ⅜ in. diam. slug tuned former. L5, L6, Trap coils, 18 turns 20 s.w.g. enamel, closewound on ⅜ in. diam. slug tuned former. L7, L8. See winding data in Table 10.10. T1 may be any step-down audio transformer with a ratio not less than 1:5. Standard loudspeaker transformers are suitable for T2 and T3. For details of the low pass audio filter see text. The commercial B & W type 2Q4 audio phase-shift network may be substituted for the components shown in the diagram.

If a different type of valve is used, it is advisable to obtain the manufacturer's recommendation about anode loading, and to calculate capacitor and inductor values in the manner demonstrated. Loading information is readily available for most of the valves likely to be encountered, but if it should prove difficult to locate for any particular valve, it can be calculated by the methods already discussed under the heading "Theory of Linear Amplification" on page 10.28.

The calculation of circuit values for the grid tank may be performed in exactly the same manner as for the anode circuit, and the results will be as shown in Table 10.9. Some manufacturers do not quote the grid load impedance in direct form, but for zero-bias valves this may be calculated from the stated grid driving-power and peak grid-to-grid input voltage by means of the simple formula

$$R_{grid} \simeq \frac{(\textit{Peak Input Voltage})^2}{2 \times \textit{Driving Power}}$$

The relatively large values of capacitance for the grid circuit are neither the result of miscalculation nor misprint; they arise from the characteristics of the valve under consideration. A two-gang receiving type variable capacitor will probably be found more convenient than a standard transmitting component, for under no circumstances should the tabulated values of capacitance be reduced, because the proper operation of linear amplifiers in general is vitally linked with the choice of correct component values in the tuned circuits.

Class AB Amplifiers Employing Tetrodes

The tetrode or pentode linear amplifier has become deservedly popular because it is capable of a much higher power gain than comparable triode valves. In the passive grid circuit arrangement it does not require neutralization and construction is thereby simplified.

A 9 MC/S PHASING EXCITER

The circuit shown in **Fig. 10.73** is a development of the S.S.B. Jr, designed by D. E. Norgaard, W2KUJ. Optimum performance is ensured in the modified circuit here presented by carrying out the r.f. phasing at a crystal controlled frequency of 9 Mc/s. An external v.f.o. at about 5·25 Mc/s enables output to be obtained in either the 3·5 or 14 Mc/s band. There is no particular disadvantage in generating the initial s.s.b. signal at a fixed frequency, because band changing must be effected by a heterodyning operation in any event. In common with all other forms of amplitude modulated telephony, band changing by frequency multiplication after modulation has been applied is out of the question. In the exciter shown in Fig. 10.73, the application of a heterodyning signal from the v.f.o. at 2, 12 or 20 Mc/s will result in output in the 7, 21 or 28 Mc/s bands respectively. The tuned circuits in the mixer and amplifier stages will, of course, have to be appropriately modified.

The audio amplifier consists of a 12AT7 (V1), with the two halves of the valve in cascade feeding into the first half of V2. Low frequency response is reduced by appropriately chosen coupling capacitors, and the low-pass audio filter which is interposed between T1 and the phase-shift network gives a high frequency cut-off which will be appreciated by other operators because of the resulting reduction in splatter. The inductors L9 and L10 may be adapted from standard television width-control coils, which should be adjusted to resonate at about 3200 c/s with a 0·1 μF capacitor in parallel. The minimum inductance varies from one manufacturer to another and with some types of coil it may be necessary to remove turns before resonance at a high enough frequency becomes attainable. The audio phase shift network is identical to that already described in Fig. 10.23. The input potentiometer VR2 should be a carbon-track component (not wirewound) and should be set initially with the aid of an ohmmeter so that the ratio of resistance either side of the slider is close to 2 : 7. If the component measures exactly 500 ohms, appropriate meter readings will be 110 and 390 ohms. When this potentiometer has been set it should be sealed and forgotten, all subsequent audio balance adjustments being made by VR3. Standard transformers designed to feed a 15 ohm speaker are ideal for use at T2 and T3.

Phase shifting at r.f. is accomplished by the coupled circuits L1 C1 and L2 C2 in the anode circuit of the crystal oscillator valve V2b.

The s.s.b. output at 9 Mc/s is link coupled from the tank circuit of the double balanced modulator to the signal input grid of the 6BA7 pentagrid mixer V4. The mixer is conventional except for the trap circuits in the screen and anode circuits which attenuate the third harmonic of the 5 Mc/s v.f.o. so that it will not cause a spurious radiation when operation in the 14 Mc/s band is required. If a grid dip oscillator is available, it will be found useful in setting the traps to frequency.

Coil winding details for all bands from 3·5 to 29·7 Mc/s are given in **Table 10.10**. If desired a band switching coil pack may be substituted for the individual coils indicated in the circuit diagram. The table also includes winding data for the tuned circuits in the output stage. This stage, which uses a 6CL6 miniature high gain pentode, follows normal practice, and is capable of providing enough excitation for a class AB1 linear amplifier.

TABLE 10.10

Coil winding data for Class A amplifier stage of phasing exciter shown in Fig. 10.73.

Band	Grid			Anode		
	Turns	Diam.	Length	Turns	Diam.	Length
3·5 Mc/s	60 *	3/8 in.	1 in.	30	1 in.	1 1/2 in.
7 Mc/s	30	3/8 in.	1 in.	20	1 in.	1 1/4 in.
14 Mc/s	16	3/8 in.	1 in.	12	1 in.	1 1/2 in.
21 Mc/s	9	3/8 in.	1 in.	8	1 in.	1 in.
28 Mc/s	7	3/8 in.	1 in.	6	1 in.	1 in.

* The 3·5 Mc/s grid inductor should be shunted by a fixed silvered-mica capacitor of 50-75 pF.

All coils are wound with 28 s.w.g. enamelled copper wire.

Alignment

Alignment is straightforward, but it does call for a certain amount of reliable test equipment. An oscilloscope is essential but it need not be expensive or complicated. A simple instrument will be found adequate. A valve voltmeter capable of making r.f. and a.f. measurements is also highly desirable, and like the oscilloscope, an inexpensive home constructed instrument will be just as effective as a highly priced commercial one. Finally, a simple audio oscillator capable of generating a good sine wave at about 1000 c/s will be found extremely useful. Suitable instruments are described in Chapter 19 (*Measurements*.)

Initially, adjust both tuned circuits in the anode of the 9 Mc/s oscillator, V2b, for maximum r.f. output; this may be checked with the valve voltmeter probe connected to each link winding L1a and L2a, in turn. Set the dust cores of L3 and L4 for maximum 9 Mc/s output at the anode of V4. A 4–5 volts signal from the v.f.o. should be satisfactory. With the v.f.o. set to 5·25 Mc/s, plug in the 3·5 Mc/s band coils and tune L7 and L8 for maximum r.f. This may be monitored conveniently by connecting the oscilloscope to the output link winding on L8. Adjust the oscilloscope gain to give a display of about 2 in. in height, and set the time base speed to a sweep rate of approximately 200 c/s.

At this point, no sideband signal will be going through the exciter and the oscilloscope deflection will be caused by the 9 Mc/s oscillator drive to the mixer beating with the 5·25 Mc/s v.f.o. input to produce a continuous r.f. output at 3·75 Mc/s. On the oscilloscope this will look like a c.w. signal—that is a smooth band of light. The two carrier balance controls are then adjusted to reduce this band of light to the smallest possible height (a thin horizontal line). The carrier is now balanced and there is no r.f. output from the exciter.

An audio signal of a fraction of a volt at 1000 c/s is next applied to the microphone input socket, care being taken not to overload the first two amplifier stages. This means that the volume control VR1 is turned fully to maximum gain and adjustment of audio level is undertaken at the audio oscillator itself.

The exciter is now producing a double sideband signal, and on the oscilloscope this will look like the pattern shown in **Fig. 10.74(a)**. The oscilloscope method of adjusting a phasing transmitter is based on the principle that a pure audio tone driving a perfect s.s.b. exciter, will produce only one r.f. signal—the desired sideband. On the oscilloscope this will look like a c.w. signal; that is, a rectangle of light with perfectly smooth edges. If there are any other signals present, unwanted sideband or carrier, they will beat with the desired signal and produce ripples on the oscilloscope pattern. The object, then, is to adjust the exciter until the pattern is a smooth band of light as shown in **Fig. 10.74(c)**. Note that the ripples due to the carrier are twice as wide (half the frequency) as those due to the unwanted sideband. By careful observation of the pattern, it is possible to tell what kind of adjustment is needed. Final suppression of the unwanted sideband is obtained by careful adjustment of the audio balance potentiometer VR2, followed by careful adjustment to the dust core of L1 and then L2. The setting of L1 will be critical, but that of L2 much flatter. It is important to move the cores only a few degrees at a time—this applies also to the setting of VR3 and the final adjustment will be critical to 1° of angular rotation of the slider. The core adjustment to L2 will affect the amplitude of r.f. output and unbalance the previously balanced modulator; this will require readjustment of the 1K ohms carrier balancing potentiometers, VR4 and VR5.

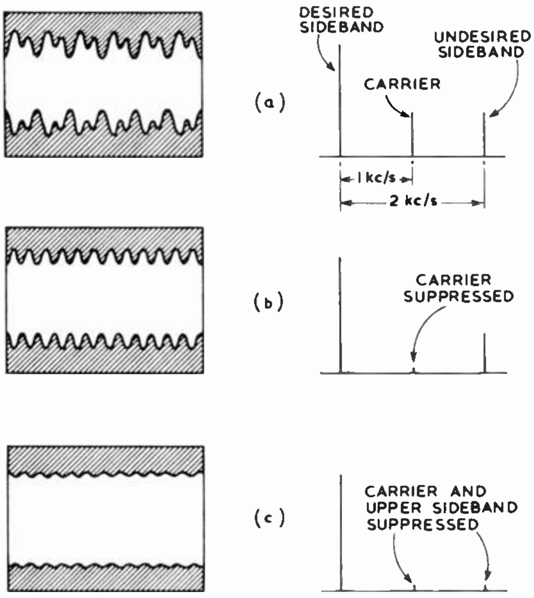


Fig. 10.74. Oscilloscope patterns for assessment of unwanted sideband and carrier suppression.

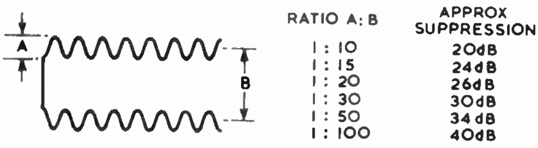


Fig. 10.75. Determination of degree of sideband suppression.

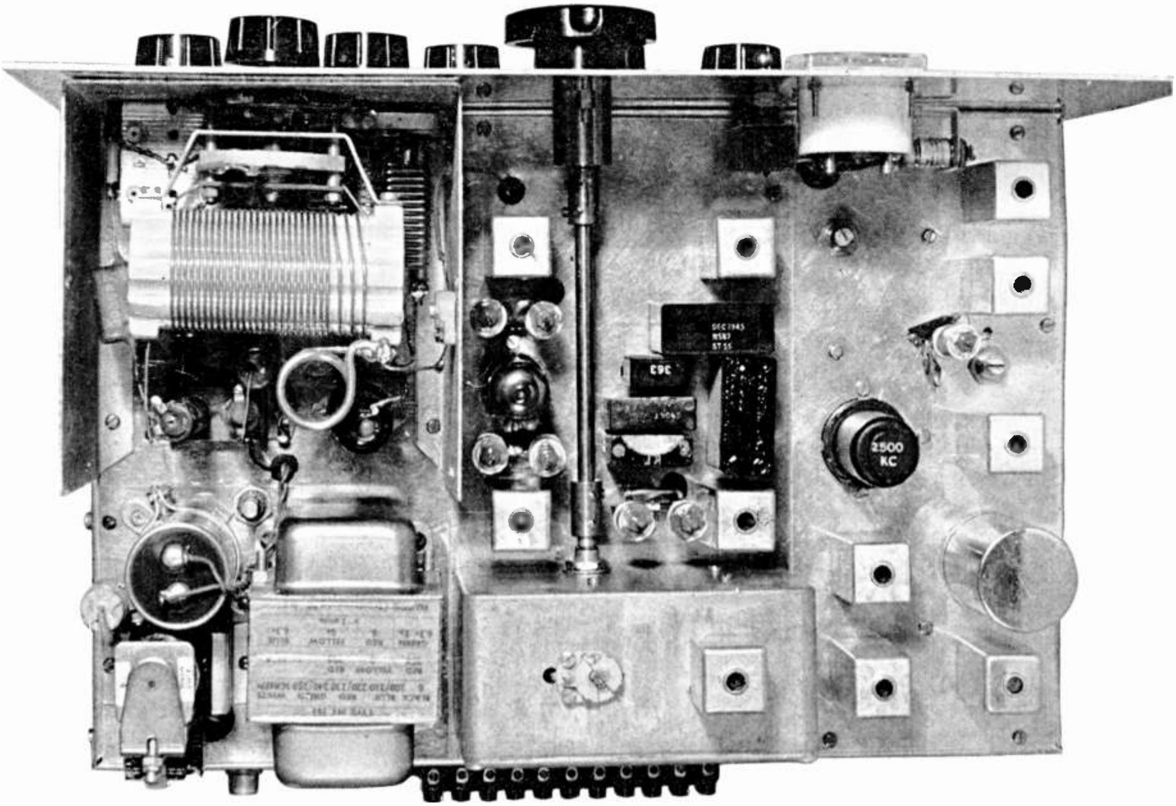
To obtain really good sideband suppression (in the range 35–40db) it is necessary to make alternate adjustments to the R.F. PHASING, AUDIO BALANCE and CARRIER BALANCE controls. This will take a considerable amount of time, but is well worth while if an above average s.s.b. signal is the aim. It is possible by observing the height of the final ripple in relation to the total trace height, to determine the sideband suppression that has been achieved. This is shown in detail in **Fig. 10.75**.

G3MVZ HYBRID S.S.B. EXCITER OR LOW POWER TRANSMITTER

The hybrid s.s.b. exciter, the circuit of which is shown in **Fig. 10.76**, employs transistors in all stages with the exception of the driver and power amplifier. Capable of an input of 25 watts p.e.p. it can be used as a low power transmitter on all bands from 3·5 to 29 Mc/s and has sufficient power output to drive a linear amplifier such as that shown on page 10.51.

Unit construction is used to permit ease of assembly and testing of each section. An unusual point of the design is the free-running carrier oscillator. A positive h.t. rail is used throughout to simplify the construction of the power supply which is an integral part of the design.

The audio section is quite conventional and employs two OC83 transistors although almost any audio transistor could be used with suitable adjustments to the circuit constants. The response is tailored to provide a reasonably flat response between 500 and 2500 c/s with sufficient gain for a crystal microphone. The audio level is set by means of the potentiometer VR1 between TR1 and TR2.



Top chassis view of the G3MVZ hybrid exciter/transmitter.

The free-running carrier oscillator (TR3, an OC170) is extremely stable and permits the carrier to be placed precisely on the skirt of the Kokusai mechanical filter for best audio quality. It is adjusted by means of the iron dust core in L1. The balanced modulator comprises a pair of matched OA79 germanium diodes. Critical adjustment of the potentiometer VR2 permits very effective suppression of the carrier. Some experiment may be needed to determine which side of VR2 provides the best suppression, circuit symmetry being the deciding factor as to the practical value of C11, a 3–30 pF Philips trimmer for the balance of stray circuit capacity.

The mechanical filter is a Kokusai type MF455K. The passband at the 6db points is 2·3 kc/s but the manufacturers' directions regarding interaction of the input and output circuits must be carefully followed. The shield provided must be fitted between the terminals of the filter, the coupling leads kept short and the stages before and after the filter correctly placed to permit this to be done.

The i.f. amplifier TR5 (OC170) is necessary to provide sufficient output from the mechanical filter for injection into the first mixer. The transformer T1 is shunted by TR5 to reduce the Q of the tuned circuit thus preventing a peaky response.

The first mixer (TR6) and the 2·5 Mc/s crystal oscillator (TR9) both use OC170 transistors, the oscillator being connected to the mixer via a low impedance cable. The correct sideband arising from the mixing of the output from the i.f. amplifier and the crystal oscillator injection is developed across T2 on a nominal frequency of 2045 kc/s. The second mixer (TR7) and the v.f.o. (TR12 and voltage amplifying stages TR10 and TR11) provide the 500 kc/s tuning range. The output of v.f.o. amplifying stages is connected to the emitter of TR7 by low impedance co-axial cable. The output coil (L8) of the v.f.o. unit is damped by a 10 K ohms resistor to provide a reasonably constant output across the tuning range of 2·955 to 3·455 Mc/s. This frequency, when mixed with the input to the mixer at 2045 kc/s, provides a tunable intermediate frequency of 5–5·5 Mc/s. This is selected by the wideband coupler T3 and fed to the third mixer, TR8. Both the primary and secondary windings of T3 are centre tapped. The coupler is stagger-tuned, the desired response being obtained by careful adjustment of the link coupling and the resonant frequency of each coil.

The third mixer provides correct sideband output in each band in conjunction with the band selection oscillator TR14 and harmonic amplifier TR13. The output coil, L6, which is common to all bands, thus simplifying the switching arrangements in relation to the low impedance link coupling, was designed for optimum performance at 23 Mc/s, loss of efficiency at the lower frequencies being compensated by the larger available output.

TABLE 10.11

C52, C53, C54	60 pF Philips trimmers.
C80	68 pF silver mica.
L1	465 kc/s i.f. transformer with secondary removed and primary stripped until on frequency.
L2	Repanco i.f. transformer XT27.
L3	50 turns, 30 s.w.g. close wound on Aladdin former, 470 pF padding capacitor. Secondary, 7 turns at earthy end.
L4	35 turns, 24 s.w.g., ¾ in. diameter former.
L5, L9	Aladdin former with slug. Pins 3 and 6, 3·5 Mc/s, 50 turns, 28 s.w.g., close wound. 60 pF trimmer + 220 pF fixed. Pins 4 and 6, 7 Mc/s, 25 turns, 28 s.w.g., close wound. 60 pF trimmer + 120 pF fixed. 14 Mc/s, 18 turns, 24 s.w.g., close wound. 30 pF trimmer + 25 pF fixed. 21 Mc/s, 14 turns, 20 s.w.g., close wound. 30 pF trimmer. 28 Mc/s, 12 turns, 20 s.w.g., close wound. 10 pF fixed.
L6	18 turns, 28 s.w.g. close wound, Aladdin former. Resonate L6 on 23 Mc/s with 10 pF silver mica, on 16 Mc/s with 60 pF Phillips trimmer, 9 Mc/s with 60 pF Philips trimmer and 68 pF s.m., and on 12·5 Mc/s with 60 pF Philips trimmer.
L7	25 turns, 28 s.w.g. Aladdin former, close wound, 65 pF padder.
L8	60 turns, 36 s.w.g. Aladdin former. 75 pF padder. Secondary, 8 turns, 36 s.w.g. at earthy end. For 3·5 and 7 Mc/s, L5 and L9 are wound on the same formers.
L10	Geloso pi-network coil and switch assembly or similar.
T1	Repanco i.f. transformer XT50.
T2	75 turns, 36 s.w.g., close wound. 220 pF padder. Coupling, 25 turns, 36 s.w.g., at centre.
T3	46 turns, primary and secondary centre tapped. 120 pF padders. Coupling link, 2 turns at cold end of each.
T4	250-0-250 volts 100 mA, 6·3 volts 2 A (twice).
X1	2·5 Mc/s.
X2	(for 20 and 80m). 9 Mc/s.
X3	(for 40m). 6·25 Mc/s.
X4	(for 15m), 8 Mc/s.
X5	(for 10m), 11·5 Mc/s.
X6	(for second 10m segment) , 11·75 Mc/s.

For ease of operation, broadband tuning is employed in the collector circuit of the third mixer and the anode circuit of the driver stage. This helps to maintain excellent stability in the stages concerned while still maintaining adequate drive on 10m. The driver (V1) and p.a. (V2) stages are conventional, employing an EF80 (6BX6) and EL81 respectively. H.t. for the p.a. is series fed through the p.a. tank coil L10. The bias on the grid of V2 is adjusted to provide class A operation at a standing anode current of approximately 25 mA.

The power supply uses two BY100 silicon diodes for full-wave rectification of the 250 volt anode supply to the final stages, and one BY100 for half-wave rectification of the 100 volt negative bias supply. The 12 volt supply for the transistor section of the exciter is obtained by a bridge circuit from the 6·3—0—6·3 volt secondary windings of the mains transformer using four GJ7M diodes.

The transmit-receive relay, which should have a coil of about 5 K ohms, is energized from the 250 volt h.t. line via a suitable series dropping resistor. It can be operated by a press-to-talk switch or a contact on a relay in the VOX unit if one is to be used.

Construction

Table 10.11 gives details of all coil and transformer windings.

The exciter is built in several sub-assemblies using miniature tag boards, each assembly being tested before being fitted into place under the 14 in. × 8½ in. × 2 in. chassis. The v.f.o. is built into an Eddystone type 845 diecast box measuring 2½ in. × 4½ in. × 2 in. and placed on the top of the chassis. Some care must be taken to line up the tuning capacitor with the tuning drive on the front panel since the use of flexible couplings does not give the necessary degree of positive tuning and direct drive with brass coupling rod is preferable.

The reduction gearing is the Eddystone 50 : 1 slow motion dial type 843, fitted with a smaller-than-standard dial. The tuning rate will be sufficiently positive for accurate netting.

The front panel measures 15 in. × 8 in. and carries the microphone input socket, the controls, and a small suitable meter to measure p.a. current.

The aluminium case is 15 in. wide, 8 in. high and 9 in. deep, which can be obtained together with front panel and chassis from suitable stockists. The small side ventilation slots are more than adequate and the unit will run with barely perceptible heat for long periods.

Additional 2 in. wide aluminium strips will be required for screening the stages under the chassis. All supply leads should be run close to the chassis edges and screened to prevent stray pick-up and leakage between sections.

GRID TUNED QVO6-20 OR 6146 AMPLIFIER

The amplifier shown in Fig. 10.77 is simple in operation, self-contained and reliable. It will handle a maximum power of 140 watts peak input on all bands from 3·5 to 28 Mc/s. Two QV06-20 (6146) valves are used in parallel, with a tuned grid circuit and modified pi-network output circuit. Input and output impedances are nominally 75 ohms.

The power supply is built on the same chassis as the amplifier, although the 250 volts 20 mA supply may be omitted if this is available from the exciter. The main h.t. voltage is 700 volts and the bias supply is −45 volts.

Grid Circuit

The input signal, between 2 and 4 watts peak, is fed through a co-axial socket on the front panel and thence to the link winding of the grid coil in use. The coil for each band has a damping resistor across the tuned winding to reduce the loaded Q and increase the margin of stability. **Details of the coils are given in Table 10.12.**

In the interests of good linearity, it is of great importance to ensure the stability of the amplifier under all conditions. Instability will result in degrading the outgoing signal and causing distortion products which will cause interference to adjacent signals.

TABLE 10.12
Inductors for Fig. 10.77.

L1, 31 turns 28 s.w.g. enam., close wound on Aladdin type F.804 former 13/64 in. diameter with iron dust core. (Nominal inductance 9 μH.)
L2, 15 turns 28 s.w.g. enam., close wound on Aladdin type F.804 former 13/64 in. diameter with iron dust core. (Nominal inductance 3·6 μH.)
L3, 14 turns 28 s.w.g. enam., winding length 0·7 in., on Aladdin type F.804 former with iron dust core. (Nominal inductance 1·35 μH.)
L1, L2, L3 have 2 turns 28 s.w.g. enam. link windings at the earthy ends.
L4, 10 turns 28 s.w.g. enam., winding length 0·5 in., on Aladdin type F.804 former with iron dust core. (Nominal inductance 1·03 μH.)
L5, 7 turns 28 s.w.g. enam., winding length 0·5 in., on Aladdin former type F.804 with no core. (Nominal inductance 0·8 μH.)
L4, L5 have 1 turn 28 s.w.g. enam. link winding at the earthy ends.
L6, L7, 7 turns 22 s.w.g. enam. wound on the ½ watt resistors R10 and R11 respectively.
L8, 28 turns 20 s.w.g. tinned copper, single spaced for 21 turns, double spaced for remaining 7 turns, on 1·75 in. ceramic former, total winding length 2·4 in., tapped at 16 turns (7 Mc/s), 7 turns (14 Mc/s), 4½ turns (21 Mc/s) and 2½ turns (28 Mc/s).
RFC1, 100 turns 36 s.w.g. enam., close wound on ⅛ in. diameter Tufnol former.
RFC2, 2·5 mH (Eddystone type 737).

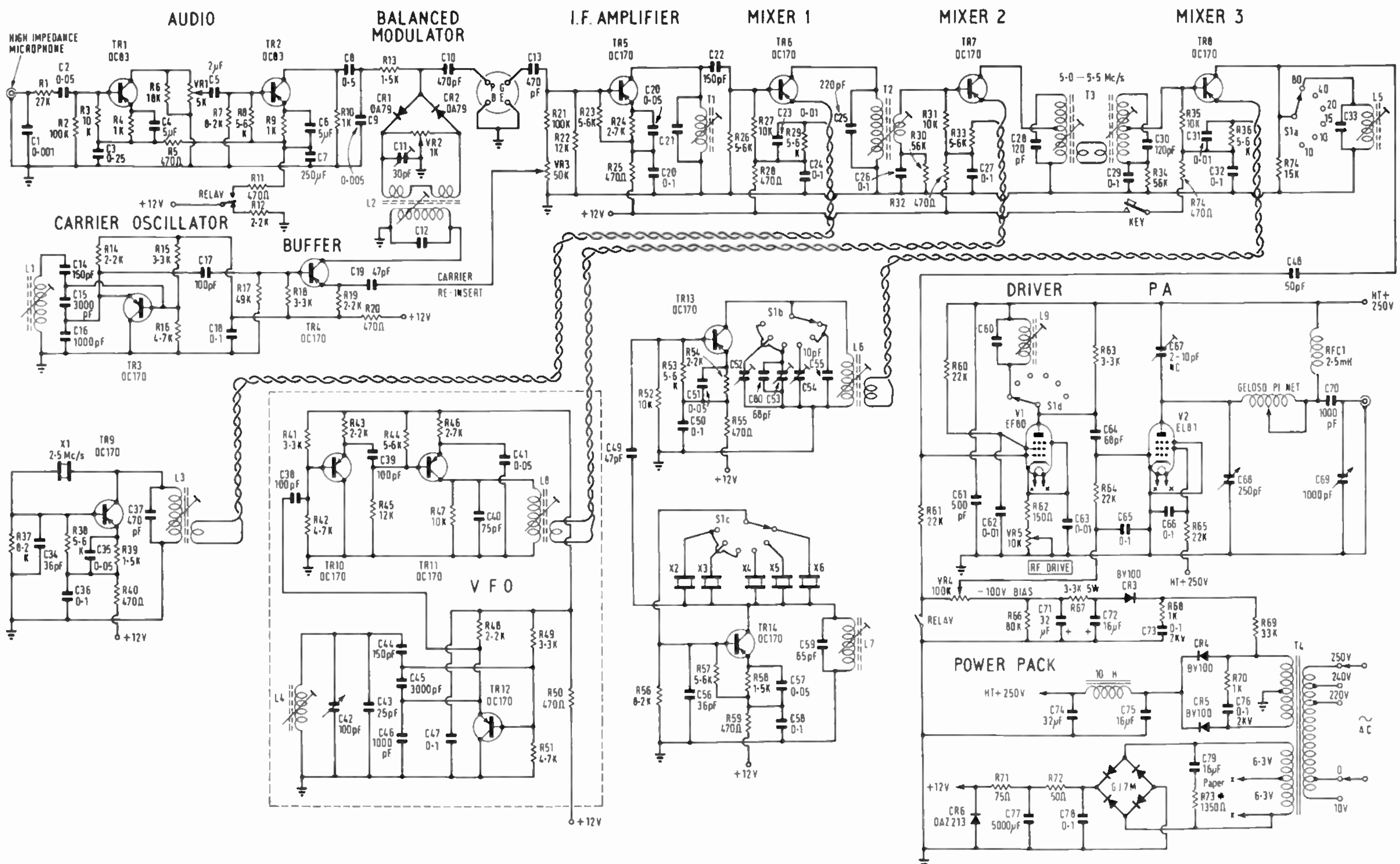


Fig. 10.76 Circuit diagram of the hybrid s.s.b. exciter or low power transmitter for 3.5 to 29 Mc/s. Unless otherwise stated all resistors are ¼ watt rating. Coil and crystal information is given in Table 10.11.

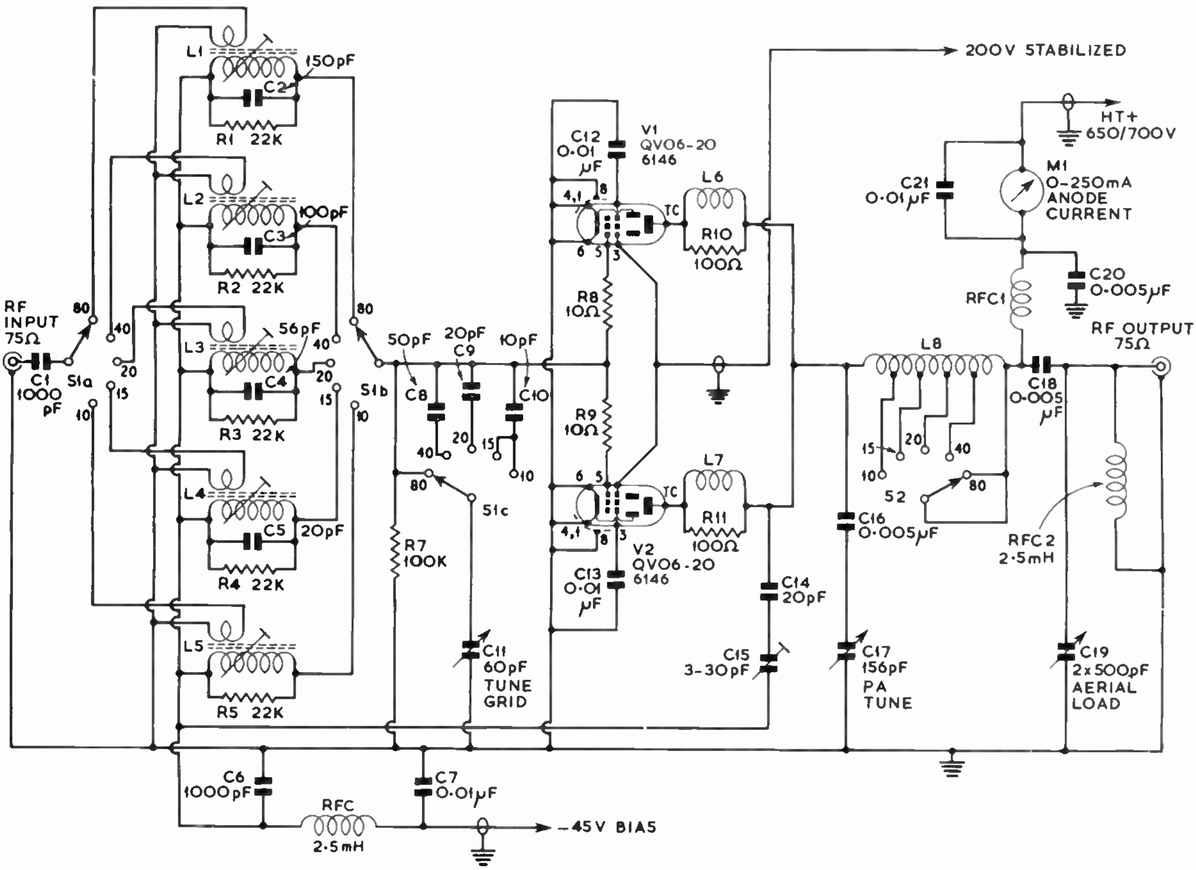


Fig. 10.77. Circuit diagram of grid-tuned bandswitched amplifier covering all bands from 3·5 Mc/s to 28 Mc/s.

Separate padding capacitors are switched into circuit on each band, reducing the electrical coverage of the grid tuning capacitor. This avoids the danger of tuning to an unwanted signal and also the need to calibrate this control. The bottom of the tuned winding on each coil is taken to a common bus-bar, with a 1000 pF capacitor from this point to earth. Bias voltage is fed to the valves through the grid coils.

Liberal decoupling is used around the valves to avoid the presence of r.f. on the supply leads. The screen grids are fed from a stabilized nominal 215 volts source, using two miniature stabilizers in series. However, this may be within range 195 volts to 215 volts depending on the stabilizers available.

Alignment of the grid circuits is best accomplished with a grid dip oscillator. Set C11 to near maximum capacity, adjust each core in turn and then check that the swing of C11 encompasses each band.

P. A. Tank Circuit

A modified pi-section anode tank circuit is used, calculated for an output impedance of 75 ohms. The anode supply passes through the p.a. coil, h.t. being fed in at the aerial end through an r.f. choke. This d.c. potential does not appear across either variable capacitor due to the series blocking capacitors but S2 carries the full d.c. potential. Using this method, the r.f. choke is no longer liable to damage, should it resonate within the bands in use. A further small r.f. choke is wired across the r.f. output socket to guard against a "hot" aerial, should the blocking capacitor fail.

Parasitic stoppers are incorporated right at the top cap connections of the valves, as well as on the grid pins below chassis. Neutralization is by means of the Philips trimmer *C*15 and series blocking capacitor C14. The trimmer is wired straight through the chassis on to the grid coil bus-bar. This trimmer and the 1000 pF capacitor to earth together with the $C_{a\text{-}g}$ and $C_{a\text{-}k}$ of the valves form a bridge neutralizing circuit which is, in theory, suitable for use over a wide range of frequencies without re-adjustment. C15 is adjusted with no drive to the amplifier and no output load. When the amplifier is driven into a load, maximum r.f. output should coincide with the dip in anode current, indicating correct neutralization.

The distortion products due to non-linearity are very small —in the region of −35db to −40db.

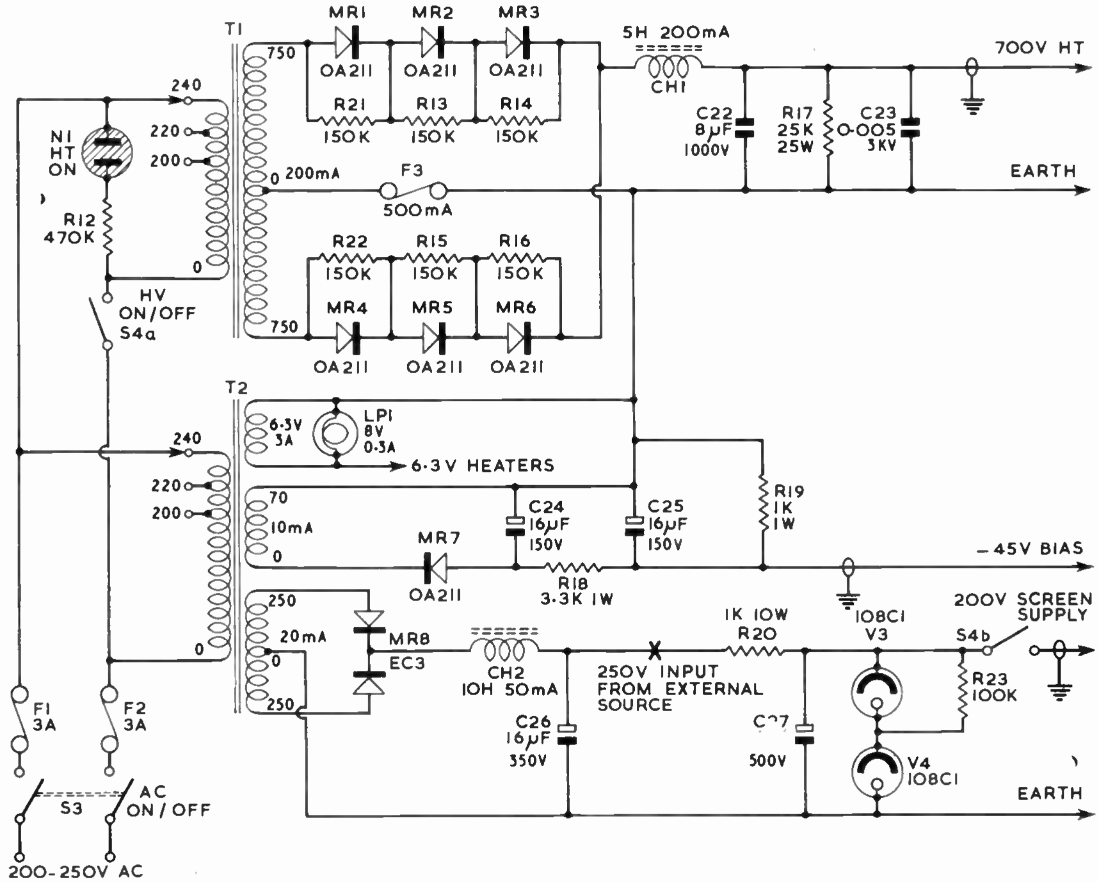


Fig. 10.78. Power supply unit for the bandswitching linear amplifier.

The power supply (**Fig. 10.78**) is conventional, although silicon rectifiers are employed to conserve space and because of their increased efficiency over valves performing a similar function. Two transformers are used, one supplying all low tension circuits, and the second, the high voltage for the p.a. anodes. This permits primary switching of the high voltage, by far the safest way of dealing with 700 volts.

Switching is in the primaries of the two transformers. However, the d.c. screen voltage is switched with the anode supply to avoid the screens drawing too high a current when the high voltage is switched off. Bias is applied at all times. Fuses are situated in the a.c. supply line and the centre tap of the high voltage secondary.

The silicon rectifiers suggested have a p.i.v. (peak inverse voltage) rating of 700 volts and a current rating of 500 mA. As the r.m.s. secondary voltage is a nominal 750-0-750, it is necessary to use six diodes, arranged in two series sections, to constitute a full-wave rectifier circuit. High stability resistors are wired in parallel with each rectifier to equalize the reverse current. Choke-input filtering is employed and, together with the low source impedance of the rectifiers, provides a supply which is fairly well regulated under static and dynamic conditions.

The screen, bias and heater supplies are obtained from the second transformer. A small full-wave contact-cooled rectifier, rated at 250 volts 20 mA, feeds two 108C1 (VR105/30) stabilizers, giving a screen voltage of 215 volts. This is slightly higher than is desired. The −45 volts bias is derived by half wave rectification from a 70 volts r.m.s. winding.

Layout

The general shape and size are not critical, depending largely on available components.

The layout of the r.f. section must be substantially similar to that described otherwise instability may result. All earth connections are taken to a common point near the valve-holders.

Ventilation of the equipment must be good as silicon rectifiers do not like a severe rise in operating temperature, unless mounted on heat sinks. Holes in the underside of the cabinet, which should be on rubber feet, permit the ingress of cool air by convection, the warm air passing out of louvres in the cabinet top.

A TT21 LINEAR AMPLIFIER

The circuit of a passive grid linear amplifier using a pair of TT21 valves in parallel is shown in **Fig. 10.79.** The drive requirements are lower than for the grounded grid configuration, less than 5 watts being required on all bands.

The measured power output at anode voltages of 800, 1000 and 1200 and a stabilized screen voltage of 300 is plotted in **Fig. 10.80.** The curves show that the output and efficiency fall as the frequency is increased above 10 Mc/s. This reduction in power output is almost entirely due to circuit losses, especially at 28 Mc/s where the valve and circuit minimum capacity is too high to permit the correct value of input capacity for the pi-network to be obtained. Nevertheless, the amplifier gives a very creditable performance as the results of tests given in **Table 10.14** show. Linearity tests were carried out at 7 Mc/s using a spectrum analyser and are summarized in **Table 10.13.** During these checks a little instability was encountered but was completely cured by the installation of the 100 ohm grid stoppers, R2 and R3, shown in Fig. 10.79. Parasitic chokes in the anode leads were not found necessary; if required, $2\frac{1}{2}$ turns of 16 s.w.g. enamelled copper wire on Erie type 8 100 ohm resistors should be satisfactory.

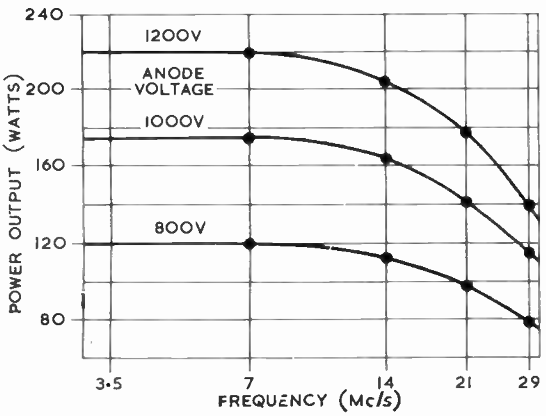


Fig. 10.80. Typical power output curves.

TABLE 10.13

Performance of TT21 Linear Amplifier
Two Tone Tests at 7 Mc/s

Anode Voltage V_a	1200	1000	800
Screen Voltage V_{g2}	300	300	300
Grid Voltage—V_{g1}	39	37	34
Peak Grid Drive $V_{g\ peak}$	39	37	34
Anode Current $I_{a(o)}$	62·5	75	94
Anode Current (2-tone) $I_{a(sig)}$	180	165	160
Screen Current I_{g2}	8	10	12
*Grid Current I_{g1}	0	0	0
Power Output (2-tone) $P_{out\ load}$	110	87·5	60
Power Output P.E.P.	220	175	120
†I.M.D. 3rd order—db	29	33	45
†I.M.D. 5th order—db	43	46	46
†I.M.D. of input signal—db	54	54	54

*Grid current limited to 0·1 mA. †Intermodulation distortion. Tests carried out using a Furzehill Spectrum Analyzer.

Construction

The construction of this amplifier is quite straightforward, and with the aid of **Fig. 10.80** should not present any problem. In general, the layout is quite conventional, but a good deal of care should be taken to avoid multiple earths. The earth returns of the anode pi-circuit, the valve cathodes and the screen bypass capacitors are connected to the same point on the chassis.

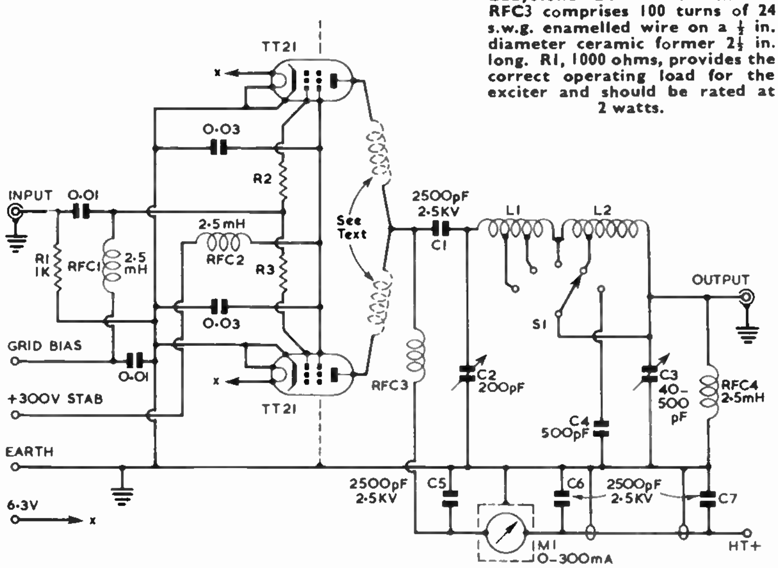


Fig. 10.79. Circuit diagram of the TT21 linear amplifier. RFC1, RFC2 and RFC4 are standard Eddystone 2·5mH r.f. chokes. RFC3 comprises 100 turns of 24 s.w.g. enamelled wire on a ½ in. diameter ceramic former 2¼ in. long. R1, 1000 ohms, provides the correct operating load for the exciter and should be rated at 2 watts.

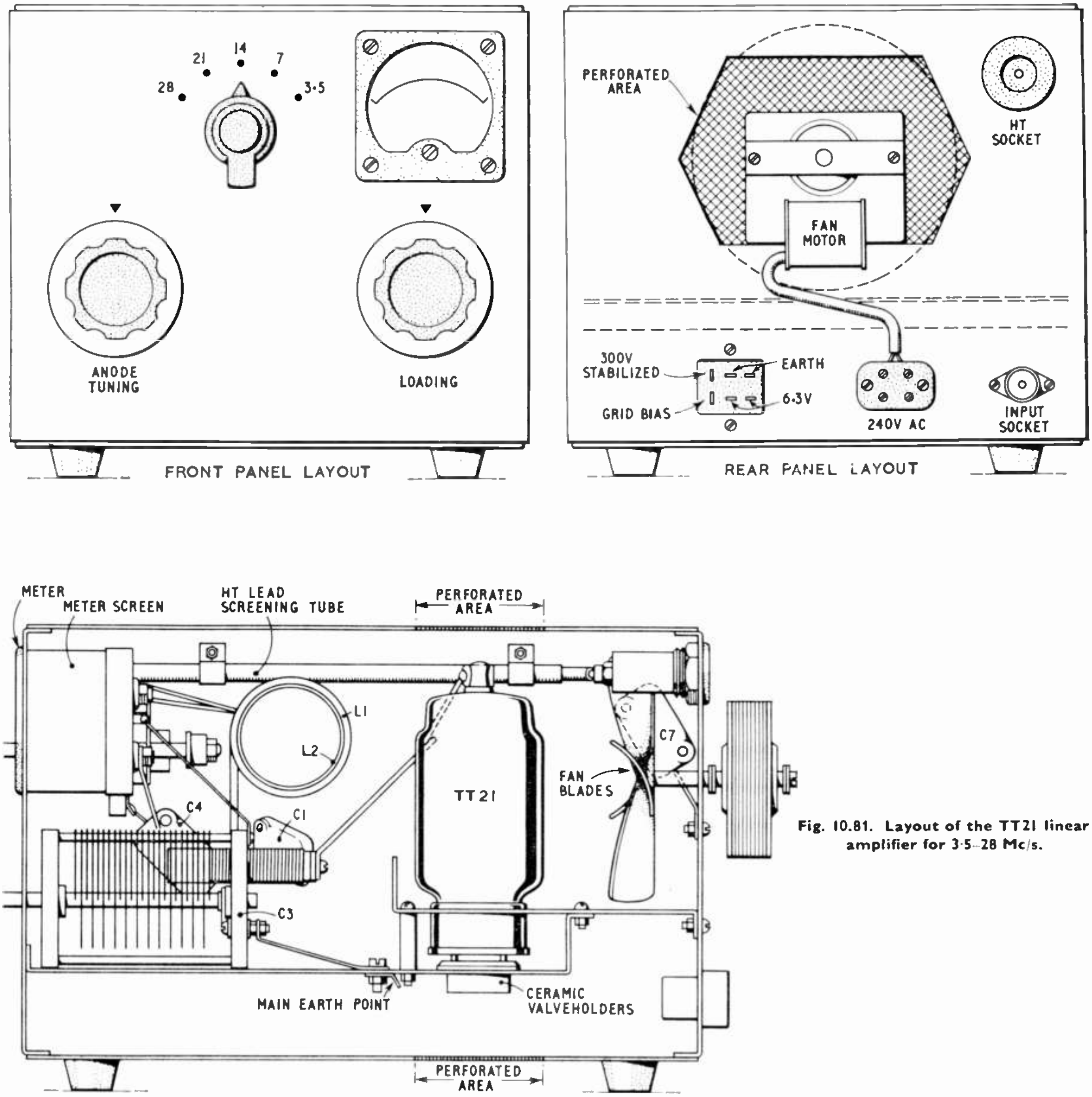


Fig. 10.81. Layout of the TT21 linear amplifier for 3·5–28 Mc/s.

The tuning capacitors C2 (200 pF) and C3 (500 pF) are the ceramic end plate type with earth returns made at only one point. In order to reduce the losses as much as possible, the anode pi-circuit returns and the feed connections are made of ⅜ in. 18 s.w.g. copper strip.

The band change switch is a large type and was chosen for its low capacity between contacts and its higher current-carrying capability. It is a standard five position ceramic wafer type with shorting plate and an extra contact to pick up C4.

The chassis is formed by two pieces which when fixed together allow the valveholders to be submounted on the lower member. The complete box should be made with the sides fixed to the chassis by means of aluminium angle, and attached to the front and back panels by their bent-over edges.

The heat generated by the two valves operating at their maximum rating will of course be more than 100 watts. This, together with the circuit heating in the relatively small cabinet, makes it necessary to have a fair degree of air

TABLE 10.14

Performance of TT21 Linear Amplifier (Single Tone Tests)

Band	3·5 and 7 Mc/s			14 Mc/s			21 Mc/s			29 Mc/s		
Anode voltage V_a	1200	1000	800	1200	1000	800	1200	1000	800	1200	100	800
Screen Voltage V_{g2}	300	300	300	300	300	300	300	300	300	300	300	300
†Grid Voltage $-V_{g1}$	39	37	34	39	37	34	39	37	34	39	37	34
*Anode Current $I_{a(o)}$	62·5	75	94	62·5	75	94	62·5	75	94	62·5	75	94
*Anode Current $I_{a(max)\,sig}$	270	260	210	250	230	190	250	240	210	235	215	210
*Screen Current $I_{g2(max\,sig)}$	18	18	19	14	16	16	15	16	16	14	16	16
§Grid Current $I_{g1(max)}$	0	0	0	0	0	0	0	0	0	0	0	0
Power Input P_{in}	324	260	168	300	230	152	300	240	168	282	215	168
‡Power Output Load P_{load}	220	175	120	204	167	114	170	140	98	130	115	80
P_{in}—P_{load} (per valve)	52	42·5	24	48	31·5	19	65	50	35	79	50	44

* For two valves. † Adjusted to obtain stated $I_{a(o)}$. ‡ Measured with Marconi Type T1205 500 watt r.f. power meter. § Grid current limited to 0·1 mA.

circulation, but a blower in the normal sense is, however, not needed. In order to get free air flow the base and cover plates are perforated in the region below and above the valves, and the lower chassis plate is drilled around each valve socket. This, together with the fairly large clearance holes through which the valves are fitted, allows a free air flow around the valves. In addition, a fan is fitted to the rear panel, which is also perforated, and draws air in through the panel and stirs the air within the cabinet, so that the temperature rise of the whole cabinet will be reasonable even after long periods of operation. The cabinet is mounted on rubber feet to ensure free air inlet through the bottom.

The fan is normally supplied with 6 in. diameter blades; these must be cut down to 4 in. diameter and reshaped to get satisfactory performance in the confined space. If space permits, the cabinet could be 2 in. higher, thus allowing the use of an unaltered fan. All the wiring of the fan motor is external to the cabinet and therefore should be free of r.f. pick-up.

The anode pi-circuit inductor is made in two parts: L1 covers the three higher frequency bands, 14, 21 and 28 Mc/s, and L2, in series with it, covers the two lower bands, 3·5 and 7 Mc/s. The construction is illustrated in **Fig. 10.82**.

L1 consists of nine turns of $\frac{1}{8}$ in. drawn copper, wound with an internal diameter of $1\frac{1}{2}$ in., and spaced $\frac{1}{16}$ in. between turns.

L2 consists of 31 turns of 16 s.w.g. on an epoxy resin former of $1\frac{1}{2}$ in. diameter; the turns are spaced to occupy $3\frac{1}{2}$ in. with spacing between turns of slightly less than $\frac{1}{16}$ in. The ends of the winding are fixed to the former by nuts and bolts. L1 is self-supporting and is mounted on the same axis as L2. The anode end of L2 is soldered to L1 at the point where the 14 Mc/s tap is made.

Tapping points for the various bands are as follows:

28 Mc/s—3 turns from anode end L1.
21 Mc/s—5 turns from anode end L1.
14 Mc/s—9 turns from anode end L1.

These connections are made of $\frac{1}{8}$ in. copper.

7 Mc/s—12 turns from anode end L2.
3·5 Mc/s—output end of L2.

The losses at the higher frequencies, especially on 28 Mc/s and to a lesser extent on 21 Mc/s, demand good soldering of these tapping points and should be carried out with a higher than average melting point solder. Ideally they should be "hard" soldered or brazed.

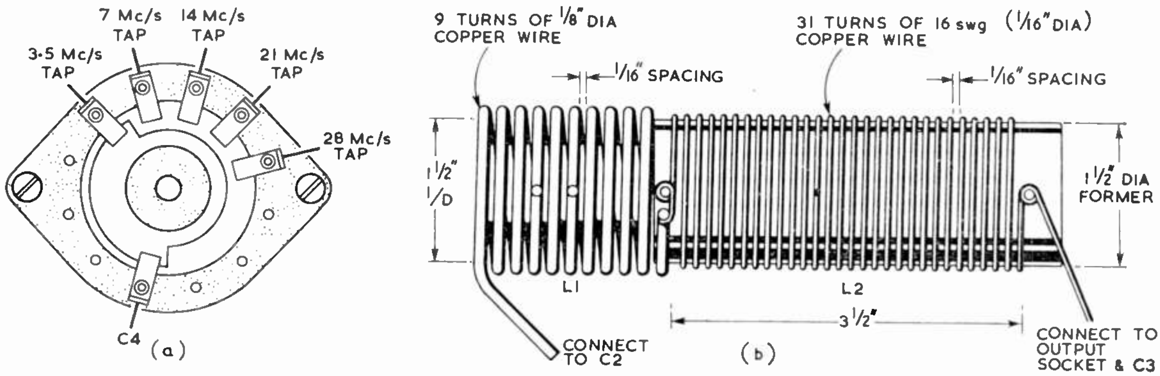


Fig. 10.82. (a) Details of the anode coil tapping switch. (b) Constructional details of the anode coil.

The r.f. choke which shunt feeds the circuit consists of a simple winding of 100 turns of 24 s.w.g. enamelled wire on a ceramic former ½ in. in diameter and 2½ in. long. It is mounted below, and at approximately right angles to, the main inductor. This choke has been found to be entirely satisfactory, and no overheating has been encountered on any of the five bands.

Drive to the amplifier is fed in through a standard coaxial socket.

In the prototype a type N connector is used for the output but any other of similar power rating will be suitable. It should be noted that the earth connection of the outer is connected directly to the main earth point with copper strip although the fixing screws make an earth connection to the cabinet at the socket.

Power Supply

As with all linear amplifiers it is important that the h.t. supplies should be adequately smoothed and of good dynamic performance in order to maintain the voltages substantially constant during the changing operating loads. This means low series impedance and high reservoir capacitance.

Probably the most satisfactory method of meeting these requirements is to use a fullwave bridge rectifier with either gas-filled valves or semiconductor diodes whose low forward impedance will keep down the series voltage drop.

To obtain good dynamic performance of the anode supply, a capacitance of not less than 24 μF will be required in the smoothing circuit. This can be readily obtained by use of TV type electrolytic capacitors in series with bleeder resistors across each unit to assist in voltage sharing. It is most important to provide proper insulation for the outer cans of the capacitors not directly earthed. They should be completely covered with insulating material to ensure the safety of the operator.

The 300 volt screen supply should be stabilized by two gas-filled stabilizers in series and can be either a separate supply or fed from the anode supply through a suitable voltage dropping resistor. The use of a combined unit ensures that at no time can the screen supply be switched on without the anode and although there is some waste of power due to the loss in the dropper resistor, it is probably the most convenient arrangement for the amateur. A suitable circuit is given in **Fig. 10.83.**

The demands on the bias supply are small and the load is stable. Apart from ease of setting the output voltage to the required value, no special requirements are involved other than adequate smoothing. A half-wave rectifier circuit with resistance and capacity smoothing giving about 50 volts to the output potentiometer would be suitable. The potentiometer should be part fixed and part variable to facilitate accurate setting of the required voltage.

The h.t. lead from the insulated safety socket at the rear is run along the left side of the cabinet and front panel to the meter, and is screened by copper tube to prevent stray r.f. pick-up. A six-pin Painton (miniature Jones) plug carries the other supplies.

Operation

The maximum time allowable for tune-up with tone on at 1200 volts should be limited to five minutes although during initial testing, periods in excess of 15 minutes without the fan running did not cause any valve distress. However, five minutes should be ample and will provide a good safety factor.

The resting current under each anode voltage condition allows for the same anode dissipation and is no worse for any of the voltages quoted. It was intentionally kept as high as possible to enable the best intermodulation distortion (i.m.d.) factor to be obtained. If these settings are significantly lowered, then some increase in i.m.d. will result.

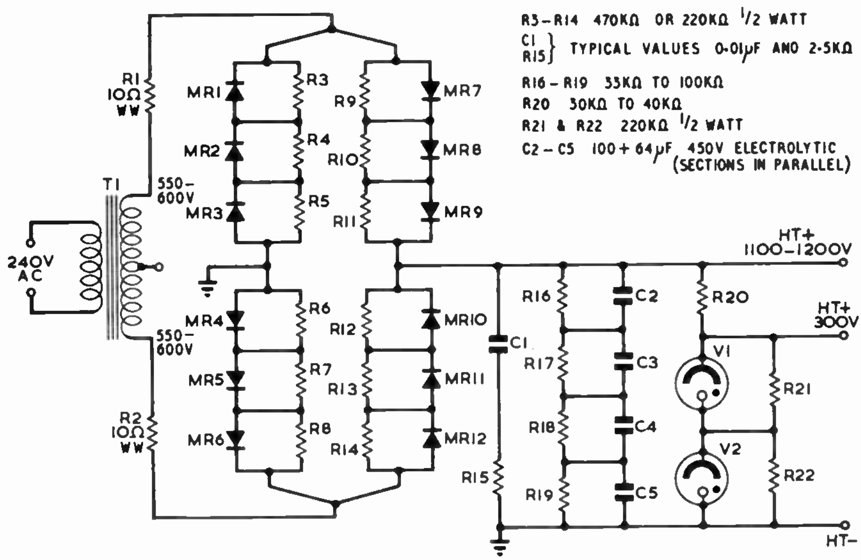


Fig. 10.83. Circuit of power supply for TT21 linear amplifier.

100 WATT OUTPUT LINEAR WITH VOLTAGE AMPLIFIER

This linear amplifier is provided with a voltage amplifier, and may be fully driven with a few volts of r.f.—for instance the output from the final mixer in an experimental exciter.

The circuit is shown in **Fig. 10.84** together with the EF80 voltage amplifier and the recommended bias supply. The grids of the QVO6-20s or 6146s must be isolated from the h.t voltage appearing on the coils and the tuning capacitor, and the 150 pF blocking capacitor must not be omitted.

As the 6146 valve has been specifically designed for class AB1 operation there is no grid current under the correct operating conditions and therefore no driving power is necessary. The output from the EF80 voltage amplifier is more than adequate to drive fully the p.a. to its maximum rating of 180 watts p.e.p. input on all bands.

The pi network values have been calculated for an R_L of 2000 ohms and are given for each band in **Table 10.15**. Construction of the tank inductance is simple, with the 160m coil L3 on a separate ceramic former positioned at right angles to the main winding L2 and adjacent to its " cold " end; the self supporting 10m coil L1 is positioned at the " hot " end of L2 to form the connecting link between L2 and the stator plates of the anode tuning capacitor VC3. It is convenient to use a standard broadcast receiver two gang capacitor of 500 pF each section for the aerial loading capacitor VC4. As this would not have a large enough value for use on the lowest amateur band, the 160m position of the pi tank band change switch is used to bring into circuit a further fixed loading capacitor of 600 pF or thereabouts. On the 160m band power is reduced to the equivalent p.e.p. of a 100 per cent modulated 10 watt d.c. input A3 transmitter, either by switching the p.a. anode feed from the 650/750 volt line to the 300 volt rail feeding the remainder of the transmitter, or alternatively breaking the heater supply to one of the 6146 valves by means of a panel mounted toggle switch.

TABLE 10.15

PI NETWORK DATA—TWO 6146 VALVES

R_L = 2000 ohms

BAND	X_{C1} = 200 ohms	X_L = 250 ohms	X_{C2} = 46 ohms
80m	220 pF	11·0 μH	900 pF
40m	110 pF	5·5 μH	450 pF
20m	56 pF	2·7 μH	225 pF
15m	38 pF	1·8 μH	160 pF
10m	28 pF	1·4 μH	115 pF

TANK COIL

L1, 4 turns 16 s.w.g. tinned copper 1 in. diam. spaced to ½ in. long, self supporting.

L2, One continuous winding of 18 s.w.g. tinned copper wound 16 turns per inch, 10 turns, 3/16 in. gap, 8 turns, 3/16 in. gap, 3 turns, 3/16 in. gap and 2 turns. Total length approximately 2 in. diameter 1½ in. supported by 4 lengths of ¼ in. diam. polystyrene rod (similar to Codar or Miniductor coils).

L3, 32 turns 22 s.w.g. enam. close wound on 7/8 in. diam. ceramic former.

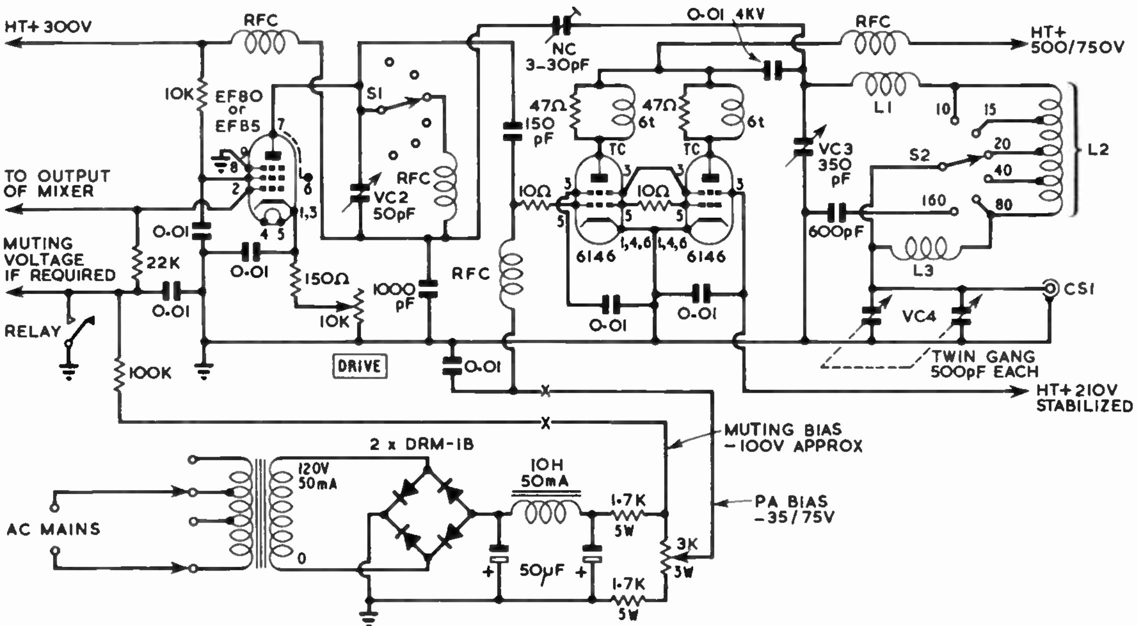


Fig. 10.84. 100 Watt linear with voltage amplifier. Modifications for operation on 160m are described in the text.

Layout

The main pi tank coil L2 is pre-assembled and supported by the connections to the switch S2 which is mounted on the panel.

The recommended screen voltage is 210 volts stabilized (two VR105s in series) while the anode supply may be any convenient value between 500 and 750 volts. The bias setting potentiometer should be adjusted until the standing no-signal anode current is at a value just within the total anode dissipation of the valves—40 watts. With a 700 volt supply this will be 50 mA. *For a clean signal, correct loading and driving is essential;* load to an indicated anode current of 200 mA and *never* talk the amplifier up beyond the point at which the milliameter pointer swing reaches a peak of half this value—100 mA. Remember at all times that the meter movement has inertia; it cannot follow at syllabic rate—if it is indicating 100 mA the true peak current will be at least twice this value.

300 WATTS OUTPUT PASSIVE GRID AMPLIFIER

A linear amplifier using two 4X150A valves is shown in **Fig. 10.85.**

The calculated value for R_L (anode load) is 1600 ohms and the values of C1, C2 and L in the pi tank are obtained by using the tables given in the *Radio Data Reference Book* or on page 10.35.

The circuit is equally applicable to other valve types, subject to modifications to the value of R and the values in the pi tank circuit to accommodate the different peak grid drive voltage and the different value of R_L. As an example, 813s and 4-125As require a peak r.f. grid drive of approximately 100 volts. A single 6146 valve with a p.e.p. output of 50 watts could not develop this across 80 ohms and R will need to be increased in value to 200 ohms. The calculated value for R_L is 3250 ohms for two 813s in parallel and 3750 ohms for parallel 4-125As—both with an h.t. supply of 2000 volts and 600 volts on the screens for operation in class AB1.

The passive grid method is equally applicable for class AB2 operation. The driving requirement is then the sum of the power dissipated in the passive grid resistor plus the driving power as given in the manufacturers' valve data. As will be seen from Fig. 10.85, the circuitry is inherently simple with only the pi network requiring tuning. The absence of the normal grid input tuned circuit and the usual neutralizing bridge makes for a particularly neat and compact layout. Any existing pi output circuit on the exciter—provided it has a variable capacity on the output side—will load satisfactorily into the passive grid load. If there is no pi tank, but a parallel tuned circuit arrangement, an output link winding with the link turns adjusted to give the maximum drive into the p.a. will be required.

400 WATT OUTPUT PASSIVE GRID AMPLIFIER

Many amateurs using s.s.b. are limited to operation with an exciter for two major reasons:

(i) the difficulty of obtaining valves suitable for a higher power amplifier at reasonable cost. and

(ii) the inherent fear of the danger of high voltages of 2000 to 3000 in the shack and the natural desire to restrict all power supplies to more familiar voltage levels around 700 or 800 volts.

There is the further attractive feature that a power supply of 1000 volts or less can be safely and compactly smoothed

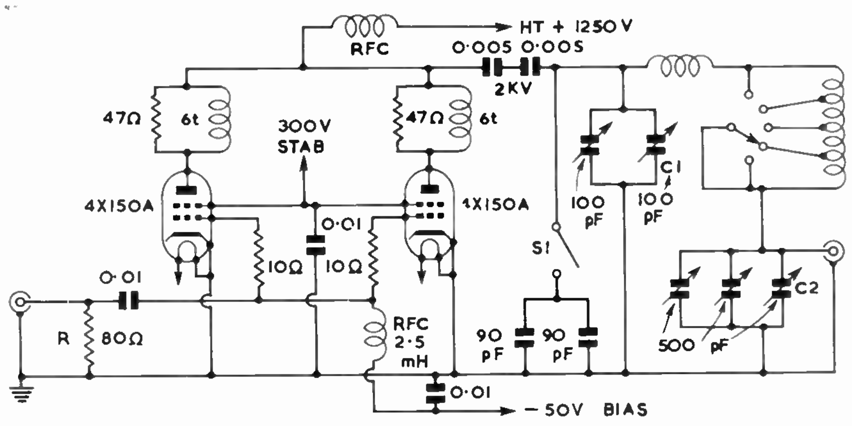


Fig. 10.85. Passive grid linear amplifier using two 4X150A valves.

C1 is a two-gang capacitor with 0·05 in. air gap. C2 is a standard three-gang 500 pF per section capacitor. R comprises 15 one watt 1·2 K ohms carbon resistors in parallel. For 80m operation, S1 is closed. The anode r.f. choke must be a type suitable for pi-network tank circuit operation—see page 6.42. It should be noted that the 4X150A valves require forced air cooling even without h.t. applied and the blower circuitry must be interlocked with the heater supply.

Operating Conditions

D.C. anode voltage	1250 volts
D.C. screen voltage	300 volts
Zero signal d.c. anode current	100 mA
Maximum signal d.c. anode current	400 mA
Maximum signal d.c. screen current	25 mA
D.C. grid voltage	−50 volts
Anode load	1600 ohms
P.E.P. output (calculated)	300 watts
P.E.P. output (measured)	270 watts *

* Including tank circuit losses.

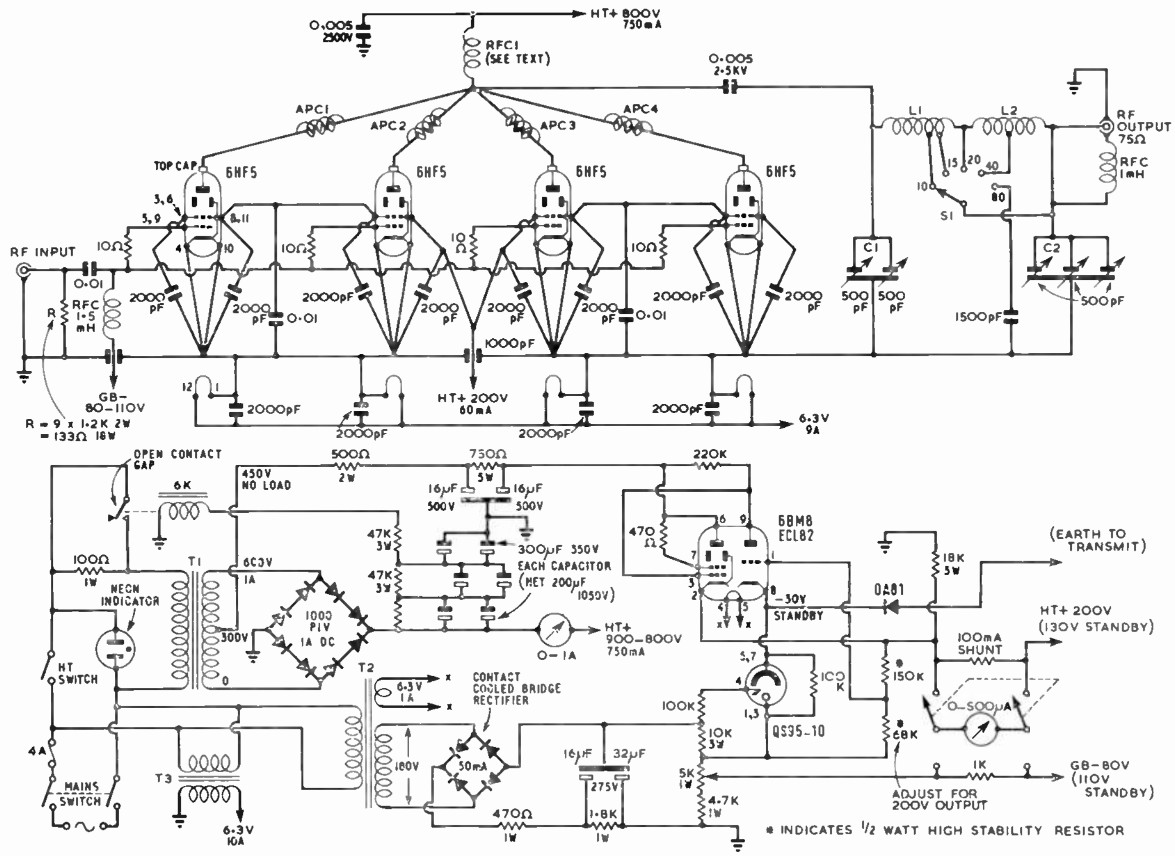


Fig. 10.86. Circuit of the 400 Watt passive grid amplifier and suitable power supply.

by low cost electrolytic capacitors in a simple series-parallel arrangement.

These considerations have caused designers to look for acceptable valve types in the low cost commercial television receiver field. The requirement for an s.s.b. amplifier is a high peak current at a low anode voltage, together with an anode dissipation capability in the range 20 to 40 watts. These characteristics are available in the modern colour television line time base output valves, and because these valves are manufactured in very large numbers the cost is relatively low.

Since these valves are designed and controlled in manufacture only for deflection amplifier service, manufacturers do not normally publish data or operating conditions for amateur single sideband service. Of very great importance is the capability of the valve to operate with an acceptably low level of third and fifth order intermodulation distortion products.

The complete circuit including the power amplifier is shown in Fig. 10.86. It will be noted that four Sylvania 6HF5 horizontal deflection valves are wired in parallel and that the amplifier uses the now common passive grid input circuit. Low voltage high current valves will require a low R_L (anode load) and as four valves in parallel will require one quarter the R_L of one, the final value will necessitate very low reactance values in the output tank circuit. This means few turns on the coil and quite unconventional values for C1 and C2. Fortunately the requirements can be met using standard broadcast type variable tuning capacitors, plus an additional fixed 1500 pF capacitor that is paralleled with C2 on the 80m band position of the pi-tank switch S1.

It is important to realize that the amplifier is not neutralized and that stability is dependent on the damping of the grid input circuit by the passive grid resistor. A low value of 50 to 75 ohms will give improved stability but requires a greater driving power to develop the wanted peak r.f. grid swing. With the value of 133 ohms given in Fig. 10.86, the amplifier can be driven fully with a single 6146 in the exciter. It is, however, of greater importance that stray r.f. feedback between the anode and grid sides of the amplifier be kept to a minimum. This means careful attention to the layout so that the input connections are kept underneath the chassis and all the components associated with the anode circuits on top i.e. the grid input side must not be able to " see " any part of the anode output side of the p.a. valves—and thorough bypassing of both the screen grid connections on each valveholder.

The mains transformer T1 provides a secondary output of 600 volts r.m.s. centre-tapped with a maximum current rating of 1000 mA. Both primary and secondary windings have very low resistance, and this, together with the low resistance silicon rectifiers and large effective value of smoothing capacity, produces a very heavy switch-on current surge that would blow the mains input fuse. It would be unwise to increase the fuse rating beyond 4 amps because this would defeat its protective function and the switch-on surge current is held to a normal value by the 100 ohm 1 watt resistor in series with the transformer primary. As the smoothing capacitors charge up towards their maximum value, the rising voltage increases the current through the 6000 ohm relay winding until the contacts finally close and short out the 100 ohm current limiting resistor. The relay is adjusted to close when the h.t. output reaches approximately 600 volts by adjustment to the contact gap—this will be wider than the normal gap setting.

Operating bias for the 6HF5 control grids and the screen supply regulator valve (ECL82, 6BM8) is provided by T2 and the 50 mA bridge rectifier with conventional smoothing and output load circuit. The p.a. bias supply is adjusted by means of the 5 K ohm potentiometer for a total zero signal anode current of 120 mA (30 mA each valve). On " stand-by " the cathode of the ECL82 triode section is at negative 30 volts, the 6HF5 bias line at negative 110 volts and the screen grid supply at positive 130 volts—this reduces the standing zero signal anode current to a very low value and enables the valves to cool down on pauses between transmission periods. For " transmit " the triode cathode is shorted down to earth via the exciter " press-to-talk " relay and muting line, the 6HF5 grid bias is then negative 80 volts (or whatever value is necessary to maintain a total zero-signal anode current of 120 mA) and the screen supply is positive 200 volts. It is necessary to incorporate the series OA81 diode in the muting line to prevent the 100 volt negative muting voltage on the G2DAF exciter rail from reaching the ECL82 (6BM8) triode cathode—the diode merely acts as a one-way gate, the characteristics not being critical, and any similar type to the OA81 is therefore suitable. It is, of course, not essential to use the regulator valve arrangement shown in Fig. 10.86, and an alternative arrangement that might appeal to some constructors is the more usual screen feed via a dropper resistor and VR105/30 voltage regulator valves. As the combined maximum signal screen current for the four 6HF5 amplifier valves is 40 mA, it will be necessary to use four VR105/30s in a series parallel arrangement to give 210 volts regulated at a maximum current capability of 60 mA.

For a maximum signal anode current of 750 mA at an anode voltage of 800, $I_a(_{peak}) = I_a(_{d.c.}) \times K$ (where K is a constant dependent on the angle of anode current flow—in this case 200° giving a value for K of 3). $I_a(peak)$ therefore equals 750 × 3 = 2250 mA. Assuming that the dynamic anode voltage swing at its lowest point is 100 volts, from the formulae $R_L = 2(V_a - V_{a\ min})/I_a(_{peak})$ the values are $R_L = 2(800 - 100)/2.250$. This equals an R_L of 622 ohms. The pi constants are then $R_L = 622$ ohms; $R_{L\ out} = 75$ ohms. The ratio $R_L/R_{L\ out} = 8{\cdot}3$ and the square root of 8·3 is approximately 2·9, the reactance ratio, ($XC1 : XC2$). Designing for a Q of 12;

$XC1 = R_L/Q = 622/12 = 52$ ohms.
$XC2 = XC1/2{\cdot}9 = 52/2{\cdot}9 = 18$ ohms.
$XL = XC1 + XC2 = 52 + 18 = 70$ ohms.

These values are a simple approximation but are quite near enough for amateur purposes. From a reactance chart the values for 80m are C1 = 900 pF total (825 µF + 75 pF stray); C2 = 2700 pF and L = 3·3 µH. Values for the other bands scale down in the same ratio as the band wavelength as follows:

Band	*C1*	*L*	*C2*
80m	900 pF	3·3 µH	2700 pF
40m	450 pF	1·6 µH	1350 pF
20m	225 pF	0·8 µH	675 pF
15m	172 pF	0·6 µH	500 pF
10m	112 pF	0·4 µH	337 pF

Construction

Chassis layout is not critical but all leads carrying r.f. should be kept as short as possible—the amplifier should be built as if it were going to operate on 4m—with the 6HF5 valves arranged in a square and the anode r.f. choke in the centre. C1 and C2 should be above the chassis on either side of L1-L2 and S1.

The main h.t. smoothing capacitors are standard television types with a rating of 300 µF at 350 volt working. As the cans are not isolated the six capacitors must be mounted on an insulated panel of paxolin or perspex. C1 is also a two gang broadcast type capacitor of 500 pF each section with not less than 25 thou. spacing between rotor and stator plates. C2 is a standard three gang broadcast type tuning capacitor of 500 pF each section. Both the pi-tank inductances L1 and L2 are mounted in the same plane and supported by the switch S1 which should have ceramic insulation and have substantial contacts suitable for carrying several amps of r.f.

TABLE 10.16

Amateur Band operating conditions (max. 400 watts p.e.p.)

V_a = 800 volts	*Single-tone*	*Two-tone*
Anode current (zero sig.)	120 mA	120 mA
Anode current (maximum sig.)	750 mA	475 mA
Power input (d.c.)	600 watts	380 watts
P.E.P. input	600 watts	600 watts
Ig_2	40 mA	24 mA
Ig_1	0 mA	0 mA
Vg_2	200 volts	200 volts
Vg_1 (peak drive)	92 volts	92 volts
Anode Dissipation	200 watts	180 watts
P.E.P. Output	400 watts	400 watts
Power Output (mean)	400 watts	200 watts
Anode Efficiency	66 per cent	52 per cent

Operation

The amplifier should be tuned for maximum r.f. output and then the loading adjusted by C2 to give the screen currents given in Table 10.16, single tone input 40 mA, two-

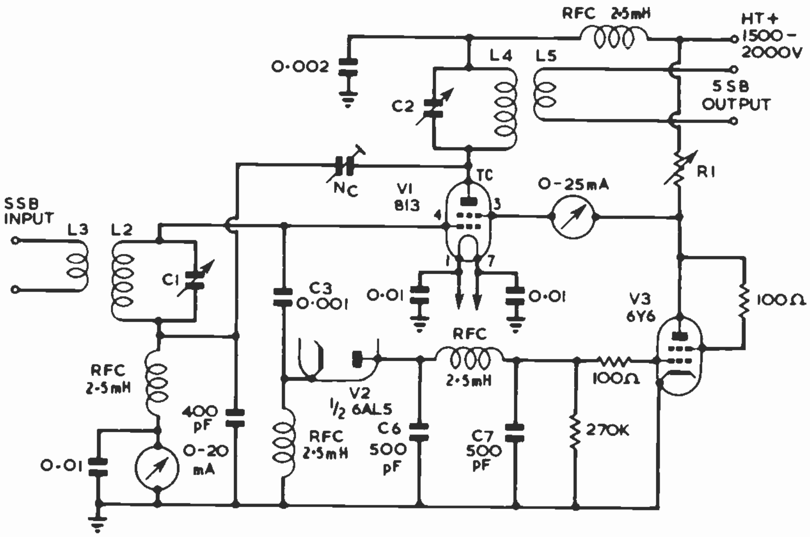


Fig. 10.87. G2MA linear amplifier. For adjustment of R_1 refer to text.

tone input 24 mA. Maximum r.f. output occurs at maximum screen current and not necessarily at minimum dip. (This is due to the relatively large anode to grid capacity of four valves in parallel—the effect is more pronounced at higher frequencies.) Operating conditions up to the maximum allowable p.e.p. output of 400 watts are also given in Table 10.16.

THE G2MA LINEAR AMPLIFIER

A contrast to the amplifiers which have already been dealt with is provided by the circuit of **Fig. 10.87**. This design, for which credit is due to David D. Marshall, G2MA, has proved so easy to construct and adjust that it has been copied successfully by amateurs all over the world. Several modifications, aimed principally at simplifying this circuit, have been described but none of them quite matches the performance of the original design, despite the fact that they call for more critical adjustment. While the peak output from the 813 will be lower than that obtainable from a similar valve in a conventional class AB2 circuit, and the linearity is marginally poorer, the G2MA amplifier may be put into operation relatively simply. Stabilization of the screen supply is not required, and any bias which may be needed can be provided by a couple of dry cells.

As will be clear from Fig. 10.87, screen voltage for the amplifier valve is taken from the anode supply via the dropping resistor R1. In the absence of excitation, the clamping valve V3 holds the screen potential down to a low value, so that the anode current passed by V1 is small, despite the absence of fixed bias. For best results, a low impedance valve designed for voltage regulation should be used in the V3 position. The 6Y6, originally recommended by G2MA, is very good, and the 6AS7 ought to be even better, but the more common tetrodes such as the 6L6 and 807 are not suitable. When drive is applied, the negative r.f. pulses applied to the cathode of V2 via the capacitor C3 will be rectified, smoothed by the filter C6, RFC and C7, and applied to the grid of the clamper valve as a negative d.c. voltage which fluctuates at the same rate as the original modulation. This bias causes the current through V3 to vary in sympathy with the modulation, and as the current falls, the drop across R1 decreases and the voltage applied to the screen of V1 rises. While the value of R1 is not critical, it must not be too high, or it may limit the maximum value to which the screen potential can rise and thereby restrict the output of the stage. For optimum results the resistor should be chosen so that the current drawn by V3 under no-drive conditions is just below the value at which anode dissipation commences; but performance will not be materially affected if it is adjusted to pass about 50 mA. In the original version of the amplifier, the designer used a fixed resistor in the grid return of the 6Y6 clamper valve, as shown in Fig. 10.87. Some users claim that adjustment may be made easier and linearity improved if a variable resistor is substituted. If this modification is incorporated, the amplifier should be operated under normal speech conditions and the variable resistor adjusted until the screen voltage rises smoothly to 400 volts on peaks.

The G2MA amplifier is no exception to the general rule that the highest obtainable anode voltage should be used if maximum output is desired but when a supply much in excess of 1500 volts is used, the power loss in R1 becomes excessive. A separate low current supply of about 1000 volts may be provided to reduce this loss, or if full wave bridge rectification is used in the power supply to the final amplifier, a half-voltage tapping may be taken off a centre tapped transformer secondary winding.

If maximum output is required, the G2MA amplifier should be adjusted in the same manner as a conventional linear, using a two-tone test signal and oscilloscope. Tolerably good results may, however, be obtained at lower output by setting the drive so that the grid current meter reads 5 mA with steady modulation. In the absence of an audio oscillator, a sustained whistle into the microphone will serve.

Loading should be increased smoothly, while observing closely the anode current meter. Beyond a certain point the current will cease to rise regularly with further increase in loading; the correct coupling occurs just before this point of inflexion is reached. The anode current should be noted, and when the amplifier is used with normal speech modulation, the meter should not be allowed to kick higher than one third to one half of the value under sustained single-tone modulation. The screen current will peak to about 10 mA. As a final test, a local amateur should be asked to monitor the signal for splatter. If any is found, the coupling should be reduced until it disappears. Some reduction in drive may also be needed. Although this procedure will enable a signal to be put on the air with a minimum of equipment, it leaves much to be desired, and adjustment with a cathode ray tube monitor is to be preferred.

THE G2DAF LINEAR AMPLIFIER

In this linear amplifier the valves are operated under zero bias conditions but without the complication of screen dropper resistors, stabilizer or clamper valves. The screen voltages are derived by rectifying a small portion of the input signal. The amplifier may be driven by the s.s.b. transmitter described on page 10.48 without modification.

The complete circuit diagram is shown in **Fig. 10.88**. The diode rectifiers may be either semiconductors of the point contact type suitable for r.f. use, or alternatively thermionic valves. Suitable rectifier valves are available at low cost and were used in the original amplifier. The main requirement is a good heater-cathode insulation and a heater rating suitable for the additional 6 volt winding generally provided on standard p.a. heater transformers. Suitable valves are the Brimar 6U4G or the Mullard EY81. The Brimar valve is preferred because the anode connection is brought out to a base pin and, as this is underneath the chassis, it is screened from the p.a. output circuits.

The pi tank coil is wound on an Eddystone 2½ in. diameter ceramic former grooved eight turns per inch, and this is

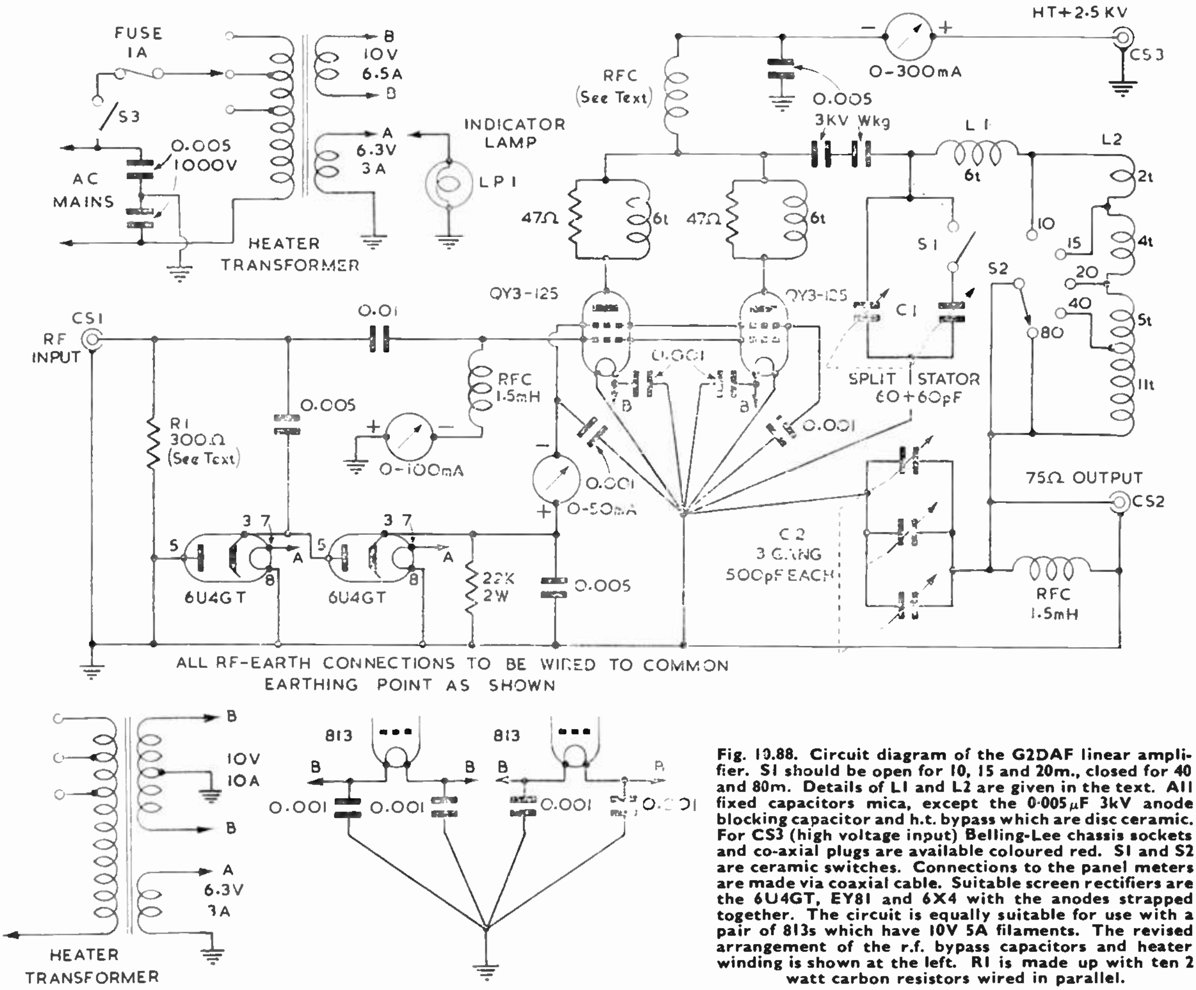


Fig. 10.88. Circuit diagram of the G2DAF linear amplifier. S1 should be open for 10, 15 and 20m., closed for 40 and 80m. Details of L1 and L2 are given in the text. All fixed capacitors mica, except the 0·005µF 3kV anode blocking capacitor and h.t. bypass which are disc ceramic. For CS3 (high voltage input) Belling-Lee chassis sockets and co-axial plugs are available coloured red. S1 and S2 are ceramic switches. Connections to the panel meters are made via coaxial cable. Suitable screen rectifiers are the 6U4GT, EY81 and 6X4 with the anodes strapped together. The circuit is equally suitable for use with a pair of 813s which have 10V 5A filaments. The revised arrangement of the r.f. bypass capacitors and heater winding is shown at the left. R1 is made up with ten 2 watt carbon resistors wired in parallel.

attached to the switch S1 before it is fitted to the panel. A gap of one groove is left between the the 15 and 20, and 20 and 40m sections. The total winding length of 16 s.w.g. tinned copper wire is 3¼ in. For the 10m band the coil of six turns of 12 s.w.g. wire is spaced to approximately 2 in. long and is self-supporting—with the axis at right angles—between the tuning capacitor and the end of the main tank coil.

The pi tank values depend on the required value of anode load (R_L) and as with any amplifier it is important that the valves are operating into the correct load. If they are not, both the power handling capability and the efficiency will suffer. Assuming that V_a is the h.t. supply voltage; $V_{a(min)}$ the instantaneous anode voltage at its lowest point; $I_{a(dc)}$ the maximum signal anode current meter reading; then $I_{a(peak)} = I_{a(dc)} \times K$ (where K is a constant dependent on the angle of current flow—in this case approximately 3) and $R_L = 2(V_a - V_{a\ min})/I_{a\ peak}$. For the amplifier under consideration $R_L = 2(2500 - 250)/0{\cdot}3 \times 3 = 5000$ ohms.

The pi constants are then $R_L = 500$ ohms; $R_{L(out)} = 75$ ohms. The ratio $R_L/R_{L(out)} = 5000/75 = 66$. The square root of 66 is approximately 8 and this is the reactance ratio (XC1 : XC2). For a Q of 12:

$$XC1 = RL/Q = 5000/12 = 416 \text{ ohms}$$
$$CX2 = XC1/8 = 416/8 = 52 \text{ ohms}$$
$$XL = XC1 + XC2 = 416 + 52 = 468 \text{ ohms}$$

These values are a simple approximation but are quite near enough for amateur purposes. From a reactance chart the values for 80m are C1 = 116 pF; L = 20 μH; C2 = 900 pF. Values for other bands scale down in the same ratio as the band wavelength as follows:

Band	*C1*	*L*	*C2*
80m	116	20	900
40m	58	10	450
20m	29	5	225
15m	20	3·5	160
10m	15	2·5	113

The type of transmitter tuning capacitor suitable for C1 generally has semi-circular rotor plates and therefore a large minimum capacity value—usually about 15 to 20 pF. This, together with circuit and valve anode capacity, will make up a total that is greater than required for the 10 and 15m bands. It is possible to overcome this in two ways, (i) redesigning the tank circuit for a higher Q value of, say, 15 or 20; (ii) reducing the minimum capacity of C1. A high value of Q in the tank coil will increase the circulating r.f. currents and therefore the losses. Accordingly the second expedient has been adopted and the tank capacitor was, in fact, made into a two-gang unit of 60 pF each section by sawing through the bars holding the stator plates. One section only is connected to the 10m coil and the anode r.f. blocking capacitor and this tunes the three higher frequency bands. The other section is switched in parallel for the 40 and 80m bands. In addition to reducing the minimum capacity value, this method also doubles the dial bandspread and makes tuning less critical on the 10, 15 and 20m bands.

The required air gap for C1 is approximately one-tenth of an inch. A standard three gang broadcast tuning capacitor of 500 pF each section is suitable for C2 and provided the amplifier is working into a load (as it should be) the plate spacing is ample to prevent flashover.

The r.f. choke comprises 300 turns of 32 s.w.g. enamelled wire wound in unequal sections—165, 65, 35, 20 and 15—on a ceramic former 1 in. in diameter and 5½ in. long with a ⅛ in. spacing between each section. Standard multi-section pie wound r.f. chokes are unsuitable for pi-tank circuits and should not be used. A standard 1·5 mH r.f. choke rated for at least 300 mA is connected across the output co-axial socket as a safety precaution that should never be omitted when high voltages are in use. Should there be failure of the r.f. blocking capacitor the h.t. current through the choke will blow the main h.t. fuse and prevent h.t. voltage reaching the aerial circuits.

Construction

Constructors of any high power linear amplifier should realize that higher power may bring increased TVI problems. All reasonable precautions should therefore be observed. These include thorough screening, with v.h.f. r.f. chokes and non-inductive bypass capacitors on all outgoing supply cables, a single earthing point for all return paths for C1, C2, the screen bypass capacitors and the heaters of the p.a. valves. All connections carrying r.f. should be made with copper strip at least ⅛ in. wide and 10 thou. thick and not more than 3 in. long—every endeavour should be made to keep these connections as short as possible and in fact the amplifier should be built as if operation is intended on 70 Mc/s. If the amplifier is using one valve with the heater

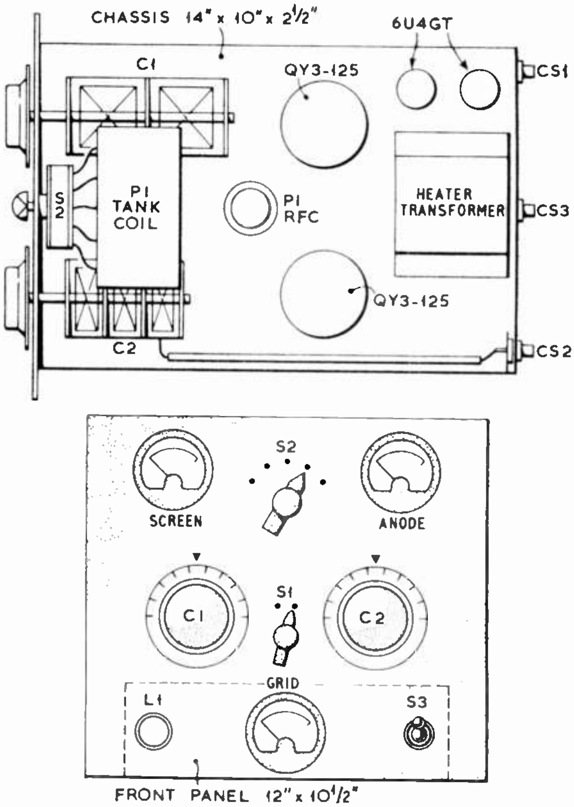


Fig. 10.89. Chassis and panel layout for the G2DAF Linear Amplifier.

winding floating and the transformer centre tap earthed, both heater pins of the valveholder must be effectively bypassed for r.f. with mica or ceramic capacitors taken to the common earthing point. On the higher frequency bands the transmitter output should be fed into the aerial via an efficient low pass filter giving at least 60db suppression at television frequencies in Band I (see Chapter 18—*Interference*).

Fig. 10.89 shows a suitable chassis and front panel layout.

Operation

Tuning and loading is exactly the same as a conventional class AB amplifier. Initially the drive level is increased until the anode current meter reads 150 or 200 mA. With the loading capacitor fully meshed, the anode tuning is adjusted for a dip in anode current. With the tank circuit at resonance, the screen current will be a high value. As the loading is increased by reducing the capacity of C2, the anode current will rise and the screen current will fall in the usual manner. The drive can now be increased until the grid, screen and anode currents are the required values.

Should the amplifier have been built using some other type of valve, the manufacturers' figures for class AB1 or AB2 working can be used initially. If an oscilloscope is available the amplifier should be driven with a two-tone input and the modulation envelope monitored on the c.r.t. It is then a simple matter to adjust excitation, tuning and loading for maximum r.f. output consistent with adequate loading to prevent flat topping or other distortion of the modulation envelope. The grid, screen and anode currents are then noted and in all subsequent operation the amplifier is adjusted to obtain these values. If an oscilloscope is not available, the loading should be adjusted so that the dip in anode current is not more than about 20 per cent of the off-resonance value, i.e. 250 mA off-resonance—load to 200 mA at resonance. A golden rule to observe is, " If in doubt, load heavily! " Under speech conditions adjust the exciter audio gain or r.f. drive so that the anode meter does not swing beyond half the steady signal value. Remember that the meter movement cannot follow at syllabic rate—if it swings up to 150 mA, the true maximum signal current is at least twice this value.

TABLE 10.17

Amateur Band operating conditions (400 watts p.e.p. output) for two QY3-125 valves. Va = 2·5 kV.

	Single-tone	Two-tone
Anode current	250 mA	175 mA
Power input (d.c.)	650 watts	440 watts
P.E.P. input	650 watts	660 watts
I_{g_2}	38 mA	22 mA
I_{g_1}	70 mA	45 mA
V_{g_2}	105 volts	75 volts
V_{g_1} (r.m.s.)	65 volts	64 volts
Driver Load (p.e.p. out)	30 watts	35 watts
Anode Dissipation	250 watts	240 watts
Power Output (mean)	400 watts	200 watts
P.E.P. Output	400 watts	400 watts
Anode Efficiency	61·5 per cent	46 per cent

Typical operation with 300 ohm grid swamping resistor.

With an anode supply of 3 kV the efficiency figures will show an improvement on those quoted. It is estimated that they will be as follows:

Single-tone efficiency = 72 per cent. Two-tone efficiency = 58 per cent.

THE 813 TETRODE

In many ways the 813 is a quite remarkable valve—in fact its performance in this amplifier is quite outstanding. With 2·5 kV anode potential it is possible to run *one* valve to the full maximum licence rating of 400 watts p.e.p. output. This is possible without degradation of linearity. Reports received during three months experimental operation on the 80 metre band all indicate that distortion products were at a level of 5 to 6 db better than with the same valve used at its full output power in conventional Class AB1.

Because the valve exceeds the rating only on voice " peaks " and has resting periods in the " troughs " the average dissipation over a period of time embracing a number of words is less than half of the peak-signal value. A valve may therefore be grossly overdriven without suffering damage.

TABLE 10.18

One 813 Valve (V_a = 2·5 kV)	Single-tone	Two-tone
Anode Current	260 mA	200 mA
Power in (d.c.)	650 watts	500 watts
P.E.P. input	650 watts	650 watts
Screen current	17 mA	8 mA
Grid current	25 mA	12 mA
Screen voltage	210 volts	170 volts
Grid Drive (r.f.)	110 volts r.m.s.	105 volts r.m.s.
P.E.P. output	400 watts	400 watts
Power output (mean)	400 watts	200 watts
Anode efficiency	61·5 per cent	40 per cent

The 813 running to a p.e.p. output of 400 watts is being operated at a level where the two-tone input power is 500 watts d.c. and the anode dissipation is 300 watts. This is greatly in excess of the manufacturers' rating of 125 watts, and while it is permissible to do this under voice input conditions where the valve is only being driven to its peak input at syllabic rate, it would be dangerous under continuous single-tone or two-tone input conditions. This means that the amplifier—at maximum input—cannot be correctly loaded or monitored on the oscilloscope, other than for a very short period of time. For this reason it is recommended that a single 813 is not driven to a power output of more than 200 watts p.e.p. if continuous 2-tone measurements are intended.

At 2·5 kV anode potential the zero signal anode current (one valve) is 45 mA and the anode dissipation 112·5 watts. This is just inside the rated dissipation of 125 watts and therefore limits the maximum h.t. supply voltage to 2·5/2·75 kV. For those who do not like high voltage power supplies, the valve will operate just as satisfactorily in regard to low distortion product level with 1·2/1·5 kV, but obviously the output power and the efficiency will be less.

The 813 is a high voltage-low current valve and therefore requires a high value of anode load (R_L). For one valve at 2·5 kV anode voltage the recommended value of R_L is 10,000 ohms, and for two valves in parallel 5000 ohms. **Table 10.18** gives the maximum signal operating conditions for both single-tone and two-tone inputs. Reference to the Mullard manual *Valves for Single Sideband Suppressed Carrier Service* will show, however, that at 2·5 kV anode potential the usual figure for two-tone efficiency for the modern tetrodes is nearer to 40 per cent and in practice any amplifier giving an efficiency figure for two-tone input of between 40 and 50 per cent is doing very well and can be considered to be operating in a satisfactory manner. Fortunately the s.s.b. operator is rated by the GPO on output power so the lower efficiency of the sideband linear amplifier is not of any particular importance. What is of particular importance is the amplifier linearity—the better the linearity the lower the distortion product level. It is far better to operate at a lower level and sacrifice some of the efficiency.

POWER SUPPLIES

It is possible to use almost any power supply to provide operating potential for a single sideband exciter and linear amplifier. Nevertheless, if a first class signal is the aim, it is well worth the effort involved either to re-build making use of those components that are suitable, or alternatively to start with a circuit specifically designed to meet the specialized requirements of s.s.b. equipment.

These requirements may be summarized as follows:

(i) Carrier oscillators, balanced modulators and v.f.o.'s must be fed from a stabilized supply—usually at 150 volts.

(ii) Any hum ripple on the 250–300 volt exciter supply to the carrier oscillator or the audio valves would unbalance the modulator and produce a hum modulated carrier. This necessitates above average smoothing and large capacity values in the rectifier filter network.

(iii) A rough " carrier " will also be produced if any 50 c/s mains ripple is induced via a cable harness (between a separate power pack and the exciter) in which one side of the heater supply is common to the h.t. negative feed cable. There is also the risk of mains hum being induced due to heater cathode proximity in the audio or carrier oscillator valves. This difficulty can be overcome by a two-wire heater circuit that is balanced to earth.

(iv) An r.f. power amplifier can only operate in a linear manner if the anode voltage remains constant in potential. The current demand is at syllabic rate and may vary over a range 50 mA to 500 mA or more (depending on the type of amplifier) while transmitting. This entails a supply with good *dynamic* regulation.

(v) A class AB1 amplifier can only operate in a linear manner if the screen supply is held absolutely constant in potential at all times. It is therefore necessary to provide the screen voltage recommended by the valve manufacturer that will hold its regulation over the " zero signal "–" maximum signal " current range.

(vi) An r.f. power amplifier can only operate in a linear manner if the negative control-grid voltage remains constant in potential. There must also be provision to vary the voltage output by a pre-set control—this is necessary to be able to set the zero signal standing anode current to the optimum value, either as given by the valve data sheet, or obtained during adjustment with the oscilloscope. This means a " stiff " bias supply.

While very good smoothing is necessary in the low voltage supply to the exciter, the conditions in regard to the high voltage supply to the p.a. valves are quite different because the anode current of a tetrode is almost independent of small changes in anode potential and hum ripple will not anode modulate the signal. The dynamic regulation over the range of 50 mA to 500 mA is, however, very important because a fall in anode voltage at the moment of peak current demand would prevent the p.a. handling the peak signal and could cause flat topping and distortion. In the high voltage supply the customary smoothing chokes are therefore omitted and a large value of reservoir capacity is placed directly across the rectifier output.

The bias supply is made " stiff " by the use of a generously rated mains transformer and rectifier together with ample capacity and a relatively heavy bleed current through a resistive load network.

AN EXCITER POWER SUPPLY

A power supply meeting the specified requirements is shown in Fig. 10.90. This is suitable for feeding an exciter using 5B/254M or 6146 driver valves, normally run with 700–750 volts on the anodes. All components can be mounted on a chassis measuring 14 in. × 10 in. × 3 in. deep. Points to note are:

(a) The heater voltage of 6·3 volts a.c. is balanced and " floating " above earth.

(b) The bias supply is stiff and also provides a wide range of adjustment.

(c) The 50 μF capacitor between the 200 volt rail and earth ensures, together with the bleed through the series resistors and stabilizer valve, a stiff screen supply with a low impedance to the demand that will vary at audio rate.

(d) The high value of effective capacity of 80 μF across the 5R4 rectifier valves gives an output voltage of 700–750 from a secondary winding of 550 volts and ensures excellent dynamic regulation.

S1 is provided so that the anode h.t. can be reduced to permit operation on the 160m band without exceeding the licensed power limit. This switch must never be used for tuning and loading purposes on the other bands. A 6146 valve fully driven with a low voltage on the anode and normal voltage on the screen will take excessive screen current, the rated screen dissipation will be exceeded and the valve may be destroyed.

POWER SUPPLY FOR A LINEAR AMPLIFIER

Fig. 10.91 shows the circuit of a power amplifier power supply. T1 is a transformer specially made for 866s with 5000 volt insulation, and the high voltage transformer T2 was manufactured by Woden. The absence of smoothing chokes together with a total of 24 μF capacity ensures good dynamic regulation. Resistors R3 and R4 provide a steady 50 mA bleed current and also ensure that the reservoir capacitors are discharged when the h.t. transformer is off.

The bleed current through R1 and R2 in series is the full 40 mA rating of the VR150s—this ensures that should the p.a. screen current exceed 40 mA the regulators will go out and the potential on the screen rail will fall and protect the amplifier valves.

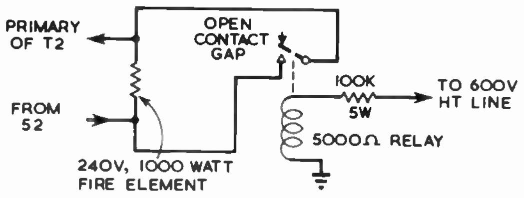


Fig. 10.92. Relay-operated surge limiting circuit.

Mercury vapour rectifiers such as the type 866 require the heater to be at full operating temperature before the anode voltage is applied. Switch on sequence is therefore S1, a wait of not less than two minutes, then S2. (When initially put into service the 866 valves should be run for at least 15 minutes with heaters only, before anode voltage is applied, to remove all traces of mercury from the cathode.)

As shown in Fig. 10.91 there is no provision of any kind to limit the initial surge current and theoretically this could strip the rectifier valves.

If surge limiting is required, this may be provided by connecting a 60 or 100 ohm wire wound resistor in series with the switch S2 and the primary winding of T2. (A 240 volt 1000 watt spiral fire element wound round a ceramic tube is very suitable because it may be left in circuit for an indefinite period—for instance while making preliminary amplifier adjustment with reduced anode voltage.) When the h.t. output voltage rises to its maximum value, the surge limiting resistor may be short circuited manually or by a 5000 ohm relay taking its energizing current via a 100 K 5 watt resistor connected to the 600 volt screen supply. A suitable circuit is shown **Fig. 10.92.**

SILICON RECTIFIER POWER SUPPLY

It is possible to obtain dual voltage output—700 volts and 300 volts—from one 350-0-350 volt secondary winding by means of silicon rectifiers in a bridge circuit. The regulation of the output voltage is directly dependent on the quality of the mains transformer which should have a low resistance secondary winding. It should preferably be rated for at least 450 mA current drain.

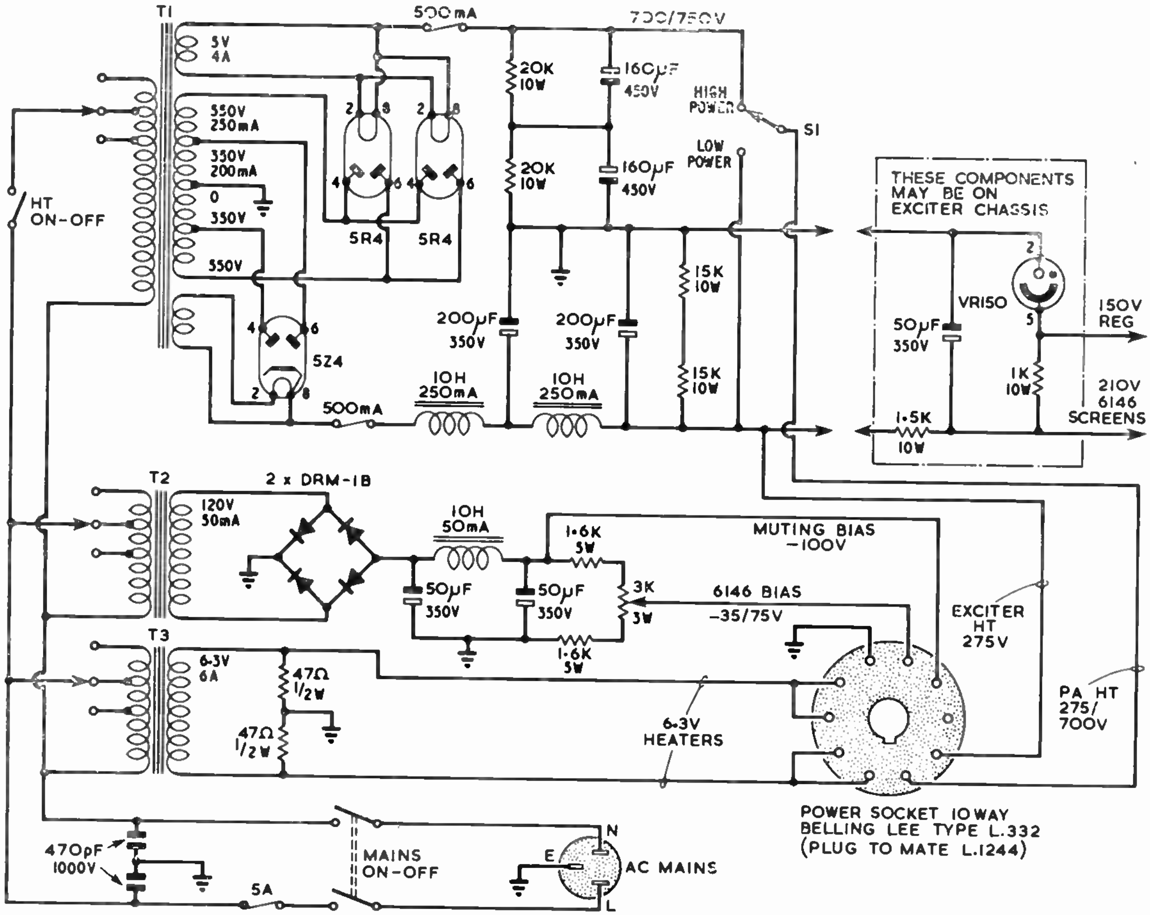


Fig. 10.90. Circuit diagram of a power supply suitable for use with an exciter.

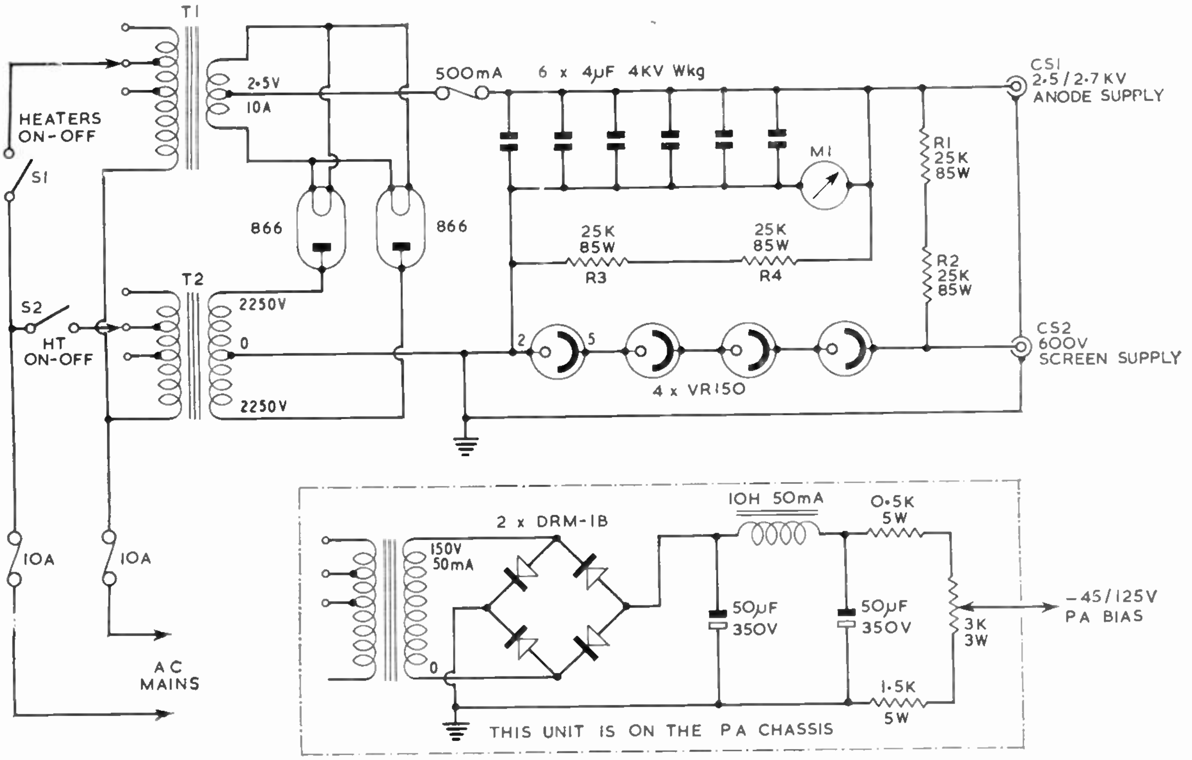


Fig. 10.91. Power supply for a linear amplifier. R1 and R2 and the VR150 stabilizers may be on the p.a. chassis. The secondary of T2 is rated at 500 mA. M1 is a 3000 volt meter.

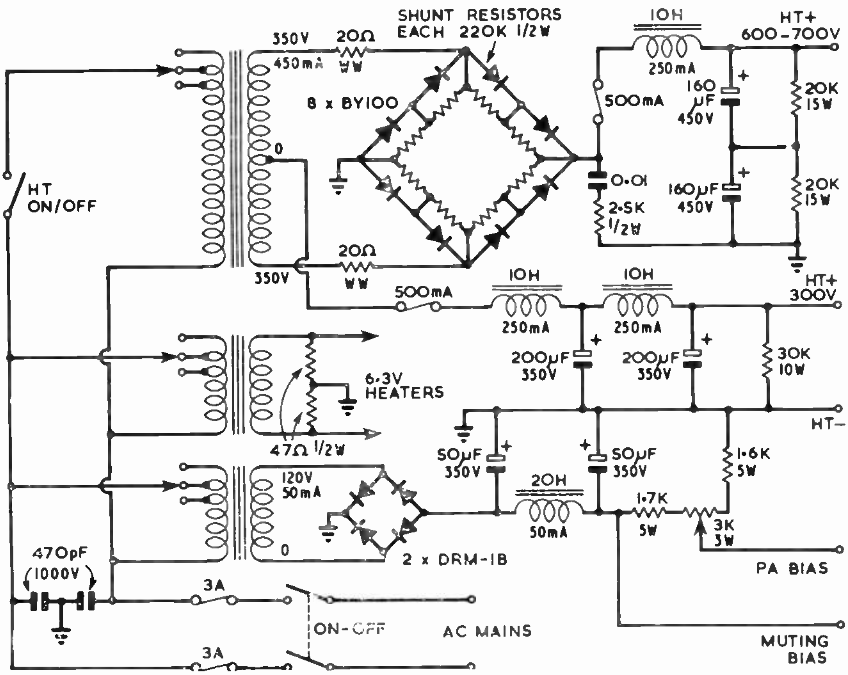


Fig. 10.93. Dual voltage power supply using silicon rectifiers.

A suitable circuit is shown in **Fig. 10.93**. The h.t. bridge requires eight silicon rectifiers and these should each be rated for not less than 800 volts peak inverse. Shunt resistors of 220 K ohms must be wired across each silicon rectifier in order to stabilize the operating conditions. The 0·01 μF capacitor in series with the 2·5 K resistor is a protection against mains supply switching transients.

Construction

Constructional details for this and the two previous power supplies are not given since these will vary with the constructor's requirements. The layouts are not critical, but if major components other than those shown are to be used, reference should be made to Chapter 19 (*Power Supplies*) for suitable parameters.

CONTROL SYSTEMS

One of the major advantages of single sideband operation is the ability to be able to carry on a natural conversation. It is customary to control the transmitter and receiver with muting bias that is switched with a change-over relay operated by a press-to-talk button or by a VOX (automatic voice control) system.

Many experienced s.s.b. operators consider that the additional components and complication of a fully automatic VOX system is not worth while; everything the voice control system can do can be done equally well with a simple push-button mounted on the microphone or a foot switch. This is, however, a matter of personal opinion and every operator is entitled to use the system that appeals to him best.

Ideally, the muting bias would be switched electronically and there would be no problem of clicking relays. There is the further problem of aerial change-over, and if the system is electronic the aerial switching must also be electronic using some form of TR switch. Unfortunately the most satisfactory TR switch is the high impedance type linked back to the linear amplifier output circuits, and a s.s.b. amplifier—drawing a considerable zero-signal anode current—is a most efficient noise generator. To stop the noise generation the p.a. must be made non-conducting during receiving periods; this is possible either by removing the screen potential or by applying additional negative voltage to the control grid. However, it is not practicable if the amplifier happens to be a zero bias triode or a tetrode in a G2DAF amplifier circuit where there is neither bias or screen voltage. There is also the problem for those amateurs using mercury vapour rectifiers in getting rid of rectifier hash.

In general, aerial change-over relays are used with a further relay as the " master control " built into the exciter. This system is simple and works very well. The only complaint is that of relay noise—this applies particularly to the aerial relay. This is often due to the use of heavy duty, wide gap, aerial change-over relays- That this is so indicates a failure to appreciate that if the control circuit is properly designed, the aerial contacts have only to pass the aerial current and not to switch it. A Post Office type 3000 (or even smaller) relay is quiet in operation and usually perfectly suitable.

The small bias supply unit may be built into the main power pack and a lead feeds through into the exciter chassis. This negative voltage not only provides bias for the driver valves but may also be used as the source of negative muting voltage.

Fig. 10.94 shows a circuit with the controlling relay mounted on the exciter, and the connections from the second pole going to a terminal block on the rear chassis apron. Any small mains transformer rated for 50 mA is suitable for T1 and if there is a centre tapped secondary (120-0-120 volts or thereabouts) two rectifiers only can be used instead of the four shown.

It will be noted that the negative output of 100 to 125 volts is fed simultaneously to both the receiver and the exciter muting rails. However, the receiver is operative because the change-over relay contact is short-circuiting the negative voltage down to earth. When the press-to-talk button is depressed, h.t. current flows through the 47 K ohms resistor and energizes the relay winding (this may be any suitable value from 3 K to 12 K ohms). The change-over contact moves over and short-circuits the exciter muting voltage to earth. At the same time the aerial contact closes and completes the circuit feeding the 12 volt energizing current to the aerial change-over relay.

Points of this system to note are:

(i) For a moment of time while the relay pole is in a mid-position, *both* exciter and receiver are muted. Because of this, the receiver goes " dead " before the transmitter comes on and there is no thump from the loudspeaker.

(ii) After a period of transmission the exciter is muted *before* the receiver can recover and there cannot be any " howl back."

(iii) The exciter is muted and there cannot be any transmission of r.f. from the power amplifier while the aerial relay is changing over. For this reason, the relay contacts do not switch, but have only to carry, the aerial current. The maximum voltage appearing across the open receiver contacts is equal to the square root of the product of the power output and the aerial impedance.

The aerial relay is a Post Office type 600 with a low resistance coil and is energized with a simple 12 volt supply made up with a battery charger type metal rectifier and a suitable low voltage mains transformer. A silicon diode could of course be used in a simple half-wave circuit.

The most suitable relay is a two pole change-over type. One pole changes over the aerial and the other pole is arranged to short-circuit the receiver input during transmission periods as a protection for the receiver input circuits. This is shown in detail in the circuit of **Fig. 10.95**.

A metal chassis is a wonderful sounding board, and any relay—even the miniature type—will make a noise if it is bolted directly to a sheet of metal. The best method is to isolate it from direct contact by means of a flexible rubber mounting. The 12 volt supply, the co-axial input sockets an r.f. ammeter if required, and the relay for the aerial system are mounted in a small aluminium box about 8 in. × 6 in. × 2½ in. deep.

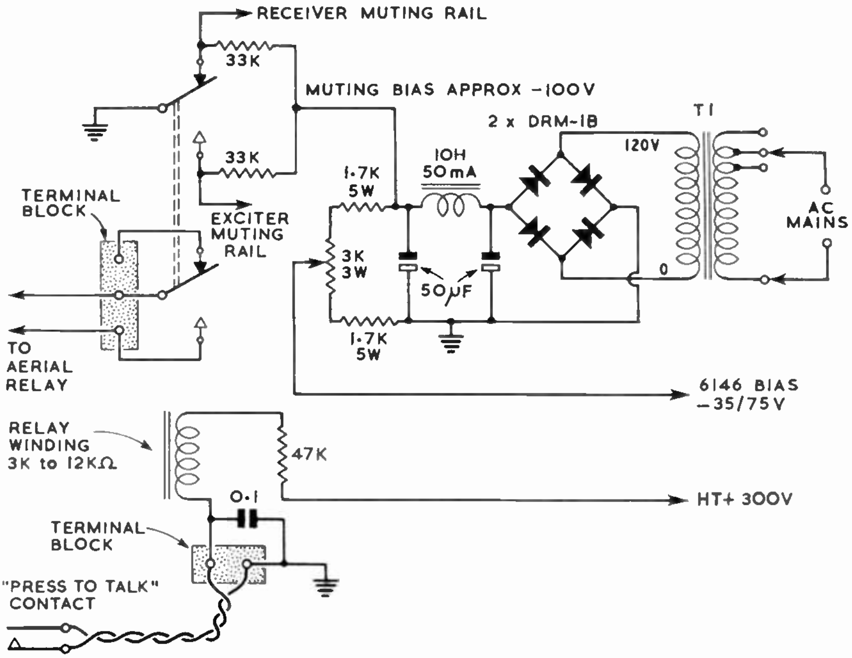


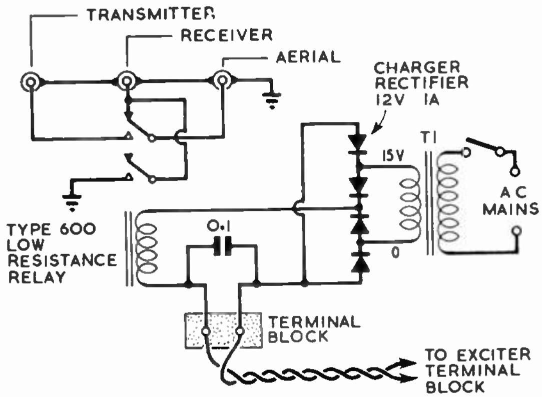
Fig. 10.94. Transmitter and receiver control system. The bias supply is in the main exciter power pack, with the 3 K ohm potentiometer and the relay on the exciter chassis. The relay is a high resistance two pole change-over type.

AN ELECTRONIC VOICE CONTROL UNIT

A voice-controlled switch is, as its name suggests, a device which will switch on some associated equipment, such as a transmitter, when a voice signal is present. The usual method of achieving this is to rectify the applied audio signal and use it to control the bias on the grid of a valve which has in its anode circuit a sensitive relay connected to the associated equipment to perform the switching.

A time delay must be incorporated so that the relay does not respond to each individual syllable of the speech but only to the presence, or absence for a specified length of time, of the input signal. The delay is generally arranged by means of a capacitor in the grid circuit which is charged by the rectified audio input and which discharges through a resistor when the input is removed. A disadvantage is that the time delay is dependent upon the voltage to which the capacitor has charged. If the last sound input was of low intensity, then the delay will be short and if of high intensity, the delay will be long. With this type of circuit, a sensitive relay is required and so precautions have to be taken to ensure that noise from the relay is not fed back through the system as this will cause instability.

Voice control circuits which do not employ relays are also in use. In these a valve is connected between earth and a negative voltage rather than a positive voltage and earth. A negative bias voltage can therefore be obtained to switch an external circuit without employing relays. Since *p-n-p* transistors require negative voltages for their normal operation, their use in a circuit of this nature should be advantageous.



10.95. Aerial change-over system. T1 may be any transformer that will give about 15 volts output. The relay is energized for transmit—this enables the receiver to be used when the transmitter is off or disconnected. The relay is a two pole change-over type.

To assess this advantage it is necessary to compare the large signal characteristics of valves and transistors. Considering the circuit shown in Fig. 10.96 with the valve anode

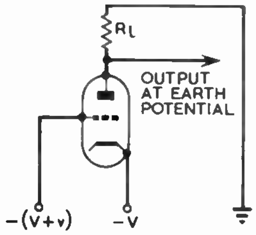


Fig. 10.96. Basic valve circuit −(V + v) greater than −V.

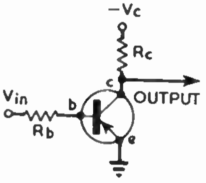


Fig. 10.97. Basic transistor circuit. (c, collector; b, base; e, emitter).

connected to earth via a resistive load and the cathode returned to a negative potential (*V*) then for a value of grid voltage more negative than the cathode potential, no anode current will flow. Since no anode current is flowing there can be no voltage drop in the anode load and so the anode is at earth potential. As the grid is brought to the same potential as the cathode, anode current will flow and the anode voltage will fall towards -*V* until the valve bottoms or the product of anode voltage and current is limited by the anode dissipation of the valve. At this point there will still be a large voltage across the valve as it has quite a high impedance. This voltage has to be added to that required to switch the external circuit when the h.t. voltage is being determined. Since a negative output voltage is required, the valve has to operate with the cathode at a high negative voltage and this may lead to a breakdown of the heater-cathode insulation.

With a *p-n-p* transistor in the grounded emitter configuration (Fig. 10.97) the situation is rather different. When a slight positive bias is applied via the transistor (Rb) to the base, the transistor is cut off. The cut-off is not complete: some leakage current (of the order of micro-amps) still flows. The collector voltage is therefore not quite equal to the supply voltage but with a low value of collector load (Rc) the difference is only fractional and can be neglected. If the input voltage is now made steadily negative, base current will flow and by transistor action the collector current will increase. In consequence the voltage drop across the collector load will increase and the collector voltage will move towards earth. A certain value of base current will be reached beyond which there will be no change of collector voltage as nearly all the available supply voltage will be dropped across the collector load. The rest is a small potential difference of the order of 0·2 volts across the transistor itself from collector to emitter. The transistor in this condition is termed "on," "saturated," or "bottomed" and when only leakage current flows it is termed "off."

The total value of current possible in the collector circuit depends upon the input current and the current gain in grounded emitter (α' or β) of the transistor. Owing to the wide variation in transistor characteristics the circuit must be designed taking a minimum value for α' to ensure that the required current is obtained with all transistors. No damage is done if the α' is higher than the value assumed, since the collector load will limit the current to the design value.

The advantages of the transistor for this application, apart from the suitability of supplies mentioned originally, are therefore:

(*a*) Full use of the available voltage, i.e. the output can swing from −0·2 volt (transistor bottomed) up to the maximum supply voltage.

(*b*) Low power dissipation in both states: OFF, high voltage, low current; ON, low voltage, high current.

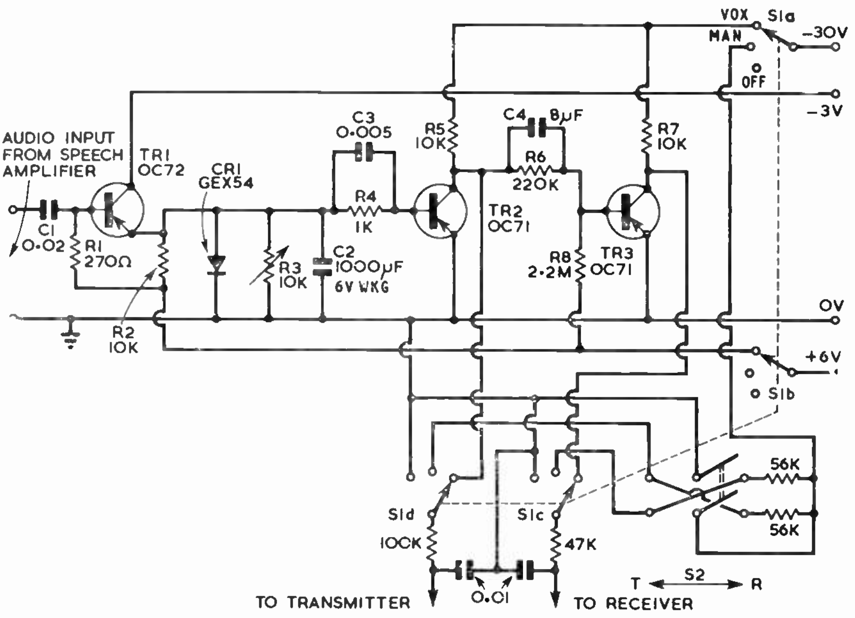


Fig. 10.98. Complete circuit diagram. The outputs to the transmitter and receiver are connected to the earthy end of the grid circuit of the controlled stages. The audio input from the speech amplifier should be of the order of 3 volts.

(Maximum dissipation occurs at half the voltage swing when half the maximum current is flowing.)

(c) No heater-cathode insulation worries.

With these points in mind, the circuit of **Fig. 10.98** was developed. The maximum voltage the transistors can safely stand is −30 volts and so this was used, it being a convenient value also for biasing off valves in receivers and low level transmitter stages. The values of collector resistors were chosen to ensure bottoming at low current levels and so reduce the drain on the battery. The necessary delay is effected by the large capacitor C2 connected from the base circuit of TR2 to earth. The value must be high because of the low resistance discharge path through the base-emitter junction.

In order that the second stage may be quickly driven into saturation, the input source must supply sufficient current to charge the capacitor to the required voltage in a very short time. Since some form of isolating circuit is also required to reduce the loading effect of the large capacitance on the preceding speech amplifier, an emitter follower is used. This stage is biased in such a manner as to rectify the input speech waveform and so fulfils three functions. An OC72 is employed in this position since a high value of α' is required.

The operation of the circuit is as follows.

With no input signal to the emitter follower (TR1), this transistor is cut off due to the positive bias on its base, with respect to the emitter, and so the only current that flows is from the +6 volt supply through the resistor R2 and the diode CR1 to earth. This holds the emitter of TR1, the capacitor C2 and hence the base of TR2 at a very slight positive potential and so ensures that TR2 is cut off. At the same time TR3 conducts, being supplied with base current from the −30 volt supply. When an alternating signal is applied to the input, the emitter of TR1 follows the negative excursions, the positive ones being suppressed owing to the biasing. As the emitter goes negative, the diode CR1 is reverse biased and so can be considered out of circuit, leaving C2 to be charged to the peak d.c. value of the emitter voltage.

The second stage is arranged to saturate when the capacitor C2 is charged to −0·5 volts and so its collector will remain at earth potential while the capacitor voltage is more negative than this value. The capacitor C3 in parallel with the base resistor of this stage helps to reduce the response time of the transistor since, for rapid changes of input, it is virtually a short circuit. The steady state conditions are maintained by the value of the base resistor. Owing to the smoothing effect of the capacitor C2 and the low impedance discharge path through the transistor, it will be found that the emitter voltage of TR1 is almost constant while an input is applied, such variations as are present being small compared with the target voltage of +6 volts. In this way, the time delay can be made largely independent of the input intensity. The discharge of C2 controls the delay before the circuit returns to its original condition.

The discharge rate is, of course, exponential and depends upon the values of R2, C2, the voltage to which C2 is charged and the target voltage (+6 volts in this case). This rate of discharge will be high initially and this is the characteristic required to ensure that TR2 and TR3 pass quickly from one condition to the other once it is obvious no further input is to be applied. However, the nearer the potential of C2 becomes to the target potential the slower will be the discharge and this would upset the switching action. Use is therefore made of a positive target voltage (+6 volts), to give a high discharge rate, and of diode CR1. Once the potential of C2 becomes slightly positive CR1 conducts and "clamps" the potential of C2 to earth. By this means, only the first rapid portion of the discharge characteristic is employed. A parallel discharge path is provided by the variable resistor R3 to allow some variation of the time delay to suit individual requirements.

The third stage is merely an inverter to give a negative bias voltage for the receiver when the transmitter is turned on. The high value of capacitor across the base resistor is unconventional but is necessary to speed up the turn off time of the receiver. This enables the receiver to be completely muted before the transmitter is fully energized and so the operation is smooth and quiet. Another approach to this would be to slow down the turn on time of the transmitter but this would tend to produce excessive clipping of the first syllable and so reduce the efficiency of the device. Owing to its recovery time, the receiver does not regain full sensitivity until the transmitter is biased off again and so no noise is generated in the change-over.

A further difficulty sometimes encountered when voice control circuits are used with a loudspeaker is the tendency for sounds from the loudspeaker to switch the circuit over. A slow form of oscillation can then develop since when the receiver goes off the sound from the loudspeaker which initiated the switching also disappears. After a length of time governed by the delay components, the receiver will come on again, the sound from the loudspeaker will once more trip the circuit and the cycle will be repeated.

In this instance, the gain of the speech amplifier feeding the voice control unit has been held to the minimum necessary for reliable operation when speaking about two inches away from the microphone. The loudspeaker is placed some 2 ft. away and so does not trip the voice control circuit. As a result no anti-trip circuit is included. If tripping does occur, it is possible to feed a voltage from the receiver so that it cancels the voltage produced by sounds from the loudspeaker which are picked up by the microphone.

Construction

The form of construction is not important, although the original unit was built mainly on a tag-strip. No difficulties should be encountered in this respect, the only point to note being that the bias line decoupling components should be placed as close as possible to the controlled stages in the transmitter and receiver.

The 30 volt battery is a small deaf-aid type, while the −3 volt and +6 volt supplies are obtained from a 9 volt grid-bias battery.

Conclusions

On-the-air tests have shown that clipping of the first syllable is negligible even for words which start with a "soft" sound such as electron or microphone. The time delay before the circuit switches to receive can be made short, so that a normal pause for breath will allow it to

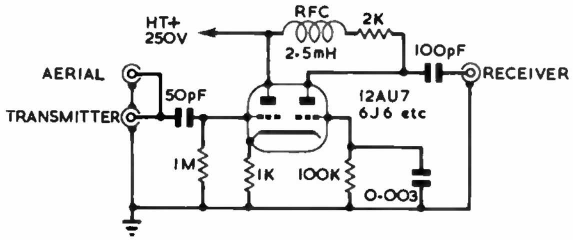


Fig. 10.99. Low impedance electronic aerial switching using a double-triode.

switch, or long so that a more definite pause is necessary. Once set, this delay is virtually independent of the intensity of the last sound.

The current drain is nearly constant at 4 mA and so small batteries can be used to power the device. This drain could be further reduced by using, with suitable circuit modifications, an *n-p-n* transistor for TR3, the inverter. The drain would then be about 1 mA on standby and 7 mA when energized.

Suitable switching is included to provide manual or VOX operation, as well as to remove the voltage supplies when the unit is out of use.

Electronic Aerial Switching

An electronic voice control unit will also necessitate the use of an electronic TR switch. This may be the low impedance type interposed between the aerial co-axial cable and the transmitter 75 ohms output socket or, alternatively, the high impedance type built as a unit adjacent to the power amplifier valves and associated pi tank coil.

The low impedance type is shown in the circuit of **Fig. 10.99.** The first half of the valve behaves as a cathode follower, coupling into the second half arranged as a grounded grid amplifier. On transmit, the heavy grid current biases the input section to cut-off and the receiver input circuits are isolated from the transmitter r.f. output voltage.

A suitable high impedance type is shown in the circuit of **Fig. 10.100.** This should be built as a unit and mounted on the p.a. chassis in order to keep the r.f. input connection reasonably short.

The anode winding is 60 turns of 38 s.w.g. enamelled wire, wound round a dust core, with the winding tapped

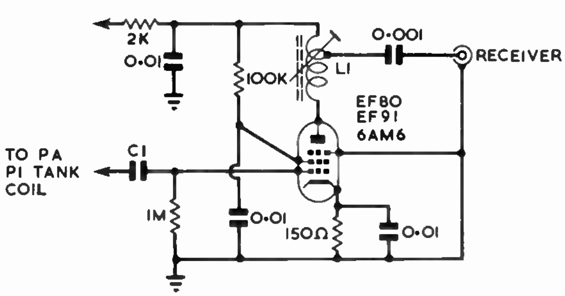


Fig. 10.100. High impedance circuit for electronic aerial switching.

eight turns from the h.t. end. Any common r.f. pentode such as the EF91 (6AM6) or EF80 is suitable. During reception the valve behaves as a low gain r.f. amplifier fed from the pi tank which should be resonant at the receiving frequency. During transmission periods the valve takes heavy grid current and is biased to virtual cut-off, thus isolating the receiver input from the transmitter r.f. output.

It should be borne in mind that electronic TR switches can be possible sources of TVI.

Control systems are dealt with in greater detail in Chapter 8 (*Keying and Break-in*).

FILTER TYPE TRANSMITTERS

It has already been shown that the requirement of all-band coverage materially increases the complexity of design. A decision to restrict operation to a limited number of the h.f. amateur bands can bring about a worthwhile gain in circuit simplicity. The newcomer to signal sideband who is embarking on the construction of a s.s.b. transmitter for the first time may well be very willing to sacrifice two of the lesser used amateur bands in return for the advantages of simple construction and straightforward alignment.

A FOUR BAND TRANSMITTER

The single sideband transmitter shown in **Fig. 10.101** may be used directly into an aerial system or alternatively as an exciter to drive a higher powered final amplifier. It has been designed to provide a high quality signal on the four most used amateur bands of 80, 20, 15 and 10m.

Initial sideband generation from a mechanical filter on a nominal frequency of 455 kc/s is heterodyned by the v.f.o. tuning 3·955 to 4·455 Mc/s to give a tunable s.s.b. output in the 80m band from 3·5 to 4 Mc/s. For 80m operation this output is switched straight through to the grid of the EF80 voltage amplifier valve V9. For the three higher frequency bands the 80m output from T4 is switched to one grid of the 12AT7 mixer valve V7. The signal is then heterodyned by the output of the switched crystal controlled overtone oscillator V8 to give the required output on 10, 15 or 20m.

Coil winding has been reduced to the minimum by making the three variable capacitors VC2, VC3 and VC4 tuning the signal frequency stages large enough in value to give the necessary 4 : 1 capacity swing to enable the 14, 21 and 28 Mc/s ranges to be resonated using one coil only in each anode circuit of V7, V9 and V10. An additional coil L4 tunes the 80m band output.

The audio stages V1 (EF86) and the first half of V2 (12AU7) are a straightforward voltage amplifier. The second half of V2 is connected as a cathode follower to present the correct impedance to the OA79 diode modulator. Output from the carrier oscillator V3 is fed via a step-down winding on T1 to the OA79 diodes and the carrier balancing potentiometer VR2. Oscillator output is also fed to VR3, the CARRIER INSERTION control, so that an r.f. signal at the carrier frequency can be fed round the sideband filter and used for carrier insertion or netting purposes.

T2 together with the 75 pF and 1000 pF resonating capacitors forms an impedance step-up transformer to provide the correct matching to the Kokusai mechanical filter. The filter is required to provide a high sideband output (i.e. the carrier

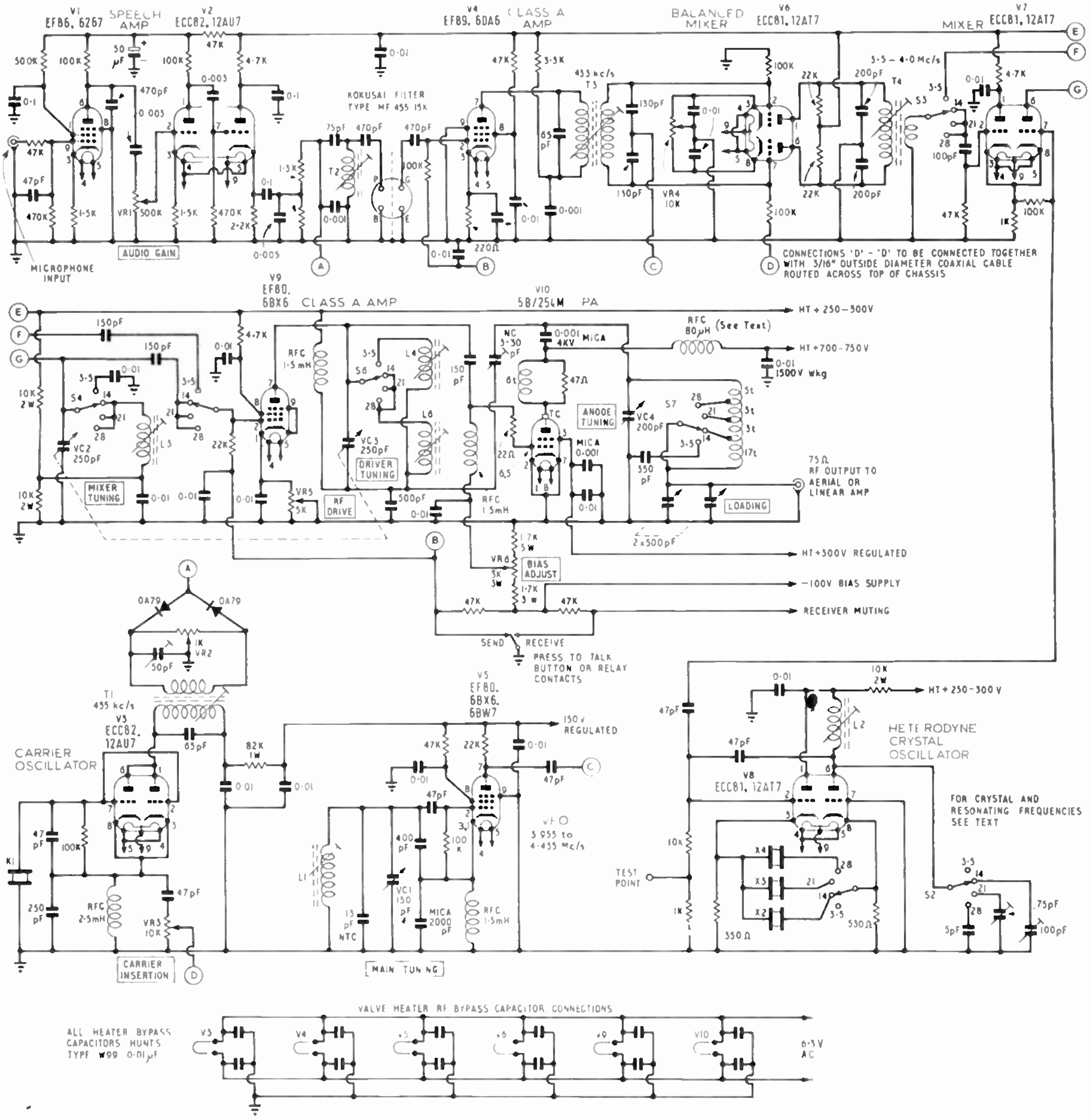


Fig. 10.101. High quality filter type transmitter or exciter for 80, 20, 15 and 10m, using mechanical filter. L2, L3, L6, 8 turns 20 s.w.g enam. spaced one wire diameter; L4, 30 turns 28 s.w.g enam., all wound on Aladdin 804 ($\frac{7}{16}$ in. diameter) formers. For details of remaining coils and transformers, see text.

crystal must be on the *low* frequency side of the filter response) because the required low sideband output in the 80m band is the low sideband of the frequency translation process and this produces a frequency inversion (i.e. the heterodyning input from the v.f.o. is *higher* than the wanted output of 3·5–4·0 Mc/s. This changes a high sideband signal to the mixer into a low sideband signal out of the mixer.)

The low level s.s.b. output from the filter is amplified by V4 operating in class A, and fed into the 455 kc/s r.f. transformer T3 feeding the grids of the balanced mixer V6 in push-pull; this valve also receives the heterodyning input from the v.f.o., V5, in parallel and this enables the mixer valve to be balanced by adjustment to VR4. With the BAND switch S1 to S6 in the first position, 3·5–4·0 Mc/s output is connected directly into the grid of the class A amplifier valve V9. In the other three positions the 80m signal from T4 is connected to the mixer valve V7 where it is heterodyned by the output of the overtone oscillator V8. The correct crystal for each band and the resonating capacitor to tune the anode coil L2 are selected by the switch banks S1 and S2. V9 is a conventional class A voltage amplifier with a tuned anode circuit coupled into the grid of the output valve V10.

The 5B/254M is the modern version of the 807 valve and will run to an input of approximately 90 watts p.e.p. with 750 volts on the anode and 300 volts regulated on the screen —bias should be adjusted for a zero signal anode current of 30 mA with the bias setting potentiometer VR6. A 6146 valve would be equally suitable and would run to the same 90 watts p.e.p. input but would require the screen voltage reducing to 200 volts regulated—a pair of VR105 regulators would be quite suitable.

The negative 100 volts bias supply also provides the potential, via the 47 K ohm feed resistors, for transmitter and, if required, receiver muting. When the press-to-talk button or relay is in the transmit position, the cut-off bias on V4 and V9 is shorted down to earth; these two valves then conduct in the normal manner. In the receive position V4 and V9 cease to conduct making the transmitter inoperative; at the same time the receiver muting line is shorted to earth allowing the receiver to commence operation.

Metering is not shown on the circuit diagram, but it is advisable to have a meter permanently wired in the 750 volt line to the p.a. anode (150 or 200 mA movement) and very useful to have a meter (0–25 mA) in the screen feed as well.

Construction

A suitable and easily worked material for the chassis is 18 s.w.g. aluminium. This may be obtained ready made in 2½ in. deep box sections, made to standardized dimensions. Three or four " boxes " may be placed side by side and bolted together to give the required surface area to accommodate all necessary valves and components. Three 12 in. × 5 in. × 2½ in. box sections placed side by side would make up a chassis of 15 in. × 12 in. × 2½ in. and this would give plenty of room. Alternatively a box section of 16 in. × 10 in. × 2½ in. could be divided into three sections by making up and fitting two cross screens 10 in. long, spaced 2½ or 3 in. either side of centre.

It is suggested that V1, V2, V3, V4, V6 and associated components are in section " A " (left hand); V5, V7, V8, V9 and associated components including switch banks S1 to S6 in section " B " (centre); V10 and tank circuits and switch S7, VR6 etc. in section " C " (right hand). Stages handling the same frequency must never be able to " look at each other "—this applies to the switch banks and coils of V7 and V9—and it will be necessary to fit two cross screens so that the 5 in. wide section " B " is further divided into three compartments. These screens also afford necessary support to the switch banks. Inspection of the chassis layout for the " G2DAF Transmitter " described later will prove to be helpful and illustrate the basic idea of using separate box sections to make up a complete chassis and give the required screening.

Before starting to drill any holes get a sheet of white drawing paper and mark out a full size outline including all the underneath screens. (Mark out a second rectangle to dimensions of the front panel.) Lay out on the drawing paper all the major components and controls, keeping a careful eye on the location of the switch shaft that will run almost the full length of the centre section, and the under-chassis partition screens, the accessibility of the valveholders and the associated components grouped round the valveholder, and the fact that connecting wires carrying r.f. must be as short as possible.

Ensure that the coils are positioned against their respective switch banks to keep connections short, at the same time making sure that the coil centres are at least 1 in. away from any underneath metalwork. Also check that the various control shafts will come out in the right place to give a balanced panel layout. The more time and thought given to this initial setting out the less likelihood of trouble later on. Finally, draw round the components with a finely sharpened pencil, removing each item and drawing in to accurate dimensions including mounting holes and drilling centres for the valveholder chassis punch. The final drawing can then be laid on top of the chassis or the panel and used as a template for marking out.

Standard Yaxley wafer switches (without shorting plates), either ceramic or paxolin, are suitable for S1, S2, S3, S4, S5 and S6 and these are ganged together on one shaft. S7 handles a considerable amount of r.f. power and should preferably be ceramic.

The i.f. transformers T1, T2 and T3 are standard Maxi-Q Type IFT.11-465 modified as follows: for the step-down transformer T1, the original top pie winding is removed together with the 65 pF resonating capacitor; 75 turns of 32 s.w.g. enamelled wire is then scramble wound close against the existing primary and soldered to the original secondary leading out tags. Filter coupling transformer T2 is one winding of a standard i.f.t. from which the original internal 65 pF resonating capacitor has been removed; the new 75 pF and 1000 pF resonating capacitors shown on the circuit diagram are wired externally, underneath the chassis. The unused secondary is removed by cutting away. T3 only requires modification to the 65 pF secondary resonating capacitor, which is removed by snipping through with cutting pliers; the two 130 pF series capacitors forming the centre-tap for the v.f.o. injection are wired to the lead out tags, underneath the chassis. Transformer T4 is wound on a standard Aladdin coil former Type 804 ($\frac{7}{16}$ in. diameter with a dust slug). The primary winding comprises 48 turns close wound of 34 s.w.g. enamelled wire with one layer of typewriter copy paper insulation, and the secondary of 24 turns of 34 s.w.g. enamelled wire overwound over the centre of the primary winding. L2, L3, L4 and L6 are also wound on the Aladdin Type 804 former, the winding details being given in the caption to Fig. 10.101.

V5 is arranged as a Colpitts v.f.o. circuit which has a high degree of frequency stability, a small warm up drift, almost constant amplitude of output voltage across the tuning range of 500 kc/s, and because of the high effective value of C in the tank circuit, a very low harmonic output. The frequency is controlled by the tuned circuit, so the stability cannot be any better than the stability of the coil and capacitor forming this circuit. As the temperature coefficient of copper wire coils and brass or aluminium plate tuning capacitor are outside the control of the amateur, the secret of success is in the mechanical stability of the assembly as a whole, the care with which the inductor L1 is constructed and finding the right value of negative temperature co-efficient compensating capacitance across this coil. The coil former should be ceramic, with a dust core mounted on a screwed brass rod fitting into a pressure loaded clutch so that there is neither end or side float of the core within the winding. Highly recommended is the ⅜ in. diameter Cambion former Type 1534-2 with 20063-B slug. The winding of 22 turns of 24 s.w.g. enamelled wire should be close wound under tension, and finally thoroughly doped with polystyrene cement. The tuning capacitor VC1 should also be of sturdy mechanical construction with a two bearing rotor; the "law" of the capacitor (i.e. "mid log," "straight line frequency," "straight line capacity") affects the linearity of the dial calibration. Rather conveniently the physically compact capacitor with semi-circular rotor plates when used in a circuit with relatively large values of shunt capacity gives a dial calibration that has good linearity across a 500 kc/s tuning range. An excellent variable capacitor suitable for the requirement is the Polar C28-142. 18/015, a two gang unit of 75 pF each section. The two sections are wired in parallel to give the required maximum 150 pF. Also associated with the tuned circuit are the 400 pF and 2000 pF fixed capacitors and these must be the best possible quality *silver mica* types, as must be the 47 pF grid capacitor. In regard to the negative temperature co-efficient capacitor across the coil, this is an Erie Type N750 K; the value shown (15 pF) is purely a nominal one and may have to be decreased or increased in value to obtain the best frequency stability—this can only be found experimentally.

The anti-parasitic choke in the anode circuit of the 5B/254M valve is made in the usual way by space winding six turns of 18 s.w.g. enamelled wire over the body of a 47 ohm ½ watt carbon resistor. The anode tuning capacitor may be the semi-circular rotor, such as the Jackson Type C604—this has an air gap of 0·019 in. The aerial loading capacitor is a standard broadcast type of 500 pF each section.

Should difficulty be experienced in obtaining a suitable 1½ in. diameter ceramic former for the pi-tank coil, this may be hand wound over four 2½ in. lengths of ¼ in. diameter polystyrene rod held in position on a mandrel made from a 4 in. length of 1½ in. diameter birch dowel rod previously prepared with four semi-circular grooves down its length. The mandrel is supported horizontally and turned as the wire is run into position; as it crosses each polystyrene rod the wire is "cemented" into position by local heat applied with a soldering iron. After the winding is completed the mandrel is withdrawn and the coil carefully inspected to ensure that there are no shorted turns. The coil is finally supported on the pi tank switch S7—this can conveniently be made up as a unit before final assembly on the chassis. Suitable coils are available ready wound from Codar and other manufacturers.

FINAL CONVERSION OSCILLATOR FREQUENCIES

Band	Crystal Frequency	Mode	Output Frequency
80m	Nil	Nil	Nil
20m	6·0 Mc/s	series-resonant	18·0 Mc/s
15m	8·333 Mc/s	series-resonant	25·0 Mc/s
10m	10·8 Mc/s	series-resonant	32·4 Mc/s

PI-NETWORK DATA—ONE 5B/254M
R_L = 3500 OHMS

Band	XC_1 = 290 ohms	X_L = 330 ohms	XC_2 = 40 ohms
80m	150 pF	16 μH	1100 pF
20m	40 pF	4 μH	275 pF
15m	30 pF	3 μH	207 pF
10m	20 pF	2 μH	138 pF

Tank coil 28 turns of 18 s.w.g. tinned copper wire, one continuous winding 14 turns per inch tapped at 5, 8 and 11 turns from hot end. Total length 2 in. diameter 1½ i.d. supported on S7.

A standard pie wound r.f. choke is totally unsuitable for use with a shunt fed tank circuit and this component must be specially wound to have a low self-capacity and no self-resonant points within the required amateur bands. The required inductance value is 80 μH and may be obtained from several manufacturers. Alternatively, it can be constructed by winding 112 turns of 26 s.w.g. enamelled wound in one layer and spaced out to 3⅞ in. on a ceramic former of ¾ in. or ⅞ in. diameter. This spacing reduces the self-capacity to a value that does not produce any self resonance at the operating frequencies. (A ceramic former from an old a.c./d.c. broadcast receiver mains dropper resistor is suitable.)

Alignment

The first step is to neutralize the 5B/254M output valve. (For a detailed description and explanation of neutralizing see Chapter 6 (*H.F. Transmitters*).) When neutralization is perfect there is no r.f. coupling between the anode and the grid circuits of the valve, and any signal injected into the 75 ohm output socket will not appear in the grid coils L4 or L6. In the absence of a signal generator an "off air" transmission can be used by adopting the following procedure. Switch all coils to a convenient band—say 20m—and with heaters on but all h.t. supplies switched off, connect a receiving aerial to the 75 ohm output socket. Couple a small pick-up loop of two or three turns to L6, the loop feeding down a length of coaxial cable into the aerial input of the receiver. Tune in a strong steady signal (commercial teletype is useful) and resonate VC4 and VC3 for maximum "S" meter reading. Now adjust the 3–30 pF neutralizing capacitor for *minimum* S meter reading; re-resonate C3 and C4 and repeat the neutralizing adjustment. It should be possible to phase out the incoming signal almost completely.

Apply h.t. and with the aid of a BC221 frequency meter or a calibrated absorption wavemeter check that the v.f.o. is oscillating and tuning the required range; and that the carrier oscillator V3 is giving r.f. output into the diode modulator. Connect a 1 mA meter or a workshop voltmeter (10 volt range) to the test point of the heterodyne oscillator V8 and with the band change switch in the 10m band position adjust the core of L2 for oscillation—this is denoted by an upward kick in the meter reading. As a further check couple an absorption wavemeter to the anode coil and ensure that the valve is in fact oscillating on the correct frequency. The core of L2 must not be touched again; all further adjustment for resonance on the other two bands is made by the 75 pF trimmer for 15m and the 100 pF trimmer for 20m, the correct trimmer and crystal being automatically selected by the position of the main band change switch assembly S1 to S6.

Finally, feed a 1 kc/s or a 1·5 kc/s audio tone into the microphone socket, turn up the audio gain control, feed the output from T4 by a pick-up loop into a receiver and peak the dust cores of T2, T3 and T4 for maximum output. Disconnect the tone input, turn the audio gain and carrier injection potentiometers to zero—there will still be an output from T4 deflecting the station receiver S meter due to carrier feeding through the balanced modulator. Carefully adjust VR2 and the associated 50 pF trimmer for maximum modulator balance (i.e. the least carrier break through and the lowest S meter reading on the receiver.) Set VR6 for a zero signal standing anode current of 30 mA.

With full drive the p.a. should be loaded for a maximum signal anode current of 100 to 120 mA and a screen current of 5 or 6 mA. Check with an oscilloscope that the output waveform is " clean " and that there is no flat-topping and finally verify that the transmitter is not being inadvertently overdriven, by getting on-the-air reports from experienced s.s.b. operators.

G2DAF S.S.B. TRANSMITTER

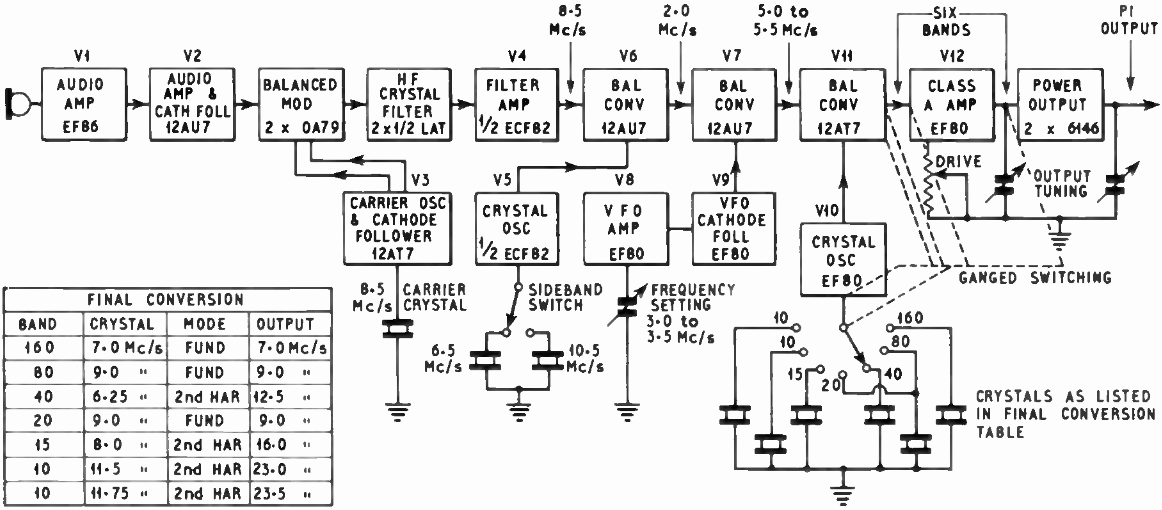
The block diagram of an s.s.b. transmitter employing a high frequency filter (8·5 Mc/s) is shown in **Fig. 10.102** and the complete circuit diagram in **Fig. 10.103.**

The audio stages, V1 (EF86, 6267) and the first half of V2 (12AU7), have cathode resistors without bypass capacitors giving negative current feedback to each amplifying stage. The second half of V2 is connected as a cathode follower to present the correct impedance to the OA79 balanced diode modulator.

Output from the carrier oscillator V3a is fed via a step-down winding on T1 to the two OA79 diodes and the carrier balancing potentiometer VR2. Cathode follower V3b feeds an r.f. output signal into VR3, the CARRIER INSERTION control, and this enables an r.f. signal at the carrier frequency to be fed round the sideband filter and used for carrier insertion or netting purposes.

The double sideband suppressed carrier output from the balanced modulator is fed into the two section crystal band-pass filter. The low level single sideband output is then amplified by V4a (the pentode section of an ECF82, 6U8) operating in class A, and fed into an r.f. transformer T3 feeding the grids of the balanced converter V6 (12AU7) in push-pull; this valve also receives the parallel heterodyning input from the sideband switching oscillator V5 (the triode section of the ECF82).

If a different carrier frequency from the one specified is used the sideband switching crystals will have to be selected accordingly, with a spacing between them of approximately twice the carrier frequency. The two preset capacitors in the oscillator grid circuit are adjusted to pull the crystal frequencies the final amount necessary to maintain zero beat when switching sidebands. If the filter gives upper sideband output, the low frequency sideband switching crystal will



FINAL CONVERSION			
BAND	CRYSTAL	MODE	OUTPUT
160	7·0 Mc/s	FUND	7·0 Mc/s
80	9·0 "	FUND	9·0 "
40	6·25 "	2nd HAR	12·5 "
20	9·0 "	FUND	9·0 "
15	8·0 "	2nd HAR	16·0 "
10	11·5 "	2nd HAR	23·0 "
10	11·75 "	2nd HAR	23·5 "

Fig. 10.102. Block diagram of the G2DAF s.s.b. transmitter using an h.f. filter with a nominal frequency of 8·5 Mc/s.

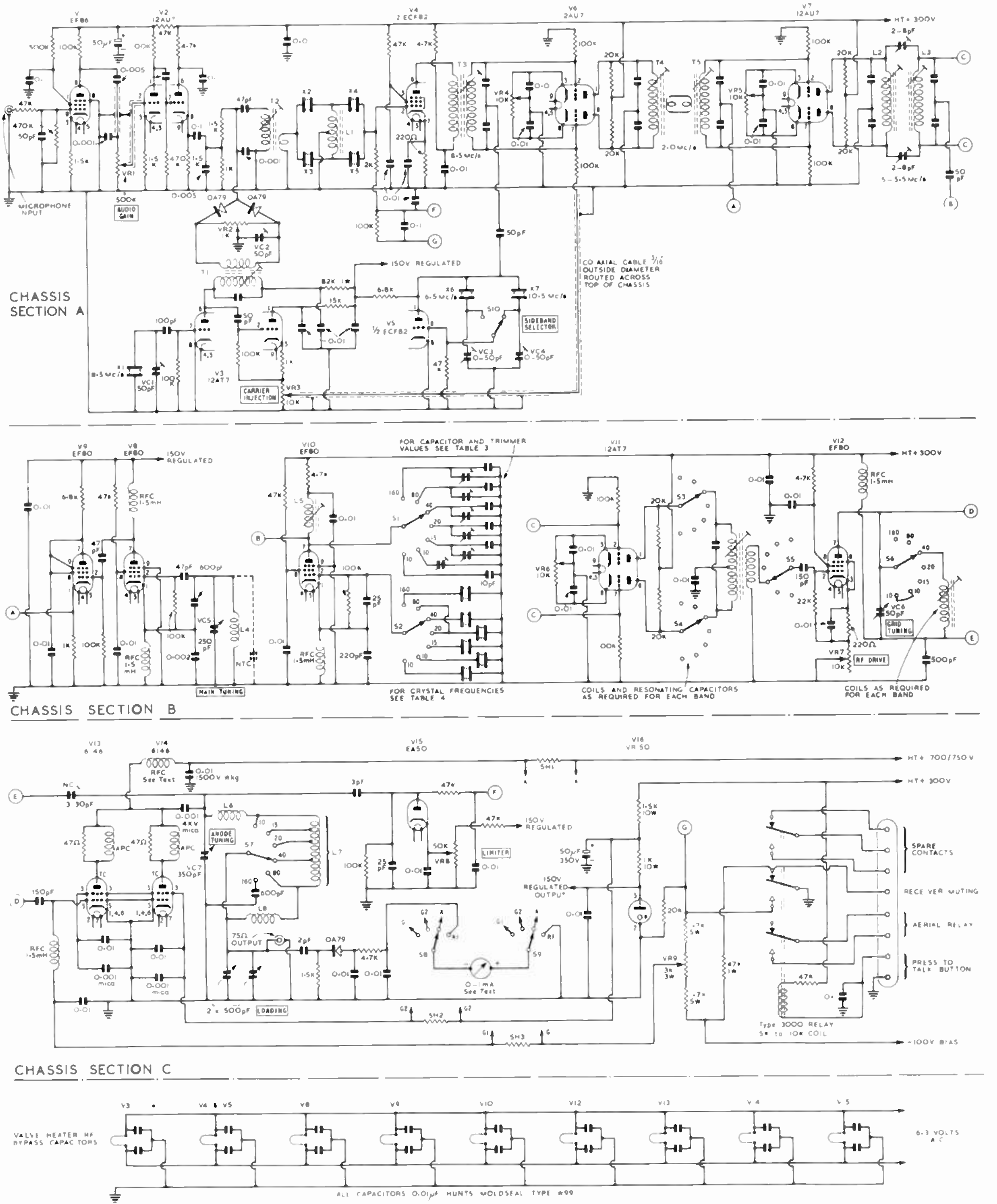


Fig. 10.103. Circuit of the G2DAF s.s.b. transmitter using the high frequency sideband generating unit. For the sake of clarity, where heater line bypass capacitors are required, these are shown separately. If required, a negative temperature coefficient capacitor, Erie type N750K, can be wired across the v.f.o. coil L4, as shown. The value will, however, have to be found experimentally, although it should lie within the range 15 to 30 pF. All band-change switches are shown in the 40m position. Values for the fixed capacitors and trimmer which form part of the resonant circuit at the anode of V10, can be found in Table 10.19. The frequencies of the crystals in the cathode circuit of V10 are given in Table 10.20. The 0·002 uF capacitor in the v.f.o. cathode circuit is silver mica. Details of the construction of L1 are given on page 10.84. The limiter (V15) provides a simple means of automatic level control.

give the required final sideband output in each band; if the filter gives lower sideband output, the high frequency sideband switching crystal will give the correct final output in each band. On the other hand, the switching can be eliminated and one crystal only used. The transmitter automatically gives the correct sideband—lower for 160, 80 and 40m, and upper for 20, 15 and 10m. Switching is included only as an operating convenience.

After the first conversion process the output at the first intermediate frequency of approximately 2·0 Mc/s is coupled via T4 and T5 and then combined with the input from the v.f.o. and cathode follower, V8 (EF80 or 6BX6) and V9 (EF80 triode connected) tunable over the range 3·0 to 3·5 Mc/s in the balanced converter V7 (12AU7) to give an output into the wideband coupler L2 and L3, at the second intermediate frequency of 5·0 to 5·5 Mc/s.

This tunable sideband output feeds into the balanced converter V11 (12AT7) which receives the heterodyning input from V10 (EF80 or 6BX6), a fundamental or second harmonic crystal oscillator. The anode load is an inductance, L5, and this is resonated to the required heterodyning frequency for each of the amateur bands by switching various values of shunt capacity by means of S1—part of the main band change switch assembly.

After frequency conversion, the switch banks S3 and S4 connect the anodes to the pre-tuned primary of the required r.f. transformer for each band, while S5 selects the secondary winding. The sideband voltage developed is amplified by V12 (EF80 or 6BX6), a class A linear amplifier the gain of which is controlled by r.f. negative current feedback across VR7, the R.F. DRIVE control. The anode coil for V12 is selected by S6 and drive fed into the grids of the parallel 6146 or QV06-20 class AB1 linear amplifier valves; the circuit is tuned by its associated variable capacitor brought out to the panel control GRID TUNING.

The p.a. valves are correctly loaded and matched to the 75 ohm output by a conventional pi-tank circuit controlled by S7 and the ANODE TUNING and LOADING variable capacitors. A pre-determined proportion of the r.f. voltage at the anodes of the 6146 valves is taken off via the 3 pF and 25 pF capacity potential divider, rectified by the diode valve V15 (EA50 or other small diode) and fed as controlling bias to the grid return circuit of the filter amplifier V4 to provide ALC. Adjustment of the potentiometer, LIMITER control, VR8, alters the delay bias on V15, it is set to allow the diode to conduct and provide a negative bias voltage that will reduce the gain of V4 and prevent the p.a. valves being inadvertently over-driven on speech peaks.

Voltage regulator V16 (VR150, OD3 or 150C3), together with the 1·5 K ohm and 1 K ohm bleeder resistors, provides 200 volts for the screens of the 6146 valves and 150 volts stabilized for the v.f.o. and cathode follower, carrier oscillator, and sideband switching oscillator.

All operating functions are effected by press-button control of the three-pole two-way Type 3000 relay the high resistance coil of which is energized by current bled from the main 300 volt h.t. rail. Although a Type 3000 relay is specified, any other type providing the same facilities and having a coil in the 5–10 K ohms range may be used.

The relay is shown in the non-energized "receiving' position, the negative muting voltage to the receiver being short circuited and the full 100 volt bias being fed to the grid return of V4 and, via the bias setting potentiometer VR9, to the grids of the two 6146s; these three valves are therefore held at cut-off and there is no output from the transmitter. When the " press to talk " button is depressed the relay contacts close allowing the 100 volt muting bias to cut off the receiver, and at the same moment of time to short circuit the bias rail " G " to earth, allowing V4 to conduct. The bias on the p.a. valves is now a proportion of the negative 100 volt supply, determined by the setting of the 3 K ohm potentiometer VR9 which is now the centre part of a potential divider between the 100 volt bias rail and chassis earth. VR9 is adjusted until the two 6146 valves are taking a total of 50 mA standing anode current.

The panel meter (a basic 1 or 2 mA movement) is re-scaled to read 0–3 mA, 0–30 mA and 0–300 mA and is connected via the two bank, single pole four-way switch, S8 and S9, to read grid, screen and anode current of the p.a. valves. In the fourth position it is used to monitor the transmitter r.f. output voltage.

Mechanical Filter Sideband Generator

If desired, the s.s.b. signal can be generated at m.f. using one of the readily available mechanical filters, such as those manufactured by Kokusai. These filters are available on a nominal centre passband frequency of 455 kc/s with either 2·0 kc/s or 3·0 kc/s bandwidth at the 6db points. Either filter would be entirely satisfactory for an s.s.b. transmitter; however the wider 3·0 kc/s filter will give more natural audio quality, make the positioning of the carrier frequency less critical, and more accommodating in regard to the operator's voice characteristics.

The circuit diagram of the complete low frequency sideband generating unit is shown in **Fig. 10.104.** It will be seen that V1, V2, V6 and V7 and the associated components are identical to those in the h.f. filter sideband generating section in the main circuit diagram in Fig. 10.101. Changes necessary to accommodate the low frequency mechanical filter are concerned with V3, V4, V5 and their associated components. The circuit is drawn in full to show exactly what is contained within the complete chassis section " A." As a further guide the chassis layout is given in **Fig. 10.105(b).**

In order to avoid any possibility of carrier leakage past the balanced modulator the customary cathode follower is omitted. The two sections of V3 are strapped together as a single triode and carrier insertion is controlled by VR3, taking a small proportion of the r.f. cathode voltage via the 50 pF coupling capacitor. Correct impedance matching between the diode modulator and the Kokusai filter is obtained by the series resonated circuit T2. This is one winding of a standard Maxi-Q intermediate frequency transformer Type IFT.11-465 (manufactured by Denco (Clacton) Ltd., 357–9 Old Road, Clacton-on-Sea, Essex) that has been modified by removing the two existing 65 pF shunt capacitors and cutting away the unwanted pie winding.

The low level output from the mechanical filter is amplified by the class A low-mu pentode V4 (EF89 or 6DA6). In order to ensure the highest possible stability this stage is bridge neutralized by the 0·001 and 0·01 μF capacitor bridge in the anode and screen return circuits.

It will be appreciated that if the heterodyning input to the converter V6 (12AU7) is made switchable with one crystal 2000 kc/s less 453 kc/s and the second crystal 2000 kc/s plus 453 kc/s, switching crystals will change the sideband output without changing the (suppressed carrier) output frequency. The required crystal frequencies are X2 1547 kc/s and X3 2453 kc/s and these are arranged as two separate Pierce oscillators using a double triode valve V5 (12AT7). To compensate for manufacturing tolerances and the effect of stray circuit capacity the crystals are finally "pulled" exactly on frequency by the pre-set 50 pF capacitors across each grid circuit to maintain zero beat when switching sidebands. The advantage of the double Pierce oscillator is that either oscillator can be selected by switching the cathodes at d.c. and this in turn enables the switch S10 to be mounted directly on the panel.

The frequencies shown for X1, X2 and X3 are those actually used in the prototype unit. It is important to appreciate however, that the Kokusai filter has a centre passband frequency of 455 kc/s plus or minus 800 c/s to allow for manufacturing tolerances. For this reason the correct carrier crystal frequencies for each filter are individually obtained from the characteristics. As the G2DAF transmitter incorporates sideband switching it does not matter whether the filter passes the upper sideband or the lower sideband and either one of the two specified carrier crystals can be used. Obviously the frequencies of the sideband switching crystals are directly determined by the frequency of the carrier crystal actually used. For any carrier frequency of value Y, the crystal X2 = 2000 kc/s minus Y, and the crystal X3 = 2000 kc/s plus Y.

T1 in the anode circuit of the crystal oscillator is required to have an impedance step down to the diode modulator. The upper pie winding of a standard Maxi-Q Type IFT.11-465 transformer is removed by cutting through with a sharp knife and replacing with 75 turns of 32 s.w.g. enamelled wire tightly coupled by scramble winding against the lower pie coil. The new winding is connected to the original lead-out wires but the original fixed tuning capacitor is removed.

In addition to the current balance obtained by the potentiometer VR2, the modulator must be reactively balanced; the amount of capacity necessary is affected by a number of variables such as out of balance in the transformer T1 stray circuit capacity and the tuning of the anode circuit. It is not therefore possible to give specific values and the fixed capacitor in parallel with VC2 may be in the range 25 to 100 pF and will have to be found experimentally. Should adjustment to VC2 fail to improve the carrier attenuation the connections from the secondary of T1 should be reversed.

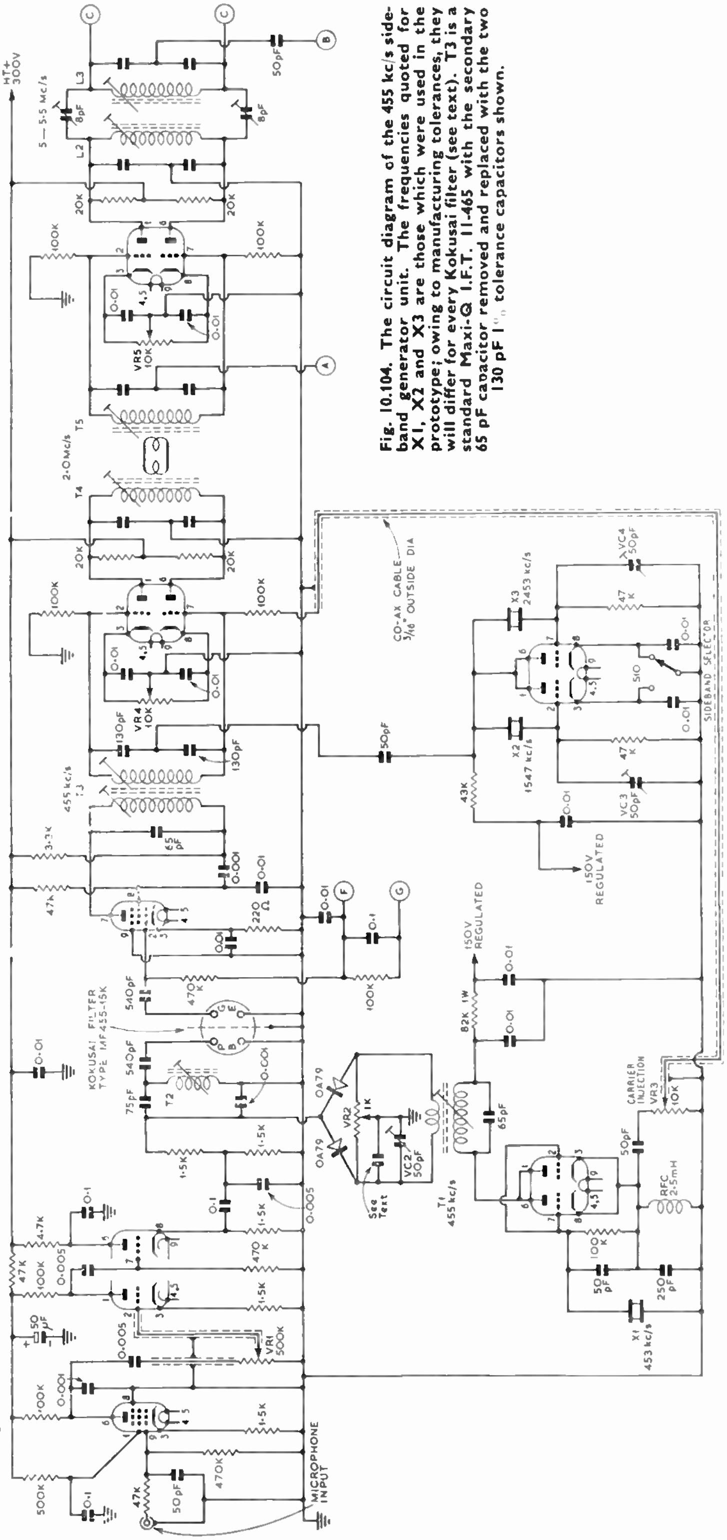
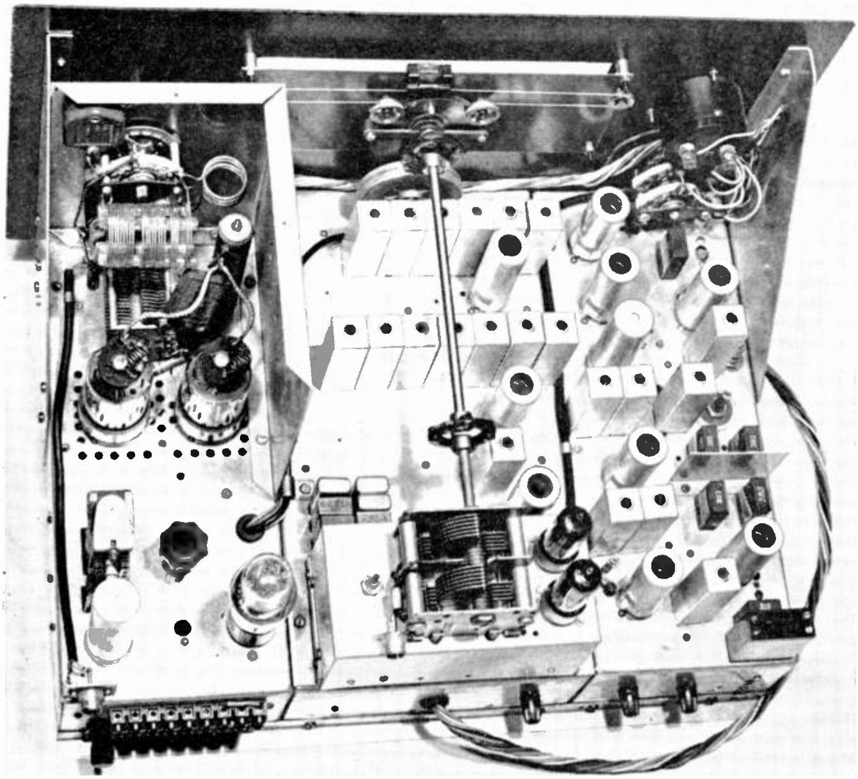


Fig. 10.104. The circuit diagram of the 455 kc/s sideband generator unit. The frequencies quoted for X1, X2 and X3 are those which were used in the prototype; owing to manufacturing tolerances, they will differ for every Kokusai filter (see text). T3 is a standard Maxi-Q I.F.T. 11-465 with the secondary 65 pF capacitor removed and replaced with the two 130 pF 1% tolerance capacitors shown.



Looking down on the G2DAF s.s.b. transmitter Mk.2 from the rear. From left to right, the linear power amplifier with pi-network tank circuit, the band pass intermediate frequency stages and the sideband generators. The components may be identified by reference to Fig. 10.105.

Component Considerations

The efficiency and satisfactory operation of any sideband transmitter is directly dependent on the " goodness " of the various resonant circuits. The design and the performance of the prototype may be everything that could be desired; however, this is scant compensation to the builder if his new transmitter fails to measure up to the expected standards. When this happens, and the usual checks do not show any obvious fault, the failure is usually in the tuned circuits.

For this reason all r.f. transformers and coils in this transmitter have been wound using standard easily obtainable 0·3 in. diameter formers and 0BA dust cores with 2¼ in. × $\frac{13}{16}$ in. × $\frac{13}{16}$ in. seamless aluminium cans. Suitable items are manufactured by Neosid and Aladdin. This ensures that all completed resonant circuits will give exactly the same Q value and coupling co-efficient as those used in the original transmitter.

A good quality variable capacitor should be used to tune the v.f.o. The type selected should preferably have ceramic insulation and a two bearing rotor. The capacitor used in the original transmitter was a twin gang 125 pF type, with both sections strapped in parallel to provide a maximum capacitance of 250 pF.

Balancing potentiometers VR4, VR5 and VR6 in the converter cathodes carry d.c. only and may be the small wire wound pre-set type of 10 K ohms in value and rated at ½ watt.

As the 1 K ohm carrier balancing potentiometer and the 10 K ohm carrier insertion and r.f. drive potentiometers are carrying r.f., these must be the non-inductive type with carbon tracks.

The germanium diodes in the balanced modulator should be a matched pair.

With the exception of the three octal bases used for V13, V14 and V16 all valveholders are B9A Noval with skirts and screening cans. If an EA50 is used for V15 it may conveniently be wired directly into circuit.

All fixed capacitors of less than 1000 pF in value should be silver mica of good quality, and where two of these are used across a resonant circuit to provide a centre-tap, they should be 1 per cent tolerance types. Unless otherwise stated, all other values of 1000 pF used in the prototype were Hunts Moldseal Type W45 of 400 volts rating. Resistors used throughout are Erie Type 8, ½ watt rating, unless otherwise specified.

The anti-parasitic chokes in the anodes of the 6146 valves are made in the usual way by space winding 6 turns of 18 s.w.g. tinned copper wire over the body of a 47 ohm ½ watt resistor. The anode tuning capacitor VC7 is a Jackson Bros. Type C604 although any similar capacitor should be satisfactory. The aerial loading capacitor is a standard two-gang 500 pF each section broadcast type.

A standard pie wound r.f. choke is unsuitable for use with

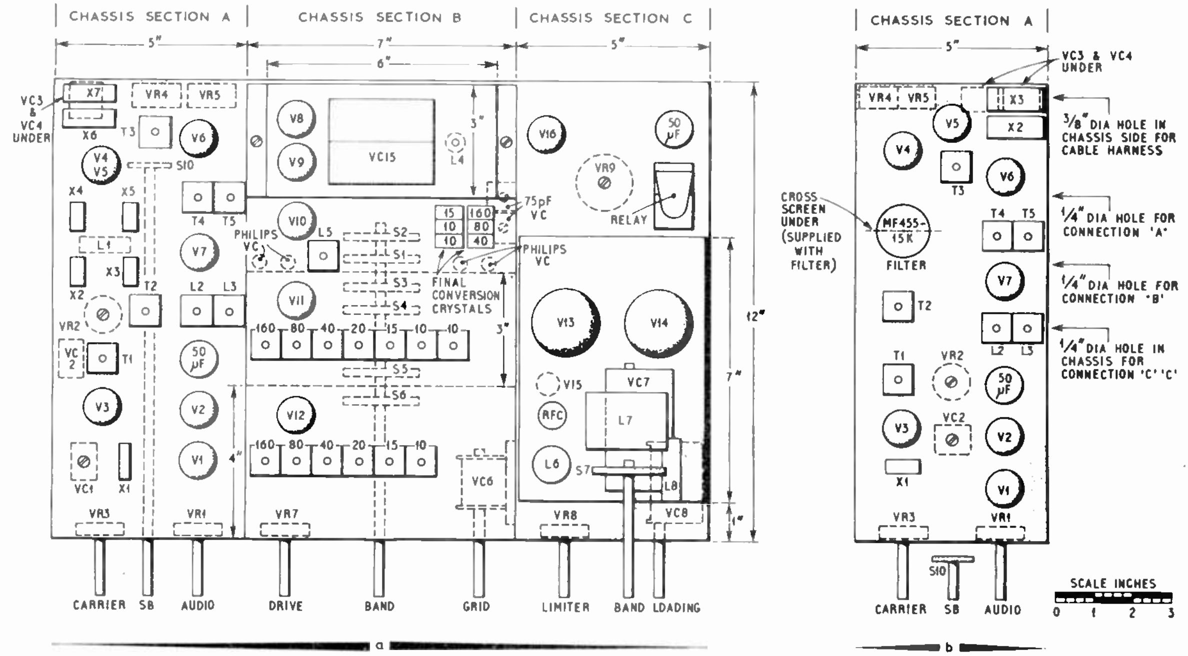


Fig. 10.105 (a). Chassis layout showing the principal components. (b) Chassis layout for the l.f. filter sideband generating unit (chassis section A). The sideband selection switch S10 is mounted directly on the panel, which is positioned ½ in. away from the chassis front apron.

a shunt fed tank circuit and this component must be specially wound to have a low self-capacity and no self-resonant points within the required amateur bands. A satisfactory choke comprises 250 turns of 36 s.w.g. enamelled wire wound in unequal sections of 150, 50, 25, 15 and 10 turns on a ceramic former ¾ in. diameter and 4 in. long with a ⅛ in. spacing between each of the sections.

All the crystals used in the h.f. filter unit (X1-7) can be FT243 types. Alternatively, a manufactured filter unit could be employed. Where necessary these were manipulated on to the required frequency by etching. The carrier crystal for the l.f. mechanical filter unit can be ordered either with the filter, or obtained later to the frequency specified.

In all cases when ordering, specify that the crystal is required for operation at the quoted frequency on the parallel resonant mode with 30 pF shunt capacity. (The only exception would be the h.f. filter crystals—these are operating on the series resonance. Any constructor unable to do his own etching would be well advised to build the l.f. sideband generator using the specified mechanical filter.)

Construction

The chassis is made up of three separate box sections of 16 s.w.g. aluminium to give a total size of 17 in. × 12 in. × 3 in. deep with a 19 in. × 10½ in. panel. This enables the transmitter to be constructed and wired in units. Elaborate screening is not necessary and adequate protection against instability in the final conversion section is achieved with two simple cross-screens that also afford support to the main band-change switch assembly.

A pair of 6146 valves will give an appreciable output and as a precaution against any possibility of stray r.f. fields, the two valves and the components in the anode circuit are screened with a 16 s.w.g. aluminium " box " measuring 7 in. × 5 in. × 6½ in. high. This is in fact two separate *L*-shaped screens with ½ in. lips top and bottom, the inner screen (nearest to panel and chassis centre line) being permanently fixed to the main chassis section with PK self-tapping screws and used to support the pi-tank switch and coil and the anode tuning capacitor; the outer screen is removable for access to the p.a. valves. It is most important that there is a free flow of air throughout the p.a. compartment; accordingly a ring of $\frac{3}{16}$ in. diameter holes is drilled round each octal valveholder and the " box " top is made from expanded aluminium mesh, also held in position by PK screws.

The chassis layout showing the position of the principal components is given in Fig. 10.105 A panel layout with suggested positioning of the control knobs and the Eddystone 898 dial in **Fig. 10.106.** It will be noted that the p.a. screening box is set back 1 in. behind the front chassis apron—this is necessary to clear the 898 drive mechanism.

The Yaxley two-pole changeover switch bank S10 must be mounted reasonably close to the two crystals X6 and X7 and the oscillator valve V5 (this applies only to the h.f. filter sideband generating section). It is supported by a small *L*-shaped bracket bolted to the underside of the chassis. A clear space should be kept down the centre of this section to clear the switch control rod and support bearing before connection to the Eddystone flexible coupler that is brought out to the front panel control knob. The use of individual coil cans on top of the chassis and switch banks centrally

TABLE 10.19

Function	Freq. or Band	Winding	Resonating Capacity
Carrier oscillator T1	8·5 Mc/s	Primary 32 turns 32 s.w.g. enam. Secondary 22 turns 32 s.w.g. enam. against cold end of primary.	50 pF s.m.
Filter input T2	8·5 Mc/s	Primary 32 turns 32 s.w.g. enam. Secondary 8 turns 32 s.w.g. enam. against cold end of primary.	See Figs 5 and 6
Filter amplifier T3	8·5 Mc/s	Secondary 45 turns 32 s.w.g. enam. Primary 25 turns 36 s.w.g. enam. over centre of secondary with one layer of paper insulation.	47 pF—47 pF 1% tolerance s.m.
V6 Conveter anode I4	2·0 Mc/s	Primary 90 turns 36 s.w.g. enam. Secondary 5 turns 36 s.w.g. enam. over centre of primary with one layer of paper insulation.	450 pP—450 pF 1% tolerance s.m.
V7 Converter grid T5	2·0 Mc/s	Secondary 90 turns 36 s.w.g. enam. Primary 5 turns 36 s.w.g. enam. over centre of secondary with one layer of paper insulation.	450 pF—450 pF 1% tolerance s.m.
V7 Converter anode L2	5·0 to 5·5 Mc/s	75 turns 36 s.w.g. enam.	65 pF—65 pF 1% tolerance s.m.
V11 Converter grid L3	5·0 to 5·5 Mc/s	75 turns 36 s.w.g. enam.	65 pF—65 pF 1% tolerance s.m.
V.F.O. L4	3·0 to 3·5 Mc/s	18 turns 19 s.w.g. enam. on $\frac{1}{2}$ in. diam. former with dust core. Cambion 1538-2-2 ceramic with 20063-K slug is recommended.	VC5 tuning of 250 pF maximum capacity.
V10 Oscillator anode L5	10m (28·5-29 Mc/s)	13 turns 24 s.w.g. enam.	10 pF s.m.
	10m (28-28·5 Mc/s)		30 pF Philips trimmer.
	15m		30 pF Phil. trim. + 50 pF s.m.
	20m		30 pF Phil. trim. + 220 pF s.m.
	40m		30 pF Phil. trim. + 100 pF s.m.
	80m		75 pF variable + 200 pF s.m.
	160m		75 pF variable + 350 pF s.m.

Function	Freq. or Band	Winding	Resonating Capacity
V11 Converter anode	10m (28·5-29, Mc/s)	Primary 7 double turns 24 s.w.g. enam. Secondary 7 turns 24 s.w.g. enam. over centre of primary.	None
V11 Converter anode	10m (28-28·5 Mc/s)	Primary 8 double turns 24 s.w.g. enam. Secondary 8 turns 24 s.w.g. enam. over centre of primary.	None
	15m	Primary 10 double turns 24 s.w.g. enam. Secondary 10 turns 24 s.w.g. enam. over centre of primary.	None
	20m	Primary 12 double turns 28 s.w.g. enam. Secondary 12 turns 28 s.w.g. enam. over centre of primary.	15 pF s.m.
	40m	Primary 25 double turns 32 s.w.g. enam. Secondary 25 turns 32 s.w.g. enam. over centre of primary.	30 pF s.m.
	80m	Primary 40 double turns 36 s.w.g. enam. Secondary 20 turns 36 s.w.g. enam. over centre of primary.	27 pF s.m.
	160m	Primary 70 double turns 38 s.w.g. enam. Secondary 35 turns 38 s.w.g. enam. over centre of primary.	60 pF s.m.

In all cases the primary is bifilar wound with the wire laid on double (i.e. 7 double turns = total primary of 14 turns). The secondary is insulated from primary with one layer of typewriter copy paper. (Three thou. thickness.) A bifilar winding is simply two interwoven windings, with the left-hand end of one winding connected to the right-hand end of the other winding to form the centre tap.

Function	Freq. or Band	Winding	Resonating Capacity
V12 Class A Amplifier anode	10m	9 turns 20 s.w.g. enam. spaced to $\frac{1}{2}$ in. long.	VC6 tuning
	15m	11 turns 24 s.w.g. enam.	VC6 tuning
	20m	18 turns 24 s.w.g. enam.	VC6 tuning
	40m	55 turns 32 s.w.g. enam.	VC6 tuning
	80m	90 turns 36 s.w.g. enam.	VC6 tuning
	160m	150 turns 38 s.w.g. enam.	VC6 tuning plus 80 pF s.m. across coil.

All coils are close wound unless otherwise stated. After winding all coils are doped with "Denfix" polystyrene cement. With the exception of L4 all coils are wound on 0·3 in. diam. formers with dust cores and $2\frac{1}{4}$ in. x $\frac{3}{16}$ in. x $\frac{3}{16}$ in. aluminium cans (Neosid or Aladdin). In all cases resonating capacitors are wired across coil and transformer connecting tags underneath the chassis. (s.m. = silver mica).

disposed underneath gives a clean accessible layout together with very short coil connecting wires. Switch sections S1, S2, S3 and S4 are supported by the rear cross screen and S5 and S6 by the front cross screen—these can conveniently be assembled before the screens are bolted in position in the central box section.

In order to bring the drive shaft in line with the driving boss of the Eddystone 898 dial assembly, the v.f.o. tuning capacitor must be raised above chassis level. For the type of tuning gang used in the prototype this required a platform

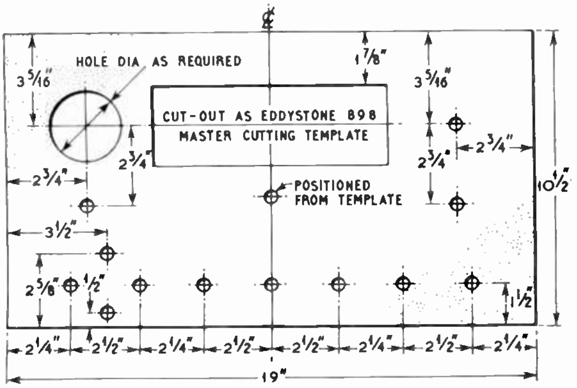


Fig. 10.106. Front layout. In order to eliminate undue flexing, the panel is cut from ⅛ in. thick aluminium alloy. A template is supplied with every Eddystone 898 dial to ensure that the mounting holes and the rectangular cut-out are accurately positioned.

2½ in. high. In order to obtain the utmost rigidity an aluminium box section 6 in. × 3 in. × 2½ in. was used—held to the main chassis by 3 in. × ½ in. × ½ in. *L*-section angle plates—and this provides ample room for the v.f.o. and cathode follower valves (V8 and V9). All the other associated components are mounted inside the box. This unit can be completely built and tested before being fitted into position.

A balanced bridge circuit is used for neutralization of the 6146 valves and this requires that the cold ends of the six coils and the rotor plates of the 50 pF tuning capacitor be taken to a common bus-bar which is insulated from earth and bypassed with a 500 pF capacitor. The 50 pF tuning capacitor (Polar Type C28-141) is mounted on a paxolin plate 2½ in. × 2½ in. × ⅛ in. thick; this in turn is bolted to the chassis side apron with the capacitor spindle extended through to the control knob with a ¼ in. diam. paxolin rod.

TABLE 10.20

FINAL CONVERSION CRYSTAL FREQUENCIES

Band	Crystal Frequency (Mc/s)	Output	Output Frequency (Mc/s)
160	7·0	Fund.	7·0
80	9·0	Fund.	9·0
40	6·25	2nd Har.	12·5
20	9·0	Fund.	9·0
15	8·0	2nd Har.	16·0
10	11·5	2nd Har.	23·0
10	11·75	2nd Har.	23·5

The pi-network values have been calculated for an R_L (anode load) of 2000 ohms and are given for each band in **Table 10.21**. The tank inductance is made up of three separate units with the 160m coil L8 on a separate ceramic former positioned at right angles to the main winding L7 and adjacent to its " cold " end with the self-supporting 10m coil L6 positioned at the " hot " end to form the connecting link between L7 and the stator plates of the anode tuning capacitor. Because of the difficulty of obtaining a suitable diameter ceramic former, the main tank coil L7 is hand wound over four 2½ in. lengths of ¼ in. diameter polystyrene rod held in position on a mandrel made from a 4 in. length of 1½ in. diameter birch dowel rod previously prepared with four semi-circular grooves down its length. The mandrel should be supported horizontally so that it can be turned as the tinned copper wire is run into position; as it crosses each polystyrene rod the wire is " cemented " into position by local heat applied with a soldering iron. After the winding is completed, the mandrel is withdrawn, and the coil carefully inspected to ensure that there are no shorted turns. A suitable ready made complete coil and switch assembly is the Codar type DAF/180/SW.

It is convenient to use a standard two gang capacitor of 500 pF each section for the aerial loading capacitor. As this would not have a large enough value for use on the lowest amateur band, the 160m position of the pi-tank band-change switch is used to bring into circuit a further fixed loading capacitor of 600 pF or thereabouts. The standard single pole, six way, Yaxley ceramic switch S7 is supported by the

TABLE 10.21

PI-NETWORK DATA—TWO 6146 VALVES
R_L = 2000 ohms

BAND	XC1 = 200 ohms	XL = 250 ohms	XC2 = 46 ohms
80m	220 pF	11·0 μH	900 pF
40m	110 pF	5·5 μH	450 pF
20m	56 pF	2·7 μH	225 pF
15m	38 pF	1·8 μH	160 pF
10m	28 pF	1·4 μH	115 pF

L6, 4 turns 16 s.w.g. tinned copper 1 in. diam. spaced to ½ in. long, self supporting.
L7. One continuous winding of 18 s.w.g. tinned copper wound 16 turns per inch, 10 turns, 3/16 in. gap, 8 turns, 3/16 in. gap, 3 turns, 3/16 in. gap and 2 turns. Total length approximately 2 in. Diameter 1½ in. supported by 4 lengths of ¼ in. diam. polystyrene rod.
L8, 32 turns 22 s.w.g. enam. close wound on ⅞ in. diam. ceramic former.

front apron of the p.a. screening " box " and the switch in turn supports the coil of L7 by means of the two ends and the three tapping connections. S7 and L7 are conveniently made up as a unit before fitting into position.

On the 160m band, power is reduced to the equivalent p.e.p. of a 100 per cent modulated 10 watt d.c. input A3 transmitter by switching the anode feed to the p.a. valves from the normal 700–750 volts line to the 300 volts rail feeding the remainder of the transmitter.

As the transmitter is controlled by a " press-to-talk " switch (this may be a press switch on the microphone or a foot switch) there is no relay clicking while the operator is actually talking. The Type 3000 relay does not require a flexible rubber shock mount and may be screwed directly to the chassis.

The meter switch, S8 and S9, is a two bank, two pole, four way Yaxley type. This may be paxolin but ceramic is prefered to eliminate any possibility of tracking. It is most important that this switch is the "break-before-make" type, otherwise h.t. would be momentarily connected to the 6146 grids and the OA79 diode as the switch poles moved over. If this type of Yaxley wafer is not available, standard "make-before-break," single pole seven-way, banks can be used with each adjacent contact left blank (i.e. the four positions of the control knob would be at 60° instead of the customary 30°.)

Any good quality moving coil meter of 1 or 2 mA full scale deflection is suitable and is more conveniently read if the original scale is re-numbered to read 0 to 3, 0 to 30 and 0 to 300 mA. The meter shunts, SH1, SH2 and SH3, may be constructed by winding Eureka resistance wire round a 100 ohm Erie (ceramic body) resistor. The value of resistance wire to shunt the meter correctly for each of the required ranges will depend on the meter internal resistance and will have to be found experimentally. (See Chapter 19—*Measurements.*)

The coil L1 in the h.f. filter is constructed by taking a length of 22 s.w.g. p.v.c.-insulated connecting wire, doubled back on itself to form two parallel wires. This is then wound on a ferrite ring core to form nine double turns (total 18 turns) and the inner of one winding connected to the outer of the other to form a bifilar winding with the junction the centre tap. The main requirement of the inductance is very tight coupling between each half, together with a perfect electrical balance. Filters have been successfully built using the ring from a Mullard LA4 pot core assembly (35mm outside diameter) and also the ring from a mains suppressor filter choke (1⅜ in. outside diameter.)

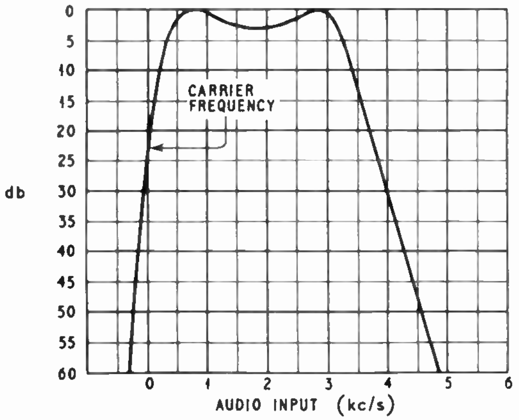


Fig. 10.107. Filter response curve plotted with an audio generator. In this diagram, zero audio frequency corresponds with the carrier frequency, nominally 25 db down the filter passband. Note that the slope on the carrier side is steeper than the actual filter response due to the bass roll-off in the audio amplifier.

To give some guidance as to the characteristics that may be expected, the plot of the filter passband of a typical filter is shown in **Fig. 10.107**.

The carrier crystal frequency is determined by plotting the filter passband and marking the 20db down points. One of the remaining FT243 crystals is etched so that its *parallel* resonant frequency is at this frequency. Finally when the transmitter is completed and tested on the air the carrier crystal can be "pulled" by means of the 50 pF trimmer capacitor across the grid circuit of V3 to obtain the best balance of voice quality.

The original filter was constructed on a nominal frequency of 8·5 Mc/s. However it does not in practice have to be on this frequency and any crystal in the FT243 range can be used from 6500 kc/s up to 8650 kc/s—provided of course that the sideband switching crystals are altered to suit the new carrier frequency. That is, X6 = carrier frequency less 2·0 Mc/s, and X7 = carrier frequency plus 2·0 Mc/s. One of the inherent advantages of heterodyning the initial single sideband output down to the first intermediate frequency of 2·0 Mc/s is the fact that there is a considerable flexibility in the choice of filter centre passband frequency.

Alignment and Operation

The most satisfactory alignment procedure is to start at the back and finish at the front—that is from the p.a. tank coil back to the carrier oscillator—using a serviceman's signal generator as the signal source.

In the interest of valve life the 6146s should be run with the power supply h.t. switch in the 300 volt position until the anode and grid tuning has been checked and the valves correctly neutralized on all bands.

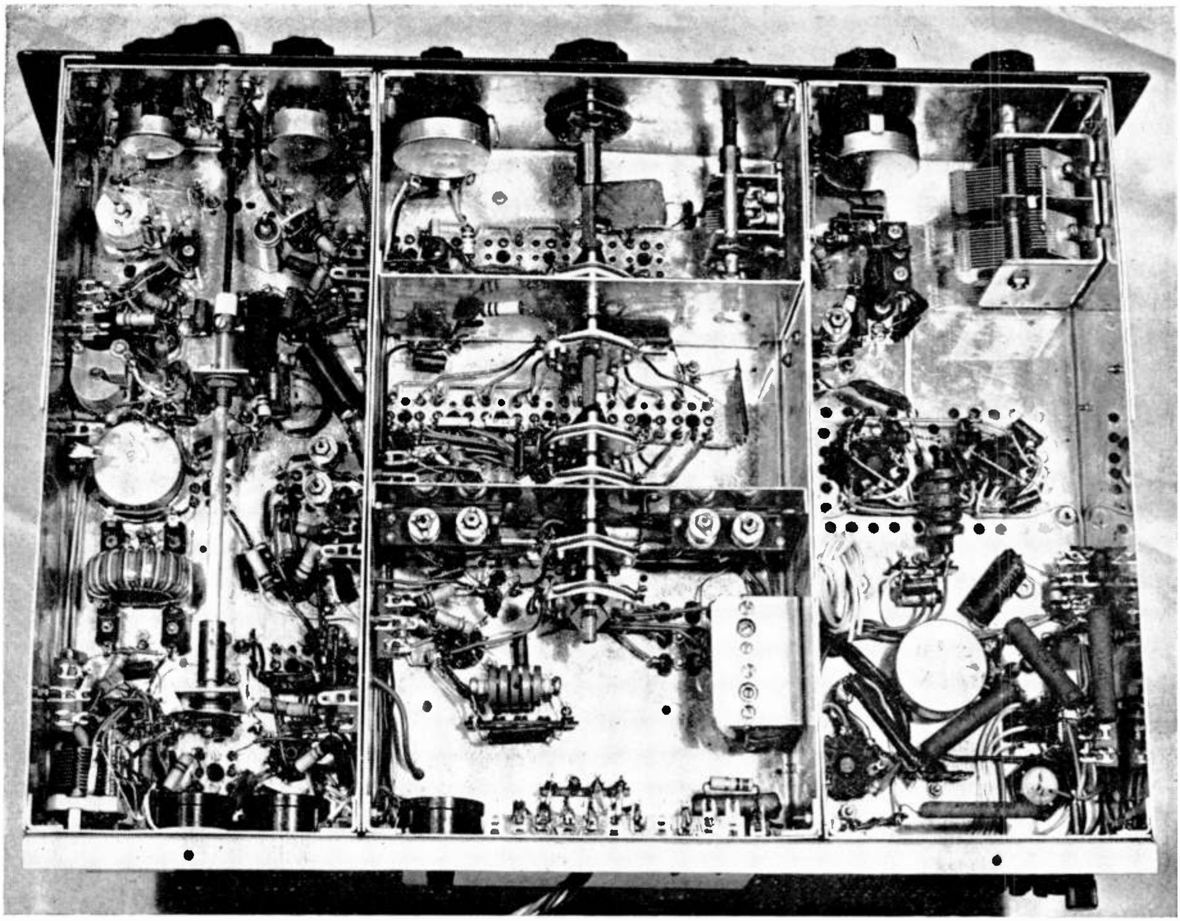
All oscillators should be checked for satisfactory operation and ideally the amplitude of r.f. output voltages should be checked with a diode probe valve voltmeter and compared with the values given in **Table 10.22.**

All coils are in screening cans and a grid dip oscillator cannot be brought up to the coil to check alignment, neither is it possible to use the usual pick-up loop feeding into an absorption wavemeter. This may cause difficulty when it is required to set each of the pre-set trimmer capacitors across the final conversion oscillator coil L5, because it is important on some ranges to know that the circuit is resonant on the correct harmonic. There are two ways to overcome this limitation:

(i) a pick-up loop tucked into the wiring close to the valve anode will pick up sufficient r.f. to deflect a sensitive absorption wavemeter using a diode and 100 μA meter; alternatively the "signal" can be fed into a receiver via a length of co-axial cable.

(ii) a pick-up loop can be constructed by winding six turns of 24 s.w.g. enamelled wire round the end of a 6 in. length of ⅛ in. diameter sleeving, the loop connections then feeding back down the centre of the sleeving and terminating with a length of twin flex or co-axial cable. This "probe" can then be pushed down towards the dust core inside the former of the circuit to be checked.

The bandpass coupler made up of L2 and L3 is required to pass all frequencies between 5·0 and 5·5 Mc/s and offer considerable attenuation to all unwanted converter products outside the required passband. This bandwidth of 500 kc/s at the 3db points is obtained by overcoupling the two circuits so that a double humped response curve is formed as shown in **Fig. 10.108.** Under these conditions, however, alignment would be most difficult because adjustment of one



A view beneath the chassis of the G2DAF s.s.b. transmitter. The toroidal core and its mounting can be seen adjacent of the sideband switch shaft on the left-hand side in the high frequency crystal filter compartment. All wiring is point-to-point, small tag strips being placed in appropriate positions to anchor the smaller components.

circuit would pull the other. To overcome this difficulty Philips concentric trimmer capacitors are used for "top capacity" coupling. Initially the two trimmers are fully unmeshed by rotating anticlockwise until the movable vanes are at the top of the screwed stem. The dust cores of L2 and L3 are then adjusted for exact resonance with a 5·25 Mc/s input signal. Finally the two Philips trimmers are each screwed inwards two complete turns.

After all other circuits are aligned and the transmitter is ready for use on the air, the bandpass coupler response can be checked by setting the bandchange switch to 20, 15 or 10m. Turn up the CARRIER INSERTION control to drive the p.a. to some pre-determined anode current—say 100 mA. The v.f.o. tuning capacitor is then traversed completely across the 500 kc/s tuning range and if all is well the current should not drop more than about 3db (from 100 mA down to 70 mA) at the two ends of the tuning scale (over each of the seven switched bands, the tuned circuit connected to each of the anodes of the final converter, V11, is resonated at the centre band frequency. This rise in response compensates for the dip in the centre of the 5·0 to 5·5 mc/s bandpass coupler to give a substantially constant gain). If the drop is greater than 3db the capacity of the two Philips trimmers should be increased slightly. It is most important that the bandpass coupler rejection to unwanted frequencies is not spoilt: over-coupling must be avoided. Use the smallest amount of "top coupling" that will give the required bandwidth.

TABLE 10.22

VALVE VOLTMETER R.M.S. CHECK VOLTAGES

Audio	Cathode of V2b	Gain control at maximum Gain control at normal	0·5V 1·5V
R.F.	Junction of each diode and VR2.	Ratio of values will depend on setting of VR2	0·6V 0·67V
	Anode of V3a		12·0V
	Cathode of V3b		8·5V
	Oscillator input to T3	Measured at T3 side of coupling capacitor	5V
	V.F.O. input to T5		5·5V
	Oscillator input to L3	Measured at L3 side of coupling capacitor	7 to 15V depending on band in use
S.S.B.	Either grid of V11	V10 removed to disable oscillator.	2·4V
	Grid of V12	10m (28·5–29 Mc/s)	8·5V
		10m (28–28·5 Mc/s)	6·6V
		15m	8·0V
		20m	8·4V
		40m	10·0V
		80m	12·0V
		160m	6·0V
	Anode of V12	10m (28·5–29 Mc/s)	95V
		10m (28–28·5 Mc/s)	97V
		15m	120V
		20m	128V
		40m	134V
		80m	155V
		160m	170V

Audio and s.s.b. measurements are made with 1·25 kc/s audio input to the microphone socket and level set to give maximum output from the transmitter unless otherwise stated. When taking measurements at the anode of V12 the 150 pF capacitor is disconnected from the grids of V13 and V14 and the R.F. DRIVE control is adjusted on each band for maximum output.
All readings taken with a valve voltmeter of 11 Megohms input impedance. NOTE: due to the input capacity of the diode probe the circuit under measurement will require to be re-resonated.

It will be noted that there are seven positions of the main bandchange switch, one each for the five bands from 160 to 15m and two for the 10m band to give a 28–29 Mc/s coverage. The anode circuit of V11 is pre-tuned at the centre band frequency for each of the switched ranges and there are seven separate coils. The variable tuning capacitor in the anode circuit of V12 is brought out to a panel control knob, so that one tuned circuit will cover the whole of the 10m band; there are six coils, the two 10m contacts on S6 being strapped across as shown.

Table 10.22 shows the r.f. and sideband voltage readings at different stages in the transmitter. When measuring the voltages at the grid of V12, it should be remembered that the input capacity of the diode probe will de-tune the anode circuit of the converter V11 and give a false low reading. When the probe is in position each circuit must be brought back to resonance by unscrewing the dust core slightly until maximum reading is obtained. Similarly when measuring the voltage at the grids of the 6146 valves the diode probe should be connected to the V12 anode side of the 150 pF coupling capacitor and the capacitor should be temporarily disconnected from the grids of V13 and V14. This is essential, otherwise any peak r.f. drive that exceeds the negative bias potential (about 50 volts) will cause the p.a. valves to draw grid current, the input impedance will fall to a few hundred ohms and severely damp the resonant circuit thus giving a false low reading. Before taking valve voltmeter reading the circuit must be correctly resonated, with the diode probe in position, by the front panel GRID TUNING control.

The various oscillator output voltages are not particularly critical and are quite satisfactory if within plus or minus 20 per cent. If they are outside these limits they can most conveniently be adjusted by changing the value of the anode feed resistors.

A convenient method of plotting the sideband filter response curve—either for the h.f. or the l.f. filter—is to use an audio signal generator. The audio generator is connected to the microphone input socket and the AUDIO GAIN control VR1 advanced until the tone input is driving the transmitter at some convenient level. The valve voltmeter probe can be connected either to the anode of V4 or the grid connection to V13 and V14. The passband should be similar to the curve shown in Fig. 10.107.

The transmitter has considerable reserve of voltage amplification. On the lower frequency bands the R.F. DRIVE control is backed off and progressively increased as the operating band goes higher. It should only need to be fully on when operating on 10m. There is also reserve of audio gain (sufficient for a low level microphone). The correct setting of the audio gain control when using a normal crystal microphone is between half and two-thirds of the maximum position.

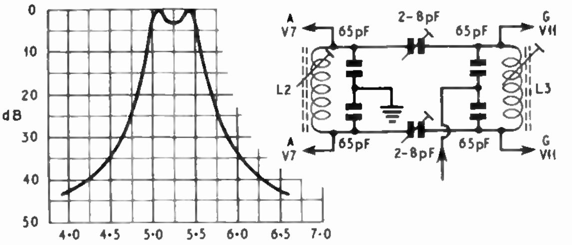


Fig. 10.108. Plotted response curve of the 500 kc/s wide bandpass coupler.

The unwanted sideband suppression at 1 kc/s (measured at the transmitter output terminal) for either filter is better than 55db. The low impedance diode modulators make the carrier balance particularly stable. The carrier suppression is better than 60db. Due to the use of double triode balanced converters in each of the frequency translation stages and negative r.f. feedback in the penultimate amplifier, the transmitter gives a clean output with a low level of intermodulation distortion products.

"AM" operation is obtained by putting in a small amount of carrier with the aid of the carrier insertion control VR3. The output is single sideband with carrier and is therefore A3H.

For c.w. operation the transmitter is best controlled by a " press-to-talk " foot switch. A key jack socket of the closed circuit type can conveniently be connected in the cathode return of the final balanced converter V11 (between the slider of VR6 and earth). The keying characteristic is clean and free from chirp. The audio gain control should be turned to zero and the CARRIER INJECTION control advanced until the required amount of carrier is available at the transmitter output.

Power Supply

An r.f. power amplifier can only operate in a linear manner and give an undistorted output if it is provided with grid, screen and anode voltages that remain constant in potential and do not vary under differing load conditions. The heavy anode current demand is at syllabic rate and the main h.t. power supply must therefore have good regulation under dynamic conditions.

The circuit of a suitable power supply is shown in Fig. **10.109**.

CONVERTING TO 144 MC/S

There are obviously many different methods of obtaining an s.s.b. output on 144 Mc/s. The most flexible and practical arrangement is however to use an existing s.s.b. exciter and convert the 14 or 28 Mc/s output to the 144 Mc/s band when this is required.

An output power of about 1 watt r.m.s. is sufficient fully to drive the usual 144 Mc/s p.a. in class AB1, and this amount of power can be obtained from a high level mixer using standard B9A valves. The basic 14 Mc/s s.s.b. signal

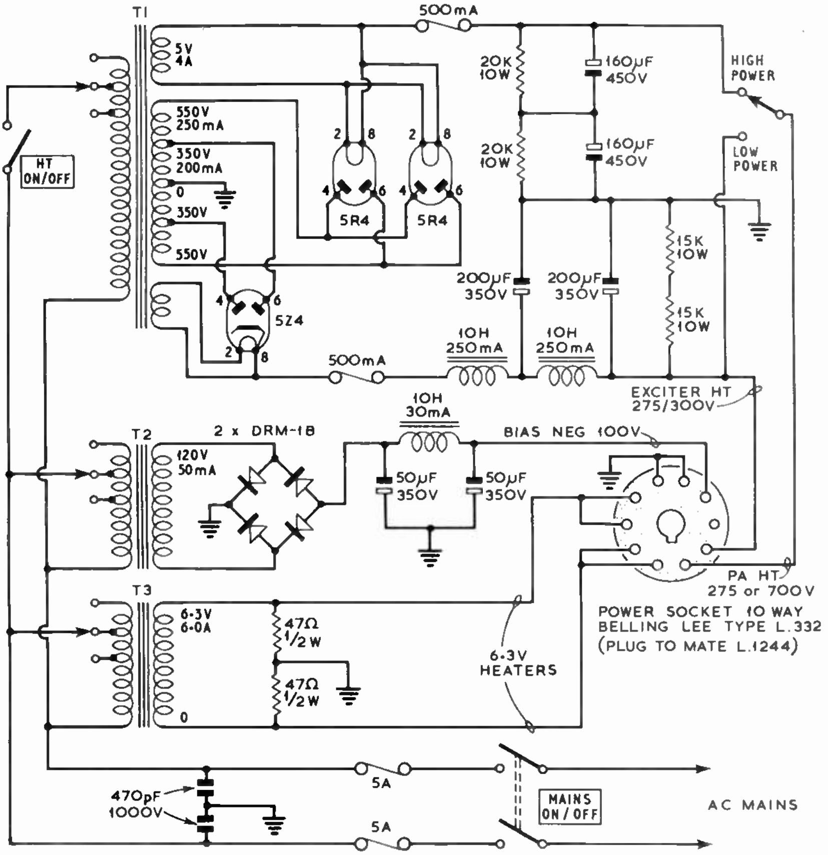


Fig. 10.109. The power supply for the G2DAF transmitter. The two 160 µF 450 volt capacitors are standard 100 + 60 µF electrolytics with the two sections strapped in parallel. T1 may be two separate transformers. The heater winding (T1) for the 5Z4 rectifier is 5 volts at 2 amps.

has to be heterodyned to 144 Mc/s by mixing it with the output from a crystal controlled oscillator. A very satisfactory yet relatively simple method of doing this has been developed by G3LNP.

The circuit diagram is shown in **Fig. 10.110** and it will be seen that the first half of the 12AT7 is a third overtone oscillator and the second half a tripler; the output at the coil L3 is therefore nine times the frequency of the controlling crystal X1. This output at about 130 Mc/s is further amplified by the EL91 buffer amplifier and fed via a link winding to the coil L6. The voltage developed across L6 is then fed into the "centre tap" of the balanced mixer grid input circuit so that the heterodyning voltage is connected to the mixer grids in parallel and can be balanced out in the output circuit L7 by adjustment of VR1, the balancing control. The 14 Mc/s output from the basic s.s.b. exciter appears across the resonant input circuit L8 and the two series connected 50 pF capacitors and feeds the sideband input into the mixer in push-pull. L7 and the 10 pF variable capacitor form a push-pull anode circuit tuned to the sum of the two input frequencies (i.e. 130 + 14 Mc/s) giving the required s.s.b. output at 144 Mc/s.

The heterodyning input is of course fixed by the crystal frequency and the tuning of the 144 Mc/s output frequency is done on the main 14 Mc/s exciter. The frequency of X1 is therefore determined by the wanted output in the 144 Mc/s band, and by the available tuning range in the exciter. The RSGB recommends that s.s.b. operation on 2m should take place near 145·4 Mc/s. This requires an heterodyning frequency of 131·0 Mc/s obtained from a 14·567 Mc/s crystal, with the 14 Mc/s exciter tuning 14·1–14·2 Mc/s.

Although a 14 Mc/s crystal can sometimes be made to operate on the third overtone, it is preferable to use a proper third overtone crystal on 43·733 Mc/s.

Finally, a word of warning: it is important that the 14 Mc/s s.s.b. input signal be completely free from any other mixer products or v.f.o. harmonics. Many exciters in use do not have a "clean" output by any means and rely on the additional attenuation of the following linear amplifier and aerial tuning circuits. This may be satisfactory on 14 Mc/s, but when the exciter is driving the 144 Mc/s mixer these circuits will not be in use and any spurious outputs will be fed into the 5763 balanced mixer and will be heterodyned by the 131·2 Mc/s input and appear in the 2m output that is driving the linear amplifier. There is therefore the danger that the transmitter will be radiating several frequencies, some of which could be outside the amateur band. An example of this would be a v.f.o. operating on 5·1 Mc/s, some of the third harmonic output of which on 15·3 Mc/s appears in the 144 Mc/s exciter output. This would produce (with a 14·567 Mc/s crystal) a spurious output on 146·5 Mc/s. If the v.f.o. were on 6 Mc/s the third harmonic would be 18 Mc/s and produce a spurious output on 149·2 Mc/s. While linear amplifier circuits operating on 14 Mc/s can discriminate against a spurious product 2 or 3 Mc/s away from the wanted output, they cannot possibly do so when they are operating at the high frequency of 145 Mc/s or thereabouts. Before putting the 2m s.s.b. transmitter on the air check carefully and make sure that the output is the one wanted and that it is "travelling alone."

The alignment of the G3LNP unit should be carried out in the following order:

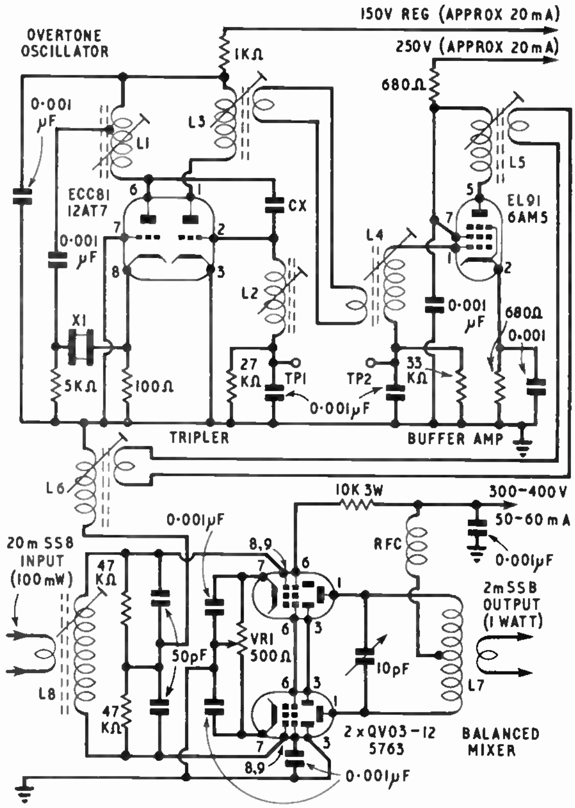


Fig. 10.110. Circuit of the unit designed by G3LNP for converting a 14 Mc/s s.s.b. signal to 144 Mc/s. L1, 2, 15 turns 36 s.w.g. enam. (L1 tapped at 1·5 turns from cold end); L3, 4, 5, 4 turns 26 s.w.g. enam. spaced wire diameter; L6 3 turns 20 s.w.g. enam. spaced wire diameter; L7, 3 turns 16 s.w.g. enam. 1 in. dia. self-supporting with 1 turn link of p.v.c. ins. wire; L8, 20 turns 24 s.w.g. enam. on Aladdin ⅜ in. dia. dust iron core former with 4 turn link of p.v.c. insul. wire at centre. L1-4 are wound on Aladdin ¼ in. dia. formers with dust iron cores and L5-6 on similar formers with brass cores. The link windings for L3, 4, 5 and 6 are each 2 turns of p.v.c. insulated wire at the cold end of the coil.

(i) Adjust the dust cores of L1 and L2 for maximum voltage at test point TP1.

(ii) Adjust Cx to give about negative 40 volts at TP1. (Capacitor value will be about 1 pF.)

(iii) Adjust the cores of L3 and L4 for maximum voltage at test point TP2.

(iv) Adjust the cores of L5 for *minimum* voltage across the 680 ohm cathode resistor.

(v) Adjust the core of L8, L6 and the 10 pF tuning capacitor across L7 for maximum s.s.b. output at 145 Mc/s.

(vi) Set VR1 for *minimum* 131·2 Mc/s output with no 14 Mc/s drive.

THE G3OQD TRANSISTORIZED SSB TRANSCEIVER

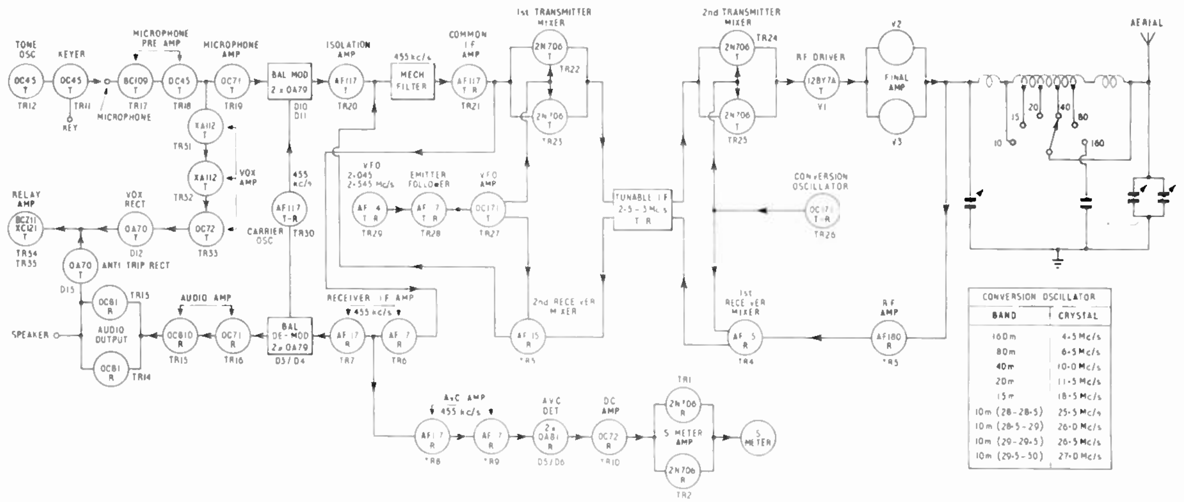
This transceiver is suitable for mobile use, or as the home station equipment.

The main features include:

(i) Coverage of all bands from 160m to 10m, with 500 kc/s tuning segments on each range.
(ii) A maximum input of 200 watts p.e.p. on s.s.b. (either sideband) or c.w.; 75 watts on a.m. (equivalent to about 50 watts input for a class C transmitter).
(iii) Operation from an a.c. power supply or from a 12 V d.c. transistorized d.c.-to-d.c. converter.
Construction can be simplified by omission of the following optional circuitry. Details are given in the relevant sections.
(iv) VOX control.
(v) Independent receiver and transmitter tuning.
(vi) A.M. and c.w. modes,
(vii) Automatic level control,
(viii) Sideband switching.

where it is mixed with the v.f.o. output, tunable over the range 2·045–2·545 Mc/s, to give a second i.f. of 455 kc/s. Selectivity is obtained in the common i.f. stage with further amplification being obtained in the receiver i.f. amplifier TR6 and TR7. A balanced demodulator is used for detection with carrier insertion provided by TR30. A.g.c. is tapped off half-way along the i.f. amplifier and further amplified by TR8, TR9 and associated d.c. amplifier TR10, the output of which is used to control the gain of TR3, TR21 and TR6.

When on transmit, the microphone input is amplified by TR17, TR18 and TR19, with TR17 and TR18 being common to both the audio amplifier and the VOX amplifier. The amplified audio signal is fed to the balanced modulator together with carrier from the carrier oscillator TR30, the resulting double sideband signal being fed to the mechanical filter, via the isolation amplifier TR20. The mechanical filter attenuates one sideband; the resulting s.s.b. output is then amplified by the common i.f. stage TR21, which provides a balanced output to the first balanced mixer TR22 and



BAND	CRYSTAL
160m	4·5 Mc/s
80m	6·5 Mc/s
40m	10·0 Mc/s
20m	11·5 Mc/s
15m	18·5 Mc/s
10m (28–28·5)	25·5 Mc/s
10m (28·5–29)	26·0 Mc/s
10m (29–29·5)	26·5 Mc/s
10m (29·5–30)	27·0 Mc/s

Fig. 10.111. Block diagram of the G3OQD S.S.B. Transceiver, using parallel 6GJ5As in the P.A. stage.

The block diagram **Fig. 10.111**, gives an overall picture of the transceiver by showing the signal paths and relationships between the various stages. Considering the receiver operation first, the input signal from the antenna is coupled into the receiver r.f. stage TR3 via the transmitter pi-tank. The first signal frequency conversion takes place in TR4, where the incoming signal is mixed with the output of the conversion oscillator TR26 to produce a tunable i.f. of 2·5–3 Mc/s. On 160m, 80m and 40m the first i.f. is produced by taking the difference frequency, i.e. $f_{i.f} = f_{osc} - f_{stg}$, while on 20m, 15m and 10m the sum frequency is used, so avoiding the necessity of switching sidebands when changing from, say 80m to 20m, thus maintaining consistent transmitter sideband quality throughout the six amateur bands.

The tunable i.f. is coupled into the second mixer TR5, TR23. High level v.f.o. signal is applied to the mixer, producing a signal at the tunable i.f. frequency. Conversion to the required amateur band is obtained in the second balanced mixer, TR24 and TR25, with oscillator injection provided by TR26. The output from this mixer is used to provide voltage drive to V1, the driver then feeding the p.a. stage in the conventional way.

Operation can be either 'press to talk' (PTT) or VOX; the VOX circuit consists of TR31, TR32 and TR33 which amplify the audio before rectification. The rectified audio is used to operate the VOX relay via the Darlington connected relay driver TR34, TR35. Anti-trip is obtained by rectifying audio output from the receiver and applying it in anti-phase to the relay driver. For c.w. a phase-shift oscillator TR12 is keyed by the keying transistor TR11.

RECEIVER SECTION

R.F. Stage

The problem of cross-modulation in bipolar transistors is due to their inherent non-linearity at large signal levels; also to the reduction in Q of the tuned circuits produced by the shunting effect of the low input and output impedances of the transistor.

These problems have been overcome to a large extent in the present design, by using the transmitter pi-tank to provide a high degree of selectivity before the r.f. stage. This arrangement offers the further advantage that the pi-tank may be resonated prior to power being applied to the p.a. The unloaded Q of the pi-tank is of the order of 200, and is therefore, capable of providing the high degree of selectivity required. When the p.a. is correctly neutralized, the output impedance on transmit will be a pure resistance in parallel with a capacitive component. This capacitive component is due to the output capacity of the p.a. valves plus stray capacity, and is constant on both transmit and receive.

Although the image rejection of the receiver is improved by the increased selectivity provided by the pi-tank, on 15m and 10m it is possible to resonate the pi-tank on the image frequency (5 to 6 Mc/s lower than the received frequency). This can be eliminated by a resonant circuit, series tuned to the image frequency, connected in shunt with the r.f. coil L1 (Fig. 10.112). If these traps are included, L1 must be re-resonated to the signal frequency. If, however, the pi-tank tuning capacitor is to be calibrated for each band, this extra complication is hardly worth while.

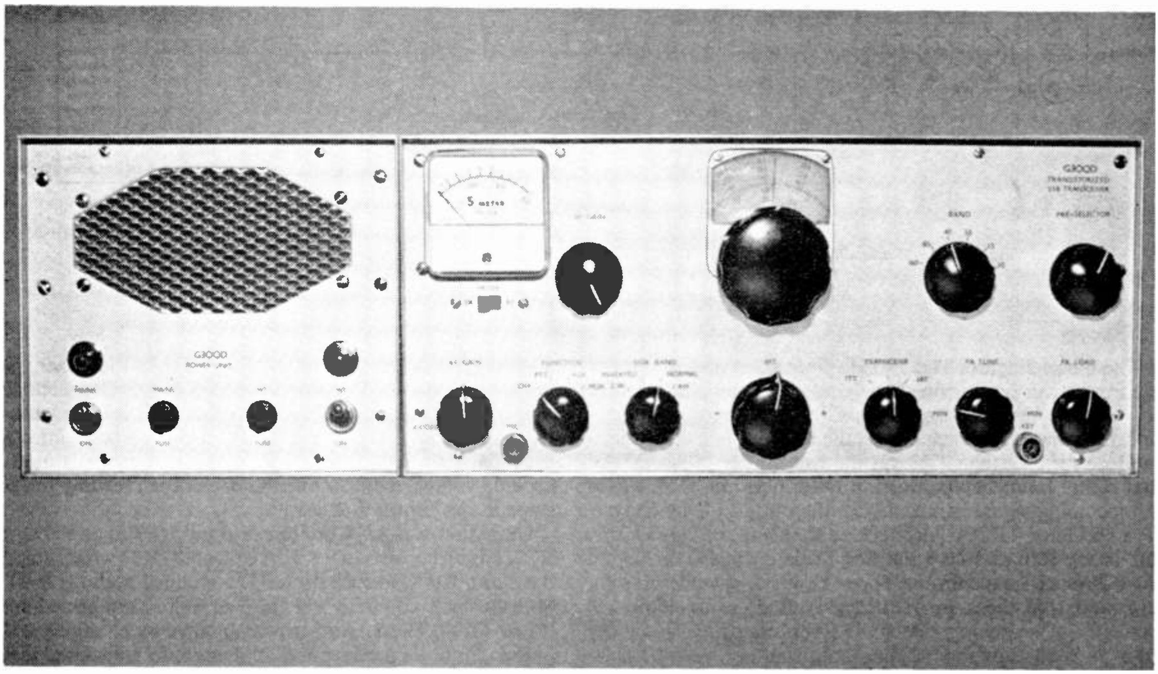
Signal injection into the receiver r.f. amplifier has been kept as small as possible, consistent with a good signal to noise ratio, by using a coupling capacitor C8 of 1·5 pF between the high impedance side of the pi-tank and the high impedance point on L1. A further reduction of strong signal effects is obtained by running the transistor at a slightly higher collector current than is usual.

When field-effect transistors were first introduced their specifications promised great improvements in cross-modulation characteristics when used in the front end of a transistor receiver. However, as the cross modulation performance of this transceiver compares favourably with that of valve equipment, it was decided to postpone the use of a Field Effect transistor.

Receiver front-end protection during transmit is provided by relay contacts RLB1, which short the receiver input to earth, thus preventing any r.f. from the transmitter reaching the front-end transistor.

The r.f. amplifier uses an AF180 in the common base configuration; neutralizing is not required. Stage gain is controlled by reverse a.g.c. action, or by the manual RF GAIN control, included so that overload from strong local signals can be prevented.

The r.f. stage is tuned to resonance by the ganged variable capacitors C10 and C11 which form the PRE-SELECTOR tuning.



Front panel layout of the G30QD transistorised SSB transceiver and matching power supply unit which also houses the loudspeaker.

First Mixer

The first receiver mixer is a single ended configuration with conversion oscillator injection applied by C14 to the emitter. The output is coupled into the tunable i.f. by a low impedance coupling tap on L17.

Second Mixer

The input to the second mixer is coupled from the tunable i.f. by the tap on L16. The mixer configuration is identical with that used for the 1st mixer. The output from this mixer is coupled into the common i.f. amplifier by the tap on the standard double-tuned inter-stage i.f. coupling transformer IFT9. This is a compromise arrangement with the transmitter signal from TR20 applied to the normal collector tap, while the receiver is coupled in on the tap orginally intended to match into the base of a transistor. This arrangement does not seem to affect receiver performance and avoids the use of a special transformer.

Receiver I.F. Amplifier

This section employs two cascaded common emitter AF117s in an un-neutralized arrangement with a.g.c. applied to the first stage TR6. A balanced demodulator is used for s.s.b. and c.w. reception and is similar in form to the balanced modulator used in the transmitter section. A suitable low impedance feed to the balanced de-modulator is obtained by using the existing base tap on the standard double-tuned transformer IFT2.

On a.m. an envelope detector, consisting of an OA70 (D2), is switched into circuit by S2A. This part of the circuitry can be omitted if the a.m. facility is not to be included.

A.G.C. Amplifier

The feed to the a.g.c. amplifier is taken from the base of TR7 via a 1000 pF capacitor C21. The a.g.c. amplifier consists of two common emitter stages using single-tuned i.f. transformers. Damping resistors connected across the primaries of these two transformers ensure that the bandwidth of the a.g.c. amplifier is flat over all the bandwidths used for reception and that the response characteristic of the filter used in the common i.f. is not altered in shape by a.g.c. action. The output from IFT5 is rectified by D5, while D6 is used as an a.g.c. gating diode, charging up C37 to provide a long discharge time suitable for s.s.b. or c.w. reception. On a.m. this capacitor is switched out of circuit by S2B to provide the fast time constant needed for this mode.

If the a.m. mode is not included C37 should be wired permanently in circuit.

The a.g.c. line is fed by the d.c. amplifier TR10, the collector voltage being stabilized at 1·2 volts by the combined action of ZD1 (5·6 V) and ZD2 (6·8 V). This prevents variations in leakage current of the transistors associated with the a.g.c. circuits, due to temperature changes, from causing variations in the no-signal gain of the receiver.

Receiver Audio Amplifier

This audio amplifier is based on the Mullard 1W circuit, modified for 12 V operation and reduced in bandwidth to remove any unnecessary hiss due to excessive frequency response.

Almost any small transistor audio amplifier can be used as an alternative to this circuit, provided that it will operate from a negative 12 V supply and that a suitable feed to the anti-trip circuit can be arranged. A packaged Japanese amplifier (Duvidal type EG-1004) is a suitable alternative.

The loudspeaker used should match the amplifier chosen. If required, a headphone socket can also be included.

S Meter

A long-tail pair amplifier using silicon n.p.n. transistors TR1, TR2 drives the S meter. The use of n.p.n. transistors was found necessary to improve the linearity of the S meter scale. The long-tail pair has excellent stability if the two transistors are mounted close together so that any temperature variation will affect both equally, thus preventing zero wander.

The sensitivity of the S meter may be changed by altering the ratio of R35 to R36 which form a potential divider feeding a fraction of the a.g.c. control voltage to the S meter circuit. If this adjustment is made it will be necessary to re-zero the S meter by use of RV1.

The meter performs a dual role, i.e. as an S meter on receive and as a p.a. cathode current or r.f. output meter on transmit.

TRANSMITTER SECTION

Microphone Amplifier

The microphone amplifier circuit has high input impedance and high gain, suitable for microphones used with valve equipment. Small shunt capacitors in the microphone circuit prevent r.f. feedback from the p.a. Microphone gain, adjusted by RV4, should be used sparingly as there is plenty of audio gain in hand with this circuit.

Negative feedback is applied to TR19, by the partially bypassed emitter resistor, giving a rising frequency response to the audio amplifier. The amplifier response may be tailored by changing the value of C62 to suit the characteristics of the microphone used.

The output of the microphone amplifier is coupled to the balanced modulator in the conventional way; relay contacts RLA2 shorting the output on receive, to prevent audio from the microphone circuits being reproduced by the receiver.

Balanced Modulator

The balanced modulator is a standard design, using a matched pair of OA79 diodes D10, D11 in a shunt modulator arrangement. Carrier balance is provided by the carbon track potentiometer PRV4 and the parallel combination of capacitor C162 and trimmer C163.

Transmitter Isolation Amplifier

This stage is included to isolate the balanced modulator output from the input of the common i.f. amplifier during reception. This is necessary in order to break the feedback loop which would otherwise exist around the common i.f. stage. Switching is provided by connecting the emitter of TR20 to the receiver supply line ($-6·8$ V) so that the transistor is reverse biased in the receive condition. Capacitor C68 prevents the transmitted signal from coming in too abruptly and impairing the c.w. keying characteristics of the full break-in position.

First Transmitter Mixer

Drive to the first transmitter mixer is obtained by a special over-wind, close-coupled to the primary of IFT8, details of which are given in **Table 10.23.**

A pair of 2N706 silicon planar transistors TR22, TR23 form a balanced mixer. This has the inherent advantages that the unwanted signal from the v.f.o. is suppressed at the mixer output and the use of two transistors provides a larger output, free from spurious components, than that available from a single-ended mixer. Output from the mixer is coupled into the tunable i.f. section by a bifilar winding L3, ensuring that a good mixer balance can be obtained. Balance is adjusted by the bypassed balance control PRV2, with oscillator injection applied to the slider.

Second Transmitter Mixer

The second transmitter mixer uses the same arrangement as the first. A step-up transformer with tuned primary winding L5 and untuned secondary winding L6 provides adequate voltage to operate the driver valve. The number of turns in L6 for each band maintains substantially constant drive from 80m to 10m, but to conform to licencing conditions the drive is reduced on 160m by use of a series resistance (not shown in the diagram) connected between the hot end of L6 and the 160m position on S1H.

P.A. and Driver Section

The valve p.a. and driver are of conventional design, employing a 12BY7A as a class A driver, bias being derived from the potential divider network R98 and R99. To simplify transmitter tune up, the driver anode coil is resonated by C103 which is ganged to the receiver PRE-SELECTOR control (capacitors C10 and C11) so as to obtain suitable tracking.

The p.a. uses two American colour television line output valves type 6GJ5A, chosen because of their low cost. This combination of p.a. and driver is used since the heaters may conveniently be connected for 12 V operation for mobile use. The transceiver is capable of 200 watts p.e.p. input on all bands, the output falling off at the upper frequencies due to increased circuit losses. As the p.a. is operated in class AB1 the maximum efficiency that can be expected is about 50 per cent, which should be obtained on the lower frequencies.

The p.a. is bridge-neutralized. Capacitor C106 is connected in series with the neutralizing control C105 to isolate it from the 300 V driver anode supply, making adjustment much easier and safer. P.A. cathode current, or r.f. output, is monitored by the combined S-meter and p.a. meter thus providing adequate tune-up facilities. The pi-tank is of standard design except for the capacitors switched in parallel with the P.A. TUNE tuning capacitor on 160m and 80m. This allows the use of a smaller variable capacitor thus conserving space in the p.a. compartment.

If high power is not a prime requirement a suitable transistor p.a. could be substituted for these stages, using one of the many published designs.

Automatic Level Control (A.L.C.)

In accordance with modern transceiver practice automatic level control has been included in this design.

The circuit functions in the following way; r.f. at the low impedance end of the driver coil is rectified by D7 to produce a positive d.c. voltage, which is applied to the top end of VR6, the pre-set ALC LEVEL control. Resistor R100 connected in series with the coupling capacitor C127, prevents the a.l.c. circuitry from upsetting the neutralization of the p.a. stage.

With the ALC LEVEL control set to minimum (slider to earth), bias is fed to the common i.f. amplifier through the gating diode D8. As the control is advanced the bias voltage increases but the positive voltage, due to rectified r.f. from the driver, causes the bias fed to the common i.f. to fall back towards its original value. A.l.c. action causes the low level voice components to be amplified more than the high level peaks and should be used sparingly, as any form of speech compression will always have a detrimental effect on the transmitted quality. It is also possible to over-run the p.a. stage because the average signal power is much greater with a.l.c. in operation. In practice the a.l.c. setting should never exceed one-third of the track of RV6.

If desired the a.l.c. circuitry can be left out of the design by permanently connecting R102 to earth, the remainder of the a.l.c. circuitry being omitted.

For a.m. operation the a.l.c. is disabled by S2B, to prevent the modulation percentage of the a.m. signal being reduced. This circuitry can be omitted if the transceiver is not to operate on a.m.

COMMON STAGES

Conversion Oscillator

The Pierce-Colpitts oscillator circuit chosen has been so arranged that only one switch pole is needed to select the appropriate crystal and the associated resonating capacitor. The 3 ohm carbon resistor R104 across the secondary of the oscillator coupling transformer reduces the effect of loading variations caused by the transmit and receive circuits. However, oscillator frequency pulling may still result if the tuned circuit (L18 in parallel with Cx and trimmer) is not correctly resonated.

Only one 500 kc/s segment of 10m (28·3–28·8 Mc/s) was included in the prototype and it will be necessary to increase the number of BAND switch positions by three, making a total of nine, to give full coverage of 10m. The crystal frequencies required for the extra sections are given in Fig. 10.110. No Cx component is required but a separate trimmer must be used with each crystal.

V.F.O.

The v.f.o. uses the W3JHR 'synthetic rock' circuit which is run at a low supply voltage of −4·7 V stabilized by ZD3. The output from the emitter follower TR28 is fed to an amplifier to provide sufficient drive for the receiver and transmitter mixers. L14 is damped by the output impedance of TR27 giving a substantially constant output across the tuning range. To prevent frequency pulling the secondary winding L15 is damped by a 3 ohm carbon resistor so that any changes in load impedance will be negligible compared with this low value. This resistor also has the function of making the output impedance of the v.f.o. which feeds the transmitter mixer appear more resistive, thus ensuring stable mixer operation.

Some slight pulling of the v.f.o. frequency may still occur between the transmit and receive positions due to earthing effects. Precautions against this are described in the section on constructional considerations.

No attempt has been made to obtain a linear tuning scale as the effort involved was not thought worth while. If this is considered an operating disadvantage then C153 should be replaced by a straight-line frequency law type; padding capacitors may also be required.

Carrier Oscillator

The carrier oscillator uses a standard 465 kc/s single-tuned inter-stage coupling transformer 1FT6, with the low impedance secondary providing the necessary match into the balanced modulator.

The supply voltage to the carrier oscillator, stabilized at 5·6 V by ZD4, is switched off during a.m. reception by S2K. This switch connects the oscillator to the transmit supply line for a.m. transmission and to the common supply line in all other modes. If the transceiver is required only for use on s.s.b. in the normal side-band mode, S2K and S2I may be omitted and the carrier oscillator permanently connected to the common supply line (junction of R125 and pole of RLA1).

Apart from the output which feeds the balanced modulator, an output to the balanced demodulator is provided by the collector of TR30, with carrier insertion for transmission of a.m. provided by RV8, which introduces carrier into the common i.f. stage when the mode switch is in the a.m. position. RV8, C160, R79 and R80 should be omitted if transmission of a.m. is not required.

The carrier oscillator crystal for the normal side-band mode should be chosen to produce a frequency corresponding to the −20db point on the lower side (l.f.) of the mechanical filter selectivity curve (upper sideband generation). If the inverted sideband mode is included, the required crystal should produce a frequency at the corresponding −20db point on the upper (h.f.) side of the filter response.

Tunable I.F.

The first i.f. of 2·5–3 Mc/s is made tunable rather than broad band, to keep spurious responses to a minimum. The transmitter drive is kept constant across the 500 kc/s band by tracking the tunable i.f. across this range.

The combination of tunable i.f. and transmitter and receiver mixers forms the bilateral mixer section. A bilateral mixer is one in which the direction of the signal path may be reversed by switching d.c. supply lines only. This form of operation is achieved by the special coupling coils used in the tunable i.f. section, enabling both transmitter mixer and receiver mixer signal paths to be permanently connected.

Common I.F. Stages 455 kc/s

Standard transceiver practice is to use the same filter arrangement for both the transmitter and receiver sections. Accordingly, the filtering is incorporated in the common i.f. stages and in this design a Kokusai mechanical filter type MF-455-10K provides the response necessary for s.s.b. operation. The filter used in the prototype transceiver has a −6dB bandwidth of 2·7 kc/s. A narrower filter can be used to improve the receiver selectivity but the transmitted signal will be impaired as a result. The bandwidth of the filter to be used is therefore a matter of personal choice.

In planning the transceiver it was decided that it would be advantageous to make provision for both a.m. and c.w. operation as well as the normal s.s.b. modes. Amplitude modulation is usually produced, in a sideband transmitter, by inserting a certain amount of carrier. If this method is used, distortion may become evident unless the modulation depth is carefully controlled. This is overcome by switching out the mechanical filter on a.m. and using top-capacity coupling between IFT9 and IFT7 to produce the bandwidth required for both sidebands. When carrier is introduced into this double sideband signal, normal a.m. is obtained, provided that IFT9 is correctly resonated so that the carrier and sidebands are in the correct phase relationship. Similarly, the reception of a.m. signals is not always possible by the exalted carrier method, so the provision of a wider bandwidth is also advantageous for a.m. reception. In operation the transceiver behaves as a normal a.m. transceiver when switched to the a.m. position.

C.W. reception is enhanced by switching a series-resonant crystal filter in series with the mechanical filter to produce a narrow bandwidth more suitable for this mode. The actual frequency of this crystal is a matter of personal choice, the required crystal frequency is calculated as follows: the series-resonant frequency of the crystal is equal to the carrier oscillator frequency plus the required pitch of the received c.w. signal. Thus, if a pitch of 800 kc/s is required and the carrier oscillator frequency is 453·2 kc/s, then the crystal should have a frequency of 453·2 kc/s + 0·8 kc/s = 454 kc/s. A surplus FT241 type has been used but improved selectivity can be expected with a new crystal, due to the high Q factor of modern hermetically sealed units. If desired, a suitable commercial c.w. filter could be used in place of the crystal, but the performance of the single crystal is adequate for most conditions.

The required filter is selected by poles S2F and S2H of the mode switch, while switches S2G and S2I prevent leakage around the mechanical filter due to the top-capacity coupling used for a.m. Because the filters used for the various modes have different insertion losses, it is necessary to equalize the gain of the common i.f. stage in such a way that the combined amplifier gain and filter insertion loss is constant. Equalization of gain is achieved by switching the value of the emitter resistor with S2E, thus adjusting the biasing conditions of TR21 to give constant gain on all modes.

It is not necessary to include the complication of switchable filters if a basic s.s.b. transceiver is required: S2E, S2F, S2H and S2I may be omitted, leaving the mechanical filter permanently in circuit.

In this design, the changeover from transmit to receive is, as far as possible, effected by switching d.c. supply lines, keeping signal switching to a minimum. This requirement has been satisfied for the tunable i.f. section by using the primary of IFT9 to provide the transmitter signal path. The receiver signal is fed from the collector of the receiver second mixer to the tap on the secondary of IFT9. This is a normal double-tuned transistor i.f. transformer and results in a mismatch between TR5 and IFT9; in practice this compromise arrangement will be found to be acceptable.

To produce a transformer suitable for matching the output of the common i.f. amplifier into the first transmitter mixer a standard double-tuned transformer is modified by the addition of a bifilar overwind, details of which are given in Table 10.23

SPECIAL FACILITIES

C.W. Tone Oscillator and Keying Circuits

When c.w. operation is incorporated into sideband equipment, some of the inherent features of the system can be used to good advantage. The problem of netting with simple c.w. transmitters is overcome by using a tone oscillator to produce the same pitch as that used for reception. This is possible with an s.s.b. transmitter, because a keyed tone will produce a pure c.w. signal, rather than the m.c.w. signal which would result if this technique were applied to an a.m. transmitter. An additional advantage gained by the use of a tone oscillator is obtained by using the output to provide side-tone facility.

The tone oscillator uses a single OC45 in a phase-shift circuit which must be adjusted in frequency to correspond with the pitch chosen for the received signal, thus making netting unnecessary. $\left(\text{Note frequency} = \frac{1}{2\pi CR\sqrt{6}}\right)$

Keying a phase-shift oscillator of this type can produce a very poor keying characteristic: a separate keying circuit is therefore used which effectively shorts the output during key-up conditions. The coupling network, C43 and R41, prevents the oscillator from being too heavily loaded when TR11 is shorting the output. The action of the keying circuit is simple; an OC45 is made either conducting or non-conducting, by switching the base between negative or positive voltage with respect to the emitter. Keying characteristics are shaped by C41 which may be adjusted to satisfy individual requirements. The shaping network will be fully effective only when the mode switch is set to either the VOX or MOX positions. In the PTT position the signal will be a little sharper due to the action of the transmit/receive relay.

Three different modes of c.w. operation are possible with this design, firstly, by using the function switch in the PTT position, the transmit/receive relay is keyed directly via D15, providing full break-in operation. Secondly, in the VOX position, a delayed form of break-in operation can be used, the time constant being decreased on c.w., as the normal VOX delay is rather too slow for average keying speeds. Finally, to prevent the VOX relay dropping out during transmission, use can be made of the MOX position which will produce a perfect c.w. signal.

If c.w. operation is not required, the c.w. oscillator and keying circuit can be omitted, or replaced by a two-tone oscillator for tune-up purposes.

Independent Receiver and Transmitter Tuning

Independent receiver tuning is a facility which enables the receiver to be tuned over a small frequency range of ± 5 kc/s, without altering the transmitter frequency. This enables the operator to follow a drifting signal or to check either side of the transmitting frequency. By switching over to the independent transmitter tuning position the transmitter will be on the frequency previously set by the independent receiver tuning position.

The action of this circuit relies on the voltage-variable capacitance nature of a silicon junction diode. The diode, VCD1, is operated in its reverse bias condition, the depletion layer capacitance decreasing with increased reverse bias voltage. The variable capacitance action produced by the bias control voltage is applied in parallel with the v.f.o. tuned circuit by C156. If desired the value (nominally 22 pF) of this capacitor may be increased, to give a greater range of frequency deviation. As the capacitance change follows a reciprocal square-root law, a linear track potentiometer controlling the bias would give a non-linear scale on the IRT/ITT control. An improvement in linearity is obtained by the use of a logarithmic potentiometer; and this, while not giving a perfectly linear scale, is quite satisfactory in practice.

D.c. switching to the i.r.t./i.t.t. circuit is obtained directly from the switched transmit and receive supply lines, and by using D9 as a switching diode the need for extra relay contacts is avoided. In the i.r.t./i.t.t. "OFF" position the diode is fixed-biased by the potential divider R119 and R120, this voltage being adjusted to give a frequency approximately central with the maximum deviation produced by RV7 at each end of its track.

If the extra complication of i.r.t./i.t.t. circuitry is not felt worth while, all the i.r.t./i.t.t. circuitry, including C156, may be omitted together with the connections to S3A and S3B.

VOX Circuitry

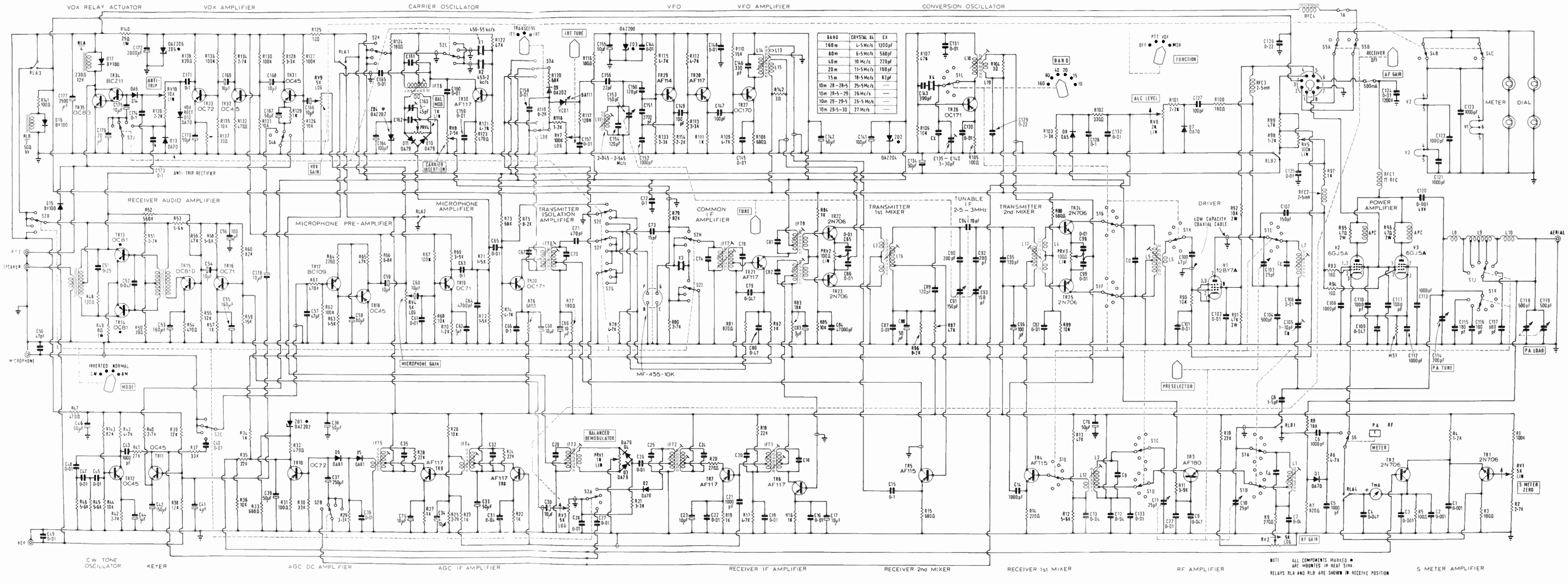
The VOX circuit comprises three stages of amplification, TR31, TR32 and TR33, with TR31 being disabled by S4A when the function switch is not in the VOX position. The input to TR31 is taken from the top of the transmitter pre-set MIC GAIN control, so that the VOX sensitivity is not affected by the setting of this control. The VOX sensitivity is pre-set by the VOX GAIN control RV9 (which adjusts the audio level fed to TR31).

The audio output from TR33 is rectified by the VOX rectifier D12 to produce a negative d.c. voltage at the bottom end of the pre-set ANTI-TRIP control RV10. The anti-trip rectifier D13 produces a positive d.c. voltage at the bottom end of RV10, derived from the audio amplifier. Adjustment of the ANTI-TRIP control RV10 ensures that the correct percentage of negative voltage from the VOX amplifier, and positive voltage from the anti-trip circuit, is fed to the VOX relay actuator to obtain reliable operation. Drive to the relay actuator is applied by the gating diode D14, which prevents the VOX time-constant capacitor C175 from discharging back into the drive circuit. The attack time of the VOX circuit is dependent on the charging time of this capacitor, which should therefore have as small a value as possible, consistent with obtaining the correct delayed VOX action.

To energise the relay, the Darlington Pair configuration, chosen for its high input impedance, allows the use of a small capacitor, which may be varied to suit individual requirements.

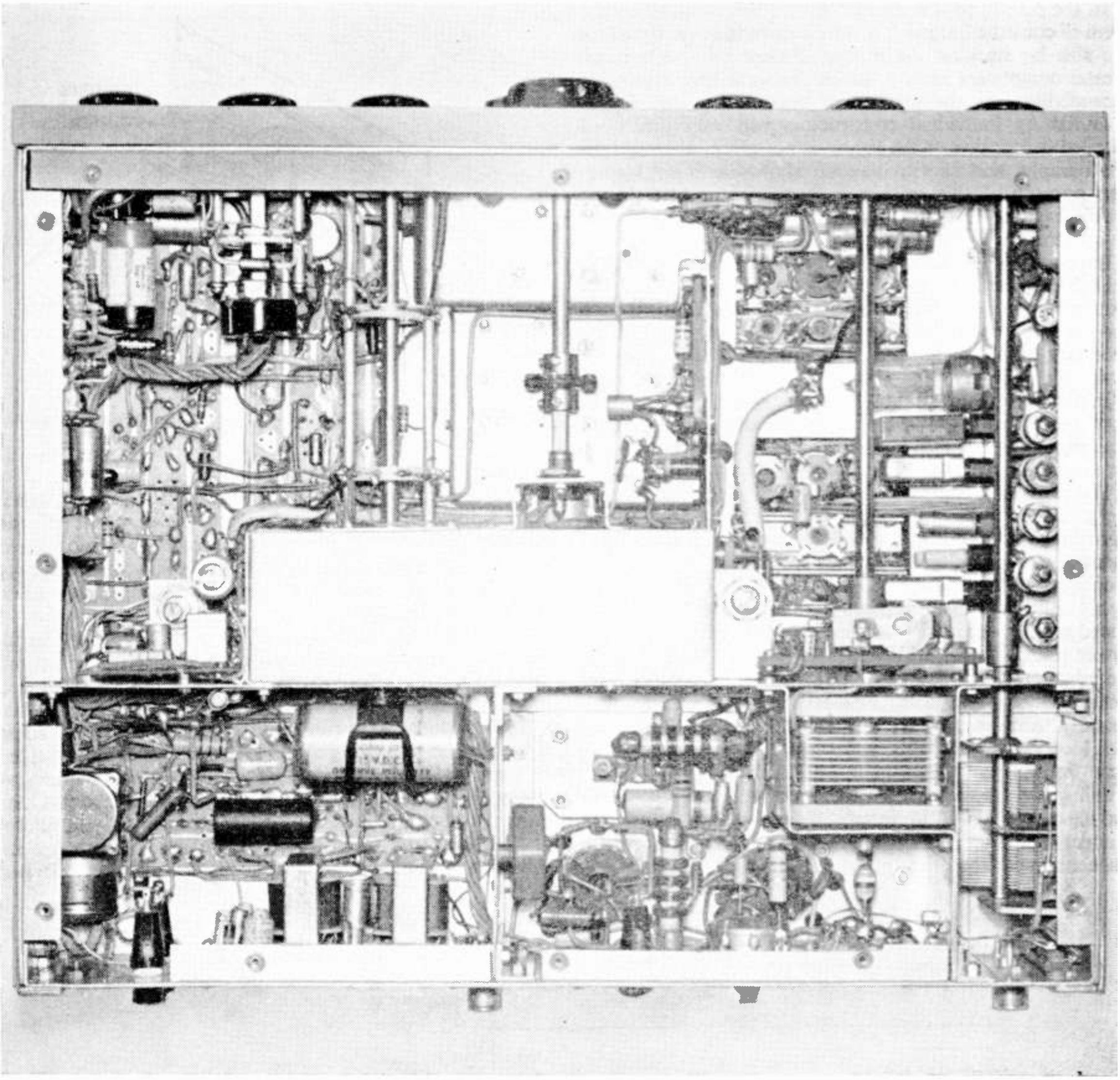
For c.w. the switch S2J shorts out the anti-trip and at the same time connects C176 in series with C175 to reduce the VOX time-constant, making it more suitable for c.w. operation. C176 may be adjusted to suit individual keying speeds. If the c.w. facility is not included, then switch S2J can be left out and C175 permanently connected.

It is essential that TR34 is a silicon p.n.p. transistor because the leakage current (Icbo) of a germanium transistor would turn the relay drive transistor TR35 on continuously.



BAND	CRYSTAL X4	CX
160m	4·5Mc/s	1200pF
80m	6·5Mc/s	560pF
40m	10 Mc/s	220pF
20m	11·5Mc/s	180pF
15m	18·5Mc/s	82pF
10m 28–28·5	25·5Mc/s	—
10m 28·5–29	26Mc/s	—
10m 29–29·5	26·5Mc/s	—
10m 29·5–30	27Mc/s	—

Fig. 10.112. Circuit diagram of the G3OQD transistorised SSB transceiver. Coil winding details are given in Table 10.23 and the method of coil assembly on page 10.96. In this diagram R174 (180Ω) should be connected between S2K moving arm and the junction of C164, ZD4, which is also the feed point for the regulated supply to the carrier oscillator.



Transceiver underchassis layout. Note the extensive screening round the i.f. and r.f. sections and accessibility of component boards mounted over the cut-outs in the chassis. Boards can be identified from the annotated top chassis diagram, Fig. 10.115.

Diodes D17 and D16 are connected across the relays RLA and RLB respectively as a precautionary measure in case voltage spikes due to the back e.m.f. across the relay coils exceed the breakdown voltage of the transistors.

Two relays perform transmit/receive switching, relay RLB being driven from contacts on relay RLA, which also provide a 12v supply on transmit for switching any relays in ancillary equipment e.g. power supplies, linear amplifiers.

If VOX is not considered a necessity for s.s.b. operation, this part of the circuitry can be omitted by leaving out all components between, and including TR31 to TR35, but retaining R140, C172, ZD5 and R125.

CONSTRUCTION

Considerable scope for modification and future development is provided by the use of Radio Spares printed circuit boards which have positions for eight transistors with connecting strips and triangular component mounting leads.

The completed boards are mounted into cut-outs made in a standard 11 in. × 14 in. aluminium chassis thus making both sides of the printed circuits fully accessible. The only panels not mounted in this way are the bilateral mixer panel, receiver front end and audio amplifier sections, which are mounted vertically to conserve chassis space (see pages 10.98 and 10.99).

In the prototype the transceiver size was dictated by the form of construction used; it is thus quite in order to reduce the size by stacking the printed circuits to give a much greater component density, but at the same time sacrificing accessibility. As the method of construction and layout favoured by individual constructors will vary greatly, a detailed description of the layout is not given, although the photographs and sketch diagram show the basic layout. The purpose of this section is to give some idea of the problems likely to be encountered and how they may best be overcome.

Instability is as much a problem with transistors as with valves; however, the cause is somewhat different, the important criterion being to prevent any common impedances existing between the separate stages. It may therefore be necessary to add some extra decoupling not shown on the circuit diagram, taking care that this does not impair the fast transmit/receive capabilities of the transceiver. All the extra decoupling should be confined to the main supply lines, thus providing the low impedance feed necessary to keep time constants short.

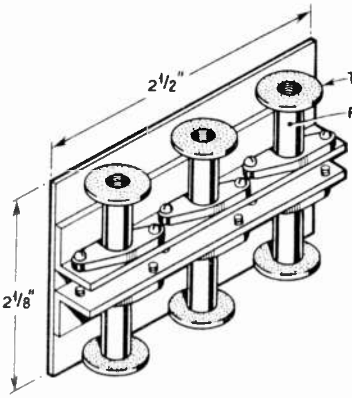


Fig. 10.114. Method of mounting coils on the switch screening plate

The receiver r.f. amplifier may require special attention so far as its decoupling arrangements are concerned. It is advantageous to make up the value of C9 from several separate capacitors earthed down to different points, also keeping the base lead of TR3 reasonably short (not greater than ½ in.). Lack of gain at the higher frequencies can often be traced to poor decoupling of the earthy side of the input and output tuned circuits, which may be remedied by making up the values of C7 and C12 from a number of smaller capacitors, again earthed to separate points. Instability can be particularly troublesome in the 455 kc/s i.f. stages due to the high loop gain produced by so many stages operating on the same frequency; this can be overcome by checking the decoupling and earthing, with particular attention to the mixer and a.g.c. sections. If i.f. instability persists, it is in order to connect damping resistors across the i.f. transformers associated with the offending stages; the loss in gain is not serious as the gain of the i.f. strip is on the high side. Selectivity will not be affected because the response of the mechanical filter is not altered by the response of subsequent stages. A particular form of audio instability, or hum, may result if the audio panel is not earthed at a single point close to the amplifier input.

The above points should make it possible to clear up all instability problems which might arise in the receiver section, but one of the inherent drawbacks with a transistorized transceiver is the proximity of the p.a. stage to the transistorized sections. This can result in r.f. from the p.a. being picked up on the wiring of the low level stages and rectified by the non-linear action of the transistors. Decoupling by electrolytics is not efficient at r.f. and such components may require bypassing with a small ceramic capacitor. The photographs show the extensive screening used to prevent r.f. radiation from the p.a. getting back into the transistor stages.

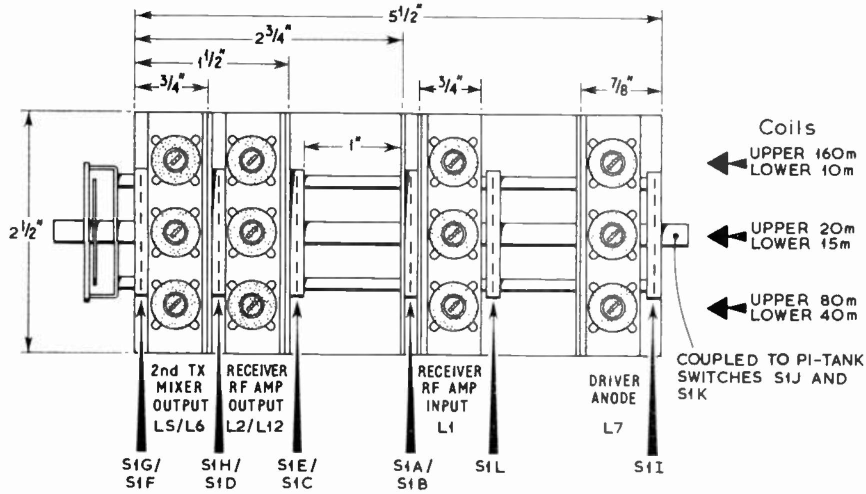


Fig. 10.113. Coil assembly showing layout of coils and switch wafers.

TABLE 10.23
COIL WINDING DETAILS

Component	BAND					
	160m	80m	40m	20m	15m	10m
L1 L1 tap (from earthy end CA	96 turns, 38 s.w.g., 14 turns 150 pF	63 turns, 36 s.w.g., 8 turns 82 pF	37 turns, 30 s.w.g., 4 turns 22 pF	21 turns, 26 s.w.g., 2 turns —	15 turns, 22 s.w.g., 2 turns —	10 turns, 20 s.w.g., 1 turn —
L2 L2 tap (from earthy end) CB L12	96 turns, 38 s.w.g. 65 turns 150 pF 14 turns, 36 s.w.g.	63 turns, 36 s.w.g. 35 turns 82 pF 8 turns, 36 s.w.g.	37 turns, 30 s.w.g. 19 turns 22 pF 2 turns, 24 s.w.g., (pvc covered)	21 turns, 26 s.w.g. 10 turns — 2 turns, 24 s.w.g., (pvc covered)	15 turns, 22 s.w.g. 6 turns — 2 turns, 24 s.w.g., (pvc covered)	10 turns, 20 s.w.g. 4 turns — 2 turns, 24 s.w.g., (pvc covered)
L5 (bifilar) CD L6 on centre of L5	60 turns + 60 turns, 38 s.w.g. 47 pF 10 turns, 38 s.w.g.	32 turns + 32 turns, 36 s.w.g. 33 pF 6 turns, 36 s.w.g.	19 turns + 19 turns, 32 s.w.g. — 22 turns, 32 s.w.g.	12 turns + 12 turns, 28 s.w.g. — 12 turns, 28 s.w.g.	9 turns + 9 turns, 26 s.w.g. — 9 turns, 26 s.w.g.	6 turns + 6 turns, 20 s.w.g. — 9 turns, 30 s.w.g.
L7 CE	100 turns, 38 s.w.g. 68 pF	63 turns, 36 s.w.g. 33 pF	40 turns, 32 s.w.g. —	16 turns, 26 s.w.g. —	10 turns, 26 s.w.g. —	6 turns, 22 s.w.g. —
X 4 CX (silver mica)	4·5 Mc/s 1200 pF	6·5 Mc/s 560 pF	10 Mc/s 220 pF	11·5 Mc/s 180 pF	18·5 Mc/s 82 pF	25·8 (28·3–28·8 Mc/s) —
L18	10 turns, 24 s.w.g.		L19, 2 turns (24 s.w.g., p.v.c. covered) at earthy end			
L14	60 turns, 36 s.w.g.		L13, 1 turn (24 s.w.g., p.v.c. covered) at earthy end. L15, 8 turns, 36 s.w.g. at earthy end			
L4 (bifilar)	4 turns + 4 turns, 24 s.w.g. p.v.c. covered		L17, 50 turns, 36 s.w.g., tapped at 20 turns from earthy end			
L3 (bifilar)	12 turns + 12 turns, 36 s.w.g.		L16, 50 turns, 36 s.w.g., tapped at 3 turns from earthy end			
L11	35 turns, 24 s.w.g., ½ in. diam. ceramic former with core					
IFT1, 2, 7, 9	Double-tuned. Not pot core type					
IFT3, 4, 6	Denco Type IFT14		IFT5 Denco Type IFT13			
IFT 8	Double tuned with 8 turns + 8 turns bifilar close-coupled to primary					
L8	4 turns, 14 s.w.g., 1 in. diam., ½ in. long, air-spaced					
L9	2 turns, 16 s.w.g., 3 turns, 8 turns, 10 turns, 18 s.w.g., 1½ in. ceramic or polystyrene rods, 2 in. long, 3/16 in. between segments					
L10	28 turns, 22 s.w.g., 1 in. diam. ceramic or polystyrene rod, ⅞ in. long					
RFC 1	10, 15, 25, 50, 150 turns, 36 s.w.g. with ⅛ in. between segments on 4 in. long, ¾ in. diam. ceramic former					

NOTE 1—Except where stated, coils are close-wound on Aladin 3/16 in. diam. former with suitable slug.
2—RFC 4:—LA2 pot core wound full with 22 s.w.g. Formvar.
3—APC:—6 turns 18 s.w.g. wound on 47 Ω 2W carbon resistor.

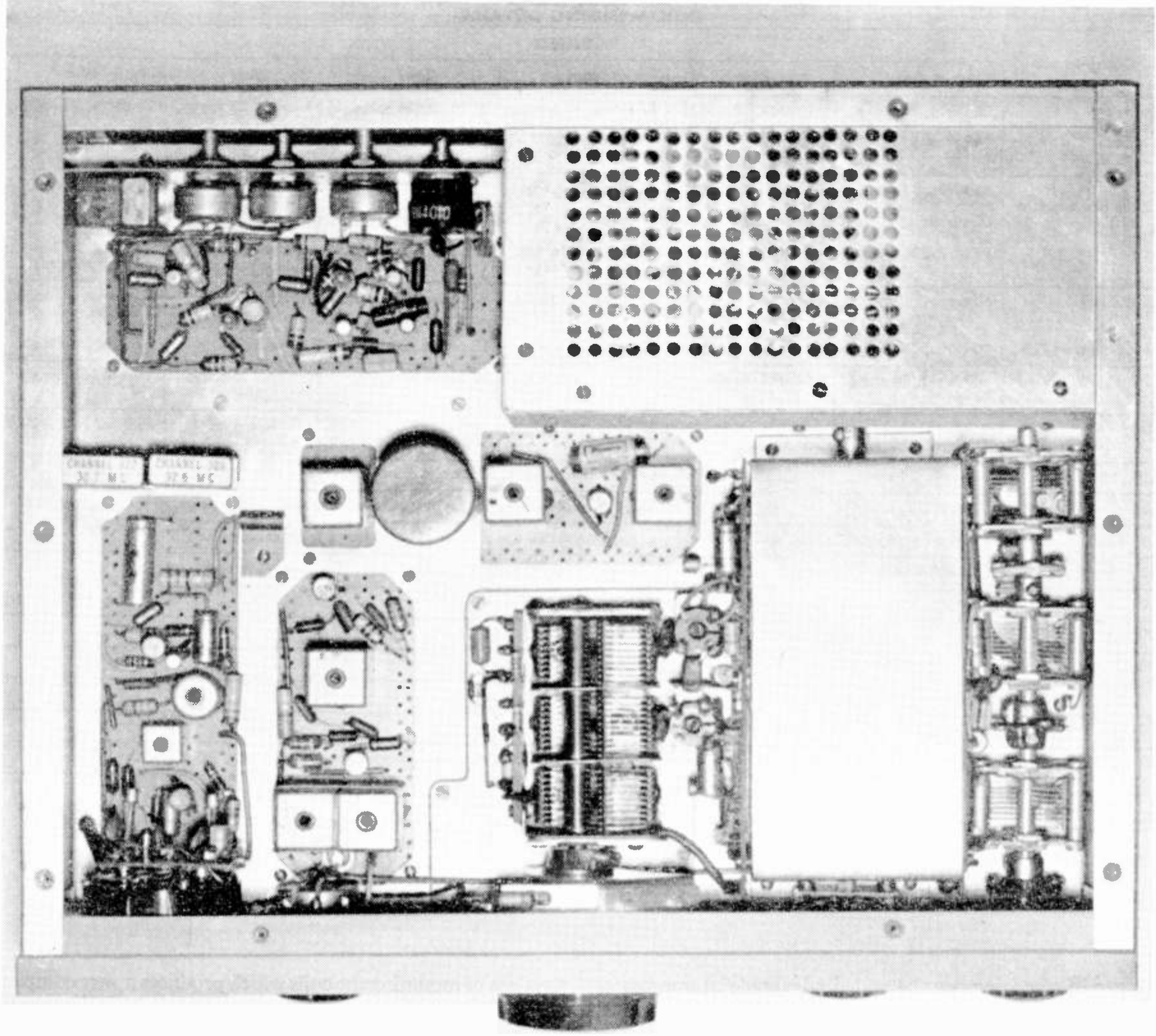
The use of screening has the added advantage of improving the overall rigidity; this being a most important factor if the transceiver is to be subjected to the vibration encountered in mobile operation. Many otherwise very stable v.f.o.'s are marred by relative movement between the frequency determining components of the oscillator due to stress in the chassis, causing frequency shifting. The v.f.o. should therefore be constructed in a diecast box of sufficient dimensions to give a clearance around the coil at least equal to its own diameter, or a smaller diameter coil former used. Relative movement between the tuning capacitor and the remainder of the v.f.o. circuitry is prevented by mounting the tuning gang on a rigid ⅛ in. aluminium panel screwed to the v.f.o. box, with the chassis sandwiched between. It is most important to have the v.f.o. box well earthed and fully screened to prevent any frequency pulling, caused by differing earth currents in the chassis, produced when changing from receive to transmit. As the output of the transistor carrier oscillator is particularly rich in harmonics, some of which will inevitably fall in the amateur bands, it is most desirable to have this stage fully screened. The screening performs a dual role by also preventing carrier leakage into the 455 kc/s i.f. stages.

The coil assembly **Fig. 10.113** used in the prototype employs a system of mounting the coils which produces a very compact unit. Each set of six coils is mounted on two 2½ in. lengths of ½ in. aluminium angle bolted to the screening plate associated with the switch wafer. The diagram in **Fig. 10.114** shows how one such coil set is constructed, with three of the six coils accessible from the top of the chassis, the others being adjusted from the underside.

Component Selection

This section indicates where component choice is critical and where substitutions are possible.

In general, none of the transistor types are particularly critical, in fact OC171's can be used for all the germanium transistors used in r.f. applications, although the use of an AF180 for the RF stage produces slightly better results at the upper frequencies. The 2N706's used for the balanced mixers can be replaced by 2N706A, 2N708, BSY38 or equivalents. A suitable alternative for the BC109 is a 2N2926 (green) or BC107, BC108. The other transistors used for audio applications are also not critical and any transistor or diode used in the design can be replaced by an equivalent the suitability of which may be verified by consulting manufacturers' published data.



Transceiver top chassis layout. This should be compared with the detailed layout diagram Fig. 10.115 on the opposite page.

The i.f. transformers chosen were surplus items, of conventional construction (not pot core type). If pot core types are to be used it may be necessary to damp the Q of the windings with a parallel resistance to prevent reduction of the i.f. bandwidth produced by the mechanical filter. A drawback with this type of i.f. transformer is that it is not suitable for addition of overwinds as required by IFT8.

All potentiometers are of carbon track construction, PRV on the circuit diagram indicating the use of miniature preset types.

The switches are almost entirely of the Radio Spares type, the exceptions being the wafers used for pi-tank switching which should be of ceramic construction. The Radio Spares switches are not the most suitable for r.f. applications because the large stray capacitance between the contacts can cause sucking from an in-circuit coil by one out of circuit. If a better switch is available this should be used in preference.

The relays used to perform the transmit/receive function must have a short operate time, particularly if full break-in operation is to be used. The relays used in the prototype are of the miniature " Siemens " type.

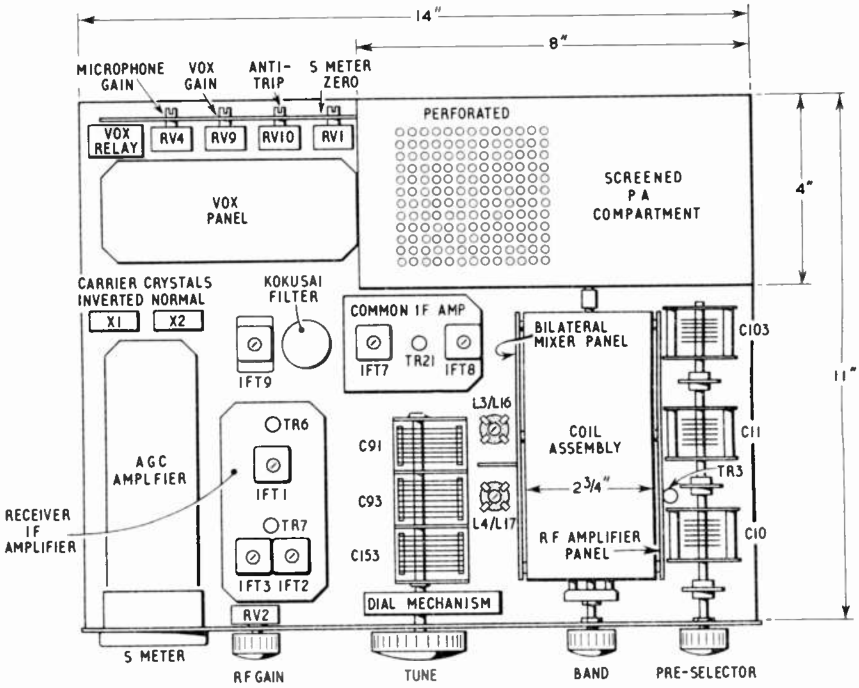


Fig. 10.115. Above chassis layout. Cut-outs are made in the chassis to mount the VOX panel and the a.g.c. amplifier, receiver i.f. amplifier and common i.f. amplifier boards. The bilateral mixer and amplifier panels are mounted vertically.

If substitution of the mechanical filter is contemplated, it should be noted that the transducer resonating capacitors are contained within the Kokusai unit to give a high input and output impedance. This is not usually the case with other makes of filter (e.g. Collins types) and external resonating capacitors must, therefore, be used as indicated in the manufacturers' instructions.

All the variable capacitors were of surplus origin. The pi-tank P.A. TUNE capacitor must have a suitably wide spacing to prevent flash-over, but the LOAD capacitor may be a standard double-gang broadcast type. A three-gang 25 pF per-section capacitor can be used for the PRE-SELECTOR control in place of the separate capacitors (ganged by insulating flexible couplers in the prototype), provided that earthy side of C103 is taken to earth instead of the junction of C104 and L7.

The 2⅜ in. square illuminated S meter is of Japanese manufacture with the lower part of scale calibrated 0–300 mA for measuring p.a. cathode current; any meter with a basic movement of 0–1 mA f.s.d. can be made suitable by recalibration of the scale in S units and mA as required.

A suitable dial assembly can be fabricated (with some ingenuity) using, as a basis a Muirhead epicyclic drive mechanism as found in the RF26, RF27 and other surplus equipments.

It is hardly necessary to point out that the final performance may be impaired if inferior components are used, some cheap electrolytic capacitors were found to go low in capacity after only a short amount of use. Mullard capacitors have given no trouble and are therefore recommended.

ALIGNMENT AND FINAL ADJUSTMENTS

Failure to obtain satisfactory results from a completed equipment is usually due to incorrect alignment and adjustment. This design, although complex, should not present any unduly difficult alignment problems, even if little test equipment is available. It is, however, desirable to have a multimeter (20,000Ω/V) and a signal generator (BC221) for calibration purposes; a valve voltmeter is also useful (see Table 10.24) but by no means essential.

Before alignment of the tuned circuits can be commenced, the d.c. operating conditions of the transistors must be correct and this should be the case, provided that the wiring is free of errors and faulty components.

As a guide to constructors new to transistors, a simple check can be made by measuring the base-emitter voltage which should be about −0·2V for germanium transistors and approximately 0·5V for silicon. Also, the Zener diode voltages (and currents) can be checked. It should be noted that the base-emitter voltage of the mixers must be measured with the oscillator drive disabled, otherwise false readings will be obtained. This effect can, however, be used to advantage by checking the emitter voltage of TR4, which should be found to vary between −0·16V on 10m to −0·08V on 160m, indicating correct oscillator injection.

The biasing of the receiver audio output transistors, TR13, TR14, may require adjustment to prevent cross-over distortion or over running—this is set by the potential divider R50 and R51.

The d.c. voltage at the junction of C78 and C79 (base of

TR21) must be checked. This is the a.g.c. control voltage and is measured with no input to the receiver and with the PRESELECTOR control detuned; an approximate reading of -1V should be obtained.

All the oscillators can be checked and the v.f.o. adjusted for correct coverage (2·045 Mc/s to 2·545 Mc/s) by using the main station receiver; it may be necessary to listen to harmonics of the carrier oscillator if coverage of 455 kc/s is not available.

Alignment should commence at the receiver back-end (455 kc/s i.f.) with the mode switch in the normal sideband position, input signal being provided by either a signal generator set to the centre frequency of the mechanical filter, or an aerial may be loosely coupled to the secondary of IFT9.

The i.f. transformers IFT1, 2, 3, 7, 8 and 9 may now be adjusted to obtain maximum output; precise alignment is not important at this stage. The a.g.c. amplifier may also be aligned by adjusting IFT4 and IFT5 for maximum S meter reading. Alignment may now proceed by adjusting the tunable i.f. (L16 and L17) to resonance, which may be achieved with an aerial loosely coupled to L17 if a signal generator is not available to provide the necessary signal.

It is now possible to adjust the r.f. coils L1 and L2 for maximum S meter reading on received signals, but use of a signal generator will prevent peaking the coils on the image frequency. Difficulty with aligning the higher frequency coils may be due to damping caused by adjacent metal work being too close. Alternatively, it may be due to suck-out as mentioned in the constructional section, in which case the LC ratio of the offending coil may be changed slightly.

Stray capacities will vary in individual layouts and may necessitate minor adjustments to the turns on the tuned windings particularly on 15 and 10m. This may also be required in the tuned circuits associated with the transmitter.

Before power is applied to the transmitter section it is important to adjust PRV2 and PRV3 to mid-track. The transmitter may now be roughly aligned, commencing with the balanced modulator. The value of C162 must be determined experimentally (about 50 pF) and the parallel combination of this capacitor and C163 should be tried on both sides of PRV4 to obtain a satisfactory balance. As the common i.f. stage has been aligned, it is necessary only to adjust the primary of IFT9 in order to obtain a 455 kc/s s.s.b. signal at IFT8. The signal may be monitored at the tunable i.f. The v.f.o. output coil L14 should be adjusted for resonance at the centre of the tuning range, and PRV2 adjusted for minimum v.f.o. signal appearing at the output of the mixer (do not set PRV2 to either end of the track: this causes excessive current in one of the mixer transistors).

By loosely coupling the station receiver to L6 it should now be possible to hear the signal on each amateur band, L5 being adjusted for maximum output in each case. PRV3 should be adjusted on 10m for minimum conversion oscillator signal appearing at the output of the second mixer, taking the same precaution as mentioned for PRV2.

Power may now be applied to the driver valve V1 and the anode coil L7 adjusted to resonance, having previously set the preselector control for reception. This adjustment must be repeated for each band to ensure correct tracking.

Before power is applied to the p.a. the meter shunt MS1 must be adjusted by applying a d.c. supply with variable series resistance to the cathodes of the p.a. valves, the current being monitored with an accurate test meter and the shunt adjusted so that the p.a. cathode current meter reads full scale when the test meter shows 300 mA. Incorrect calibration will result if the shunt is adjusted with the meter out of circuit.

With power applied to the p.a. the bias control RV5 should be adjusted for a standing cathode current of 35 mA. Final p.a. adjustment consists of setting the neutralizing capacitor C105 (on 10m) so that the dip in anode current corresponds with maximum r.f. output. This condition should hold good for all bands without having to readjust C105. If this is not the case then some form of feedback other than that due to anode-to-grid capacity must be present. The layout and wiring should therefore be checked for stray capacitive or inductive feedback which varies with band changing.

TABLE 10.24
R.M.S. CHECK VOLTAGES USING VALVE VOLTMETER

V.F.O. Output (TR27 collector) 0·6 V
Carrier Oscillator (TR30) output 2·3 V

Band	Driver Grid	Driver Anode	Conversion Oscillator Collector
160m	0·4 V	25 V	1·3 V
80m	1·3 V	45 V	1·3 V
40m	3·5 V	45 V	2·0 V
20m	4·0 V	50 V	2·3 V
15m	3·0 V	40 V	3·0 V
10m	3·3 V	35 V	4·0 V
	Re-resonate coils when valve voltmeter connected		

S.S.B. measurements are made with 1·5 kc/s audio input to the microphone socket and the level set to give maximum output.

The transceiver should now be operational and final adjustments can be made to give optimum performance. With a signal applied at the centre frequency of the mechanical filter, the 455 kc/s i.f. strip may now be accurately adjusted by tuning IFT 1, 2, 7 and 8 for maximum S meter reading. A convenient signal is provided by tuning the receiver to the output from a calibration oscillator etc., so that the same pitch is heard in both normal and inverted sideband positions of the mode switch. Maintaining this setting and switching to a.m., the secondary of IFT9 should be adjusted for maximum S meter reading on this signal.

For correct a.m. transmission the pre-set CARRIER INSERTION control is adjusted on 80m to drive the p.a. to a cathode current of 100 mA and the primary of IFT9 adjusted to obtain undistorted a.m. An oscilloscope connected to the transmitter output will assist in this operation although good results can be obtained by monitoring on the station receiver. Since the a.m. drive is set on 80m there will be a progressive fall-off at the higher frequencies; operators wishing to obtain full input power on a.m. may incorporate an RF DRIVE control in the cathode of V1, using a 10KΩ potentiometer decoupled by a 0·1 μF capacitor. If this modification is adopted PRV8 is set on 10m and the drive control turned down to prevent over-driving the p.a. on the lower bands.

To ensure accurate netting, it is important to check that the v.f.o. and conversion oscillator frequencies are not being pulled due to the change from receive to transmit. This may conveniently be checked by monitoring both the v.f.o. and conversion oscillator frequencies while changing from receive to transmit.

Calibration of the tuning dial and IRT/ITT control may be accomplished by the use of a BC221 frequency meter. It should be noted that the 1RT/1TT calibration will be reversed when changing from, say 80m to 20m.

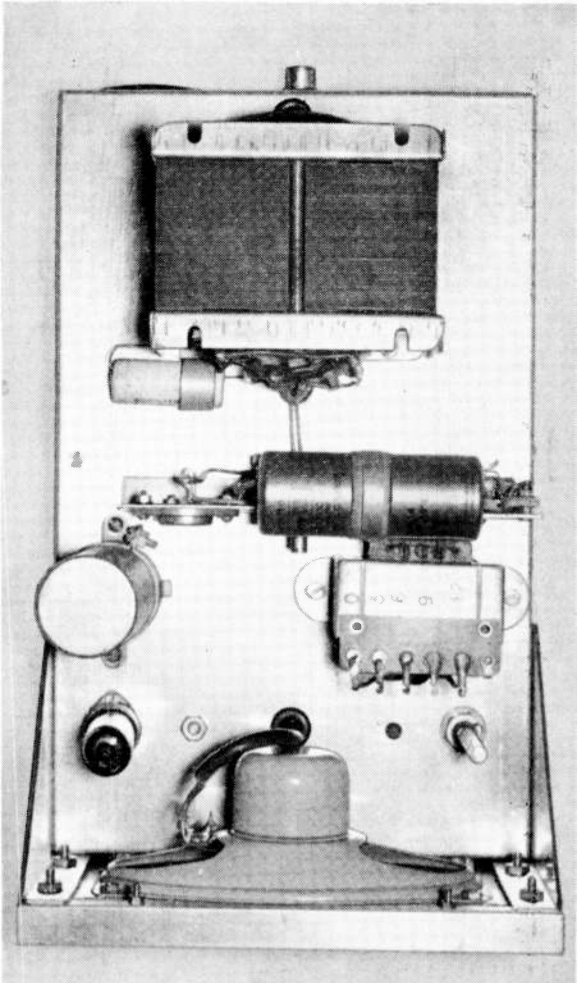
A.C. Mains Power Supply

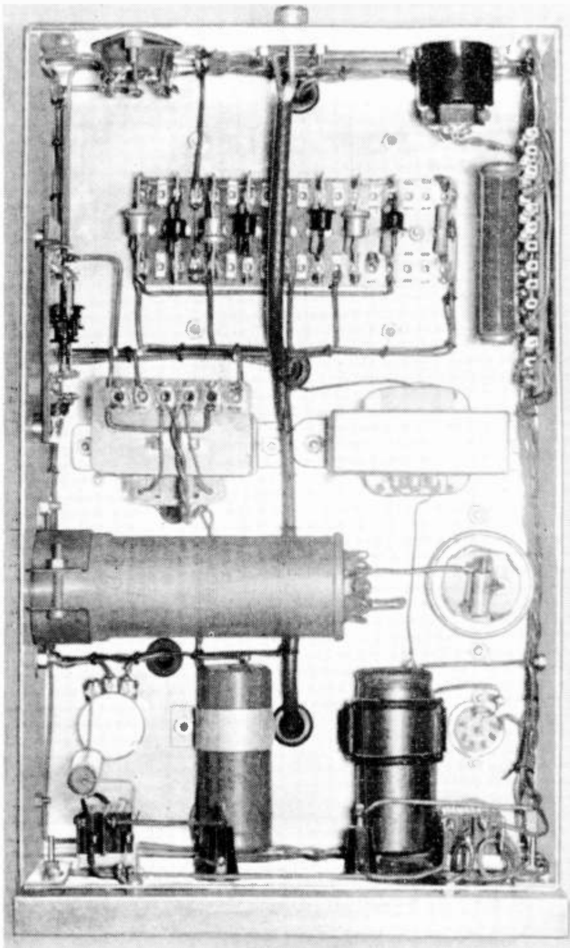
The power requirements of the transceiver are:

750 V @ 300 mA peak, for p.a. anode supply.
12 V @ 2A for heater supply (including dial lights; 1·5A for valves alone).
−100 V @ 3 mA for p.a. and driver bias.
210 V @ 20 mA regulated, for p.a. screens.
300 V @ 50 mA for driver, anode and screen supply.
−12 V @ 150 mA regulated, for transistor stages.

For home station operation the transceiver is powered from a conventional a.c. power unit which meets the above requirements. The circuit is shown in **Fig. 10.116**. All voltage requirements except the stabilized 12 V supply are provided by a standard transformer. Any suitable transformer may be used provided that the power unit is capable of producing the required outputs.

High voltage supply for the p.a. valves is obtained by bridge rectification of the 600 V a.c. output. No smoothing choke is used as this would ruin the dynamic regulation needed for the fluctuating demands of a linear p.a. used for s.s.b. The voltage available from a capacitor input filter is approximately 1·4 times the a.c. input voltage during off load





Top (left) and underchassis (right) views of the a.c. power supply unit which also houses the loudspeaker.

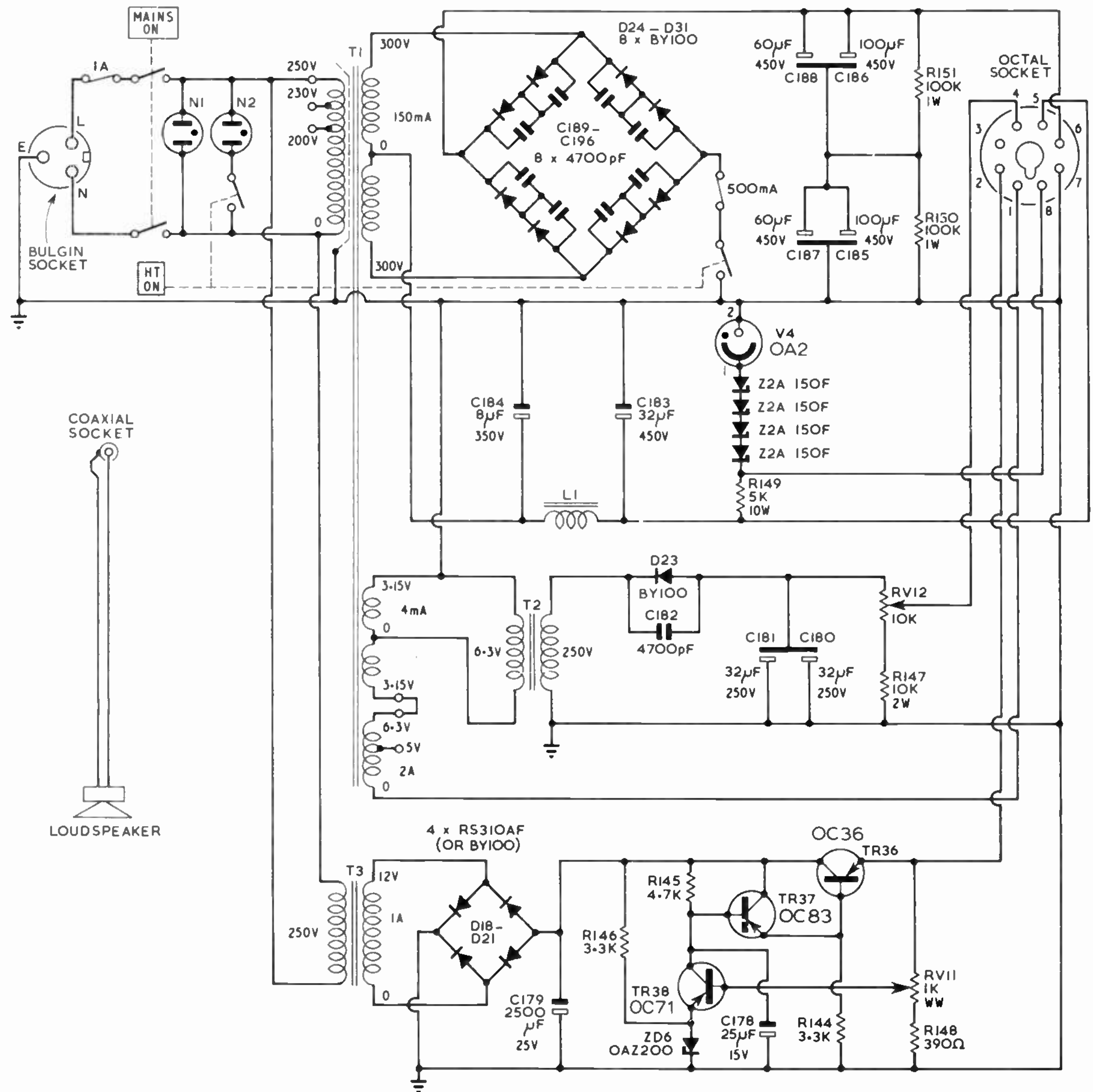


Fig. 10.116. A.C. power supply. T1 is a Douglas MT33AT transformer (300—0—300V, 150mA: 3·15—0—3·15V, 4A: 0—5—6·3V, 2A). T2 (6·3V, 1A) and T3 (12V, 1A) are small heater transformers. The smoothing choke L1 is rated at 10H, 60mA. Indication of "mains on" and "h.t. on" conditions are given by the 250v a.c. neon lamps N1 and N2 respectively. RV11 should be set to provide 12 V d.c. on load, at the emitter of TR36.

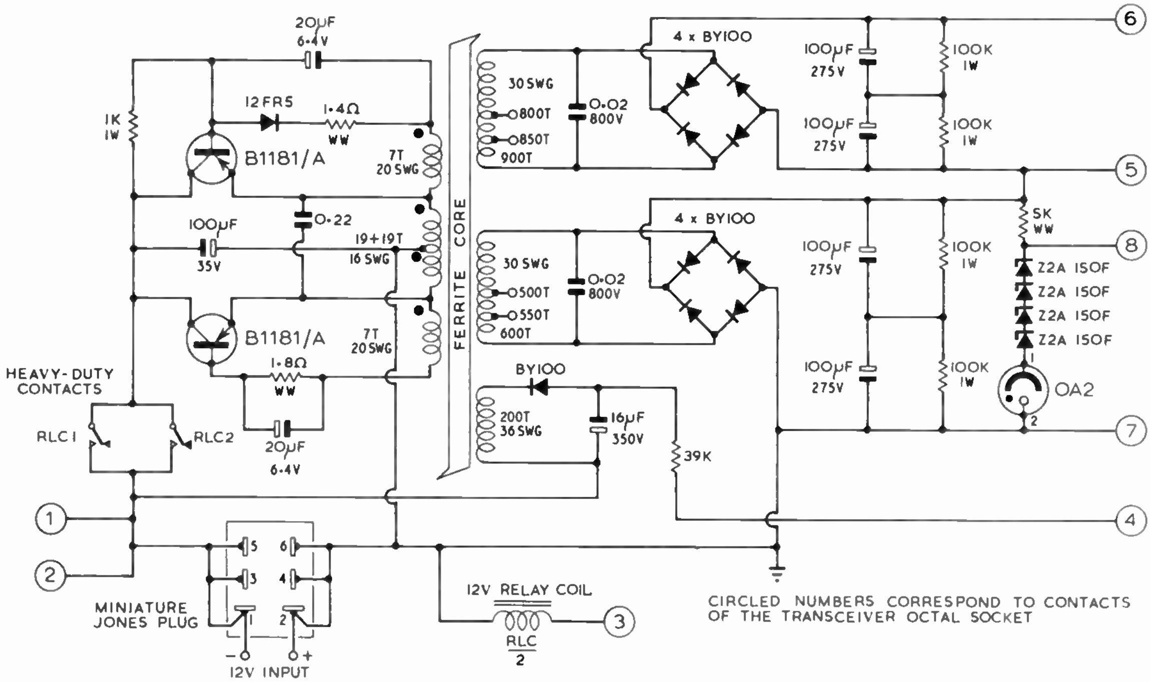


Fig. 10.117. D.C. to D.C. converter for mobile operation. The transformer windings are all of Formvar wire, each layer being covered with thin plastic tape. The primary winding of 19 + 19 turns, 16 s.w.g., bifilar, and the feedback winding of 7 + 7 turns, 20 s.w.g., bifilar, are phased as shown on the circuit diagram to form a continuous winding. Secondary winding details are as follows: 450V winding, 900 turns, 30 s.w.g., tapped at 800 and 850 turns wound on half of ferrite core: 300V winding, 600 turns, 30 s.w.g. tapped at 500 and 550 turns. This and the bias winding of 200 turns, 36 s.w.g., are wound on half of the ferrite core. Suitable alternatives for the B1181/A transistors are Texas Instruments types 2G229, 2G230 or 2G231. The 12FR5 diode is rated at 12A, 50 p.i.v., and the Zener diodes Z2A150F at 15V, 1W.

conditions, falling to an average value of 750 V during transmission. The reservoir capacitor combination should have a total capacitance of at least 80 μF with a voltage rating not less than 900 V. Equalizing resistors must be included to prevent inequality between the leakage currents of the individual electrolytics causing the voltage across one to exceed its maximum rating. Convenient use is made of the transformer centre tap on the h.t. winding to provide the requisite 300 V supply for the driver valve. This 300 V supply also provides the stabilized 210 V required for the p.a. screen grids. Stabilization is obtained with the series neon stabilizer and Zener diode combination. Use of series Zener diodes alone to give the same voltage would be a better arrangement.

As transformers with bias windings are not easily obtainable, a readily available 6·3 V heater transformer is used with the secondary winding connected across a 3·15 V supply provided on the main transformer. An output of 125 V a.c. is then available at the secondary and is rectified by D23. Although there is a bias control on the transceiver chassis the bias control RV12 is fitted so that the bias voltage supply from the a.c. mains p.s.u. can be adjusted to the same value as that produced by the mobile p.s.u. This makes it unnecessary to re-adjust the transceiver bias control when changing from home station operation to mobile and vice versa.

A suitable supply for the transistorized stages is derived from a small 12 V 1A transformer feeding the regulator. The series regulator is of standard design the output voltage being adjusted to −12 V, prior to use, by the pre-set voltage control RV11. Input voltage supplied by the bridge rectifier can be any value from −17 V to −30 V without affecting the output voltage.

As can be seen from the photographs, the power supply has been constructed to form a matching unit with the transceiver, and following current commercial transceiver practice, the loudspeaker is also mounted in this unit. Layout details are not given for this unit, as it is in no way critical and will depend upon the relative sizes of individual components used.

Mobile Power Supply

The output available from the power unit Fig. 10.117 must be some 200 W maximum to allow for the momentary heavy current which occurs during tune-up operations, a current approaching 20A (allowing for some slight inefficiency) will be taken from a 12 V car battery. The transistors used must be capable of switching this 20A input current to produce the required d.c. to a.c. conversion. A pair of Bendix B1181/A transistors are ideally suited for this application, having a collector junction current rating of 25A and a 160 V breakdown figure. Suitable alternatives are suggested in the components list.

By using a ferrite core for the combined saturating and output transformer, a high efficiency (about 90 per cent) can be obtained without resorting to the complication of a two transformer circuit.

A suitable core may be obtained from the line output transformer removed from a discarded TV receiver. The old windings should first be stripped off, revealing the square ferrite core which may be carefully separated into two halves. In preparation for winding it is then necessary to tape the core with plastic insulating tape to prevent the winding coming in contact with the core. The procedure then adopted is to first wind the large number of turns for the two secondaries on to the two separate halves of the core, prior to glueing them together with " Araldite." Any gap between the two halves is kept small by tying the transformer together while the glue sets. It is now possible to put the heavy gauge primary and feedback windings on to the assembly. A tag panel and a coating of " Araldite " now completes the construction.

Simple mechanical construction of the power supply is offered by the use of a standard Eddystone diecast box, length $7\frac{3}{8}$ in., width $4\frac{11}{16}$ in., depth 3 in., to house the complete unit. A finned transistor heat sink, mounted on top of the box, is used for mounting the power transistors.

The capacitors connected across the transformer windings are for transient suppression to prevent the voltage rating of the transistors being momentarily exceeded, resulting in breakdown. Control of the power unit is by the heavy duty 12 V relay which is actuated by the VOX circuitry during transmit, giving a worth-while improvement in overall operating efficiency and freedom from interference during reception. The only disadvantage, a minor one, is a slight delay which may be noticed at the beginning of " overs " is due to the time taken for the reservoir capacitors to charge.

All the circuits given have assumed positive earth operation. As some modern cars are fitted with negative earth it may be necessary to alter the power unit for this kind of operation. However, powering of the transistor stages will present a problem somewhat more difficult to overcome. A suitable solution is to use a small continuously-running transistor d.c.-to-d.c. converter, providing a similar output to that supplied to the regulator circuitry. By moving the regulator circuitry (TR36 TR37, and TR38) on to the transceiver chassis, a completely compatible system is possible. Using an extra transistor d.c.-to-d.c. converter may seem a disadvantage but immediate switching from positive to negative earth is possible with this arrangement. The design could be based on two small receiving transistors (OC81's should be suitable) and a transformer wound on a small ferrite pot core.

A RECEIVER WITH NOISE IMMUNITY AND FREQUENCY SYNTHESIS

There have been several recent developments in the field of communications receiver design of interest to the radio amateur, particularly in connection with frequency synthesisers, and the treatment of man-made impulse noise and cross-modulation. The amateur band receiver to be described was developed by P. G. Martin (G3PDM), and is an attempt to apply some of these latest techniques to a practical design. **Fig. 10.118.** shows the general form of the circuit.

Cross-modulation and overload usually occur in the mixer stages of a receiver, and are particularly severe when a high-gain r.f. stage is used. Much can be gained by having the main selective filter close to the aerial, separated from it by a single mixer stage and *no* r.f. stage. A suitable mixer valve, with low noise performance and large signal handling capabilities is the RCA 7360 balanced beam-deflection valve. Used alone, this yields a receiver sensitivity of 0·4 μV for a 10dB signal-to-noise ratio at SSB bandwidths, together with the ability to handle interfering signals as strong as 2 volts without noticeable cross-modulation or overload. As the 7360 is a balanced device, it has further advantages over conventional mixers. The anodes are balanced with respect to the control grid, so i.f. rejection can be 60dB without any front-end selectivity; oscillator radiation is low, as the deflection electrodes are balanced with respect to the control grid; and mixing at local oscillator harmonics is reduced.

The 7360 valve is far superior to any present-day semiconductor device for mixer applications, so this receiver is a hybrid design. In general, valves are used for r.f. circuitry, and transistors for pulse and switching functions. A total of 15 valves, and five transistors and 22 diodes are used.

The receiver, **Fig. 10.119** is a single conversion design (in order to minimise the number of mixers) with an intermediate frequency of 1·62 Mc/s. Three half-lattice crystal filters provide the main selectivity for SSB reception, giving a 6 dB bandwidth of 2·5 kc/s and a 60dB bandwidth of 4·0 kc/s. Three i.f. amplifiers are needed to make up for the absence of an r.f. stage, and these use frame-grid pentodes (Mullard EF183), with their excellent a.g.c. characteristics. A second 7360 valve (V5) is used as a product detector, again in a balanced circuit for maximum linearity, and this supplies the ECL83 audio amplifier (V6). No provision is made for a.m. reception except by exalted carrier techniques, but for CW reception an active RC audio filter is built around the triode section of the ECL83 (V6A). A tunable rejection-type Q-multiplier (V9) and a 100 kc/s crystal calibrator (V10) are also provided.

To render the receiver virtually immune to man-made impulse noise, a special noise silencer has been developed. The silencer takes wideband noise (and signals) from the mixer anode circuit and amplifies them in two high-gain wideband amplifiers (V7, V8). The output is envelope detected by D5 (OA47) and as the noise pulses will invariably be greater in amplitude than the instantaneous value of all the signals in the band, the detector output is passed to the noise pulse shaper TR1 (2N3702 or similar). The square silencing pulses are now applied to the special FET noise gate TR3 (2N3819 or 2N3823), which shorts the main receiver signal path for the duration of each noise pulse. The gate/drain capacitance of the FET, about 3 pF, causes small transient pulses to appear on the noise gate output each time it is switched. These are cancelled out by means of the inverted silencing pulses—at the collector of the phase-splitter TR2 (2N3702)—which are coupled to the gate output through a 7 pF trimmer VC2.

By using a single-conversion design, frequency stability cannot be achieved in the same way as in double-conversion receivers with a tunable first i.f. and a crystal-controlled first oscillator. The single mixer requires oscillator injection over the range 3·42 to 31·62 Mc/s, and this can be achieved in any of four ways, of varying stability, complexity, expense and freedom from spurious responses.

(i) A simple band-switched LC oscillator can be used, tuned by a variable capacitor ganged to the front-end tuned

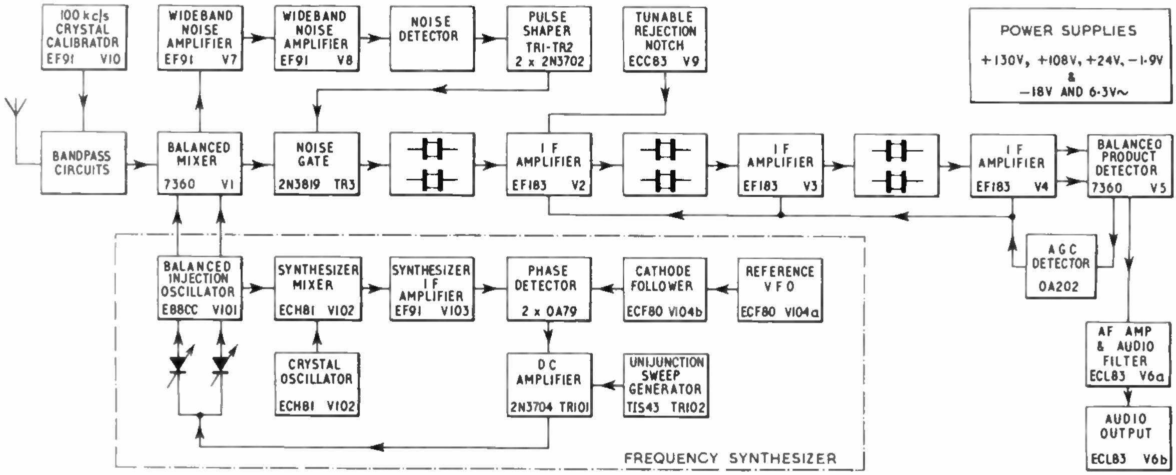


Fig. 10.118. Block diagram of the G3PDM receiver.

circuits, as in many conventional single-conversion receivers. This method is the easiest, but lacks stability, particularly on the h.f. amateur bands.

(ii) The crystal-mixer v.f.o. technique has the inherent stability of a double-conversion receiver, but the likelihood of strong spurious responses and heterodynes due to high order products from the oscillator mixer reaching the receiver mixer.

(iii) Phase-locked frequency synthesis has the stability of method (ii), and the advantage of method (i) that only one oscillator is intimately connected to the receiver mixer.

(iv) Full frequency synthesis has the highest stability of all, being based on a v.f.o. phase-locked to a multiple of an extremely stable low-frequency reference oscillator. At the present time this method is prohibitively expensive.

The receiver uses a phase-locked frequency synthesiser. The bandswitched LC oscillator V101 (E88CC) is phase-locked to the stable reference v.f.o. V104 (ECF80), using a band-switched crystal oscillator V102 (ECH81 triode section) to bring the frequency of the LC oscillator down to that of the v.f.o. The wideband synthesiser i.f. amplifier V103 (EF91) is tuned to cover the v.f.o. range of 5·88–6·38 Mc/s. The synthesiser mixer V102A (ECH81 heptode section) provides signals in this range, which was chosen for freedom from spurious responses. The v.f.o. uses the Vackar circuit, and includes a cathode follower for isolation purposes and a low-pass filter to attenuate harmonics. The v.f.o. and synthesiser i.f. outputs are applied to the phase detector D105, D106 (OA79), whose output in the steady state is a d.c. voltage proportional to the phase difference between the two input signals. The d.c. amplifier TR101 (2N3704) brings the detector output to a suitable level for applying to the varactor diode; D101–4 (BA111 in the local oscillator tank circuit. When phase-lock is lost, as when changing bands, a unijunction saw-tooth oscillator TR102 (TIS43, 2N2160 or similar) sweeps the LC oscillator across its full tuning range until phase-lock is achieved.

The main receiver h.t. line is only +130 volts, so that heating is minimised. This voltage is obtained from a bridge rectifier and a rewound transformer. A regulated +108 volts line is obtained from V11 (OB2), and a regulated +24 volts line from two 12 volts Zener diodes. A voltage doubler from the 6·3 volts a.c. heater line provides −18 volts for bias and for the noise silencer circuits.

Front panel controls on the receiver are IF GAIN, AF GAIN, BAND, PRESELECTOR, MAIN TUNING, NOISE SILENCER THRESHOLD, MODE (CAL, LSB, LSB CW, USB CW, USB, RTTY), and NOTCH FREQUENCY. The PRESELECTOR control tunes the capacitively-coupled tuned circuits in the mixed grid circuit. The coils will tune the amateur bands with a tuning capacitance of 6/20 pF, but a 100 pF variable capacitor is used so that the receiver can be used to cover most frequencies outside the amateur bands without additional coils.

The MODE switch selects the appropriate carrier insertion crystal, except in the RTTY position where the oscillator frequency is determined by a tuned circuit, and provision is made for automatic frequency control by means of the varactor diode D2 (BA111). A varactor diode (D4) also tunes the Q-multiplier. Automatic v.f.o. frequency correction is provided when switching sidebands, by a 10 pF trimmer which is switched into the circuit by forward-biasing diode D108.

The front-end coils (L1 and L2 series), are made by

Electroniques, the aerial coils being designed for a 75 ohm input impedance. A 1·62 Mc/s i.f. trap may be added in the aerial lead to increase i.f. rejection to over 100dB. The 7360 mixer (V1) operates with fixed bias of about −2 volts, adjustable for best cross-modulation performance by means of VR1 The balanced anode circuit of the 7360 consists of a special i.f. transformer with centre-tapped capacitance and inductance in its primary circuit. The inductive centre-tap is not earthed, as accurate balance is achieved by means of VC1.

The deflection electrodes of the 7360 require a *balanced* oscillator injection of 7·5 volts r.m.s. (20 volts peak-to-peak). This is obtained by means of the cross-coupled Hartley oscillator V101 (E88CC); six oscillator coils are provided (one for each of the amateur bands).

The noise gate output feeds an i.f. transformer which has a capacitively centre-tapped secondary suitable for driving the first half-lattice crystal filter (X1, X2). Each crystal filter is loaded by a resistor, rather than the more usual tuned circuit. The third i.f. amplifier (V4) has a conventional S-meter bridge circuit attached to its cathode, the reference voltage for this being obtained from the cathode of the output valve V6B. VR2 is used to set the meter zero with the a.g.c. line earthed.

The 7360 product detector (V5) is of the self-excited type, using the Colpitts configuration. Diode D1 (OA79) prevents the control grid from going positive and causing excessive distortion. In the RTTY mode, the tuned circuit L4, C1 is used: this must have a low L/C ratio. The varactor diode D2 has a fixed bias of −5 volts across it in the reverse direction, so that this a.f.c. input can vary from 0 to +25 volts with respect to chassis.

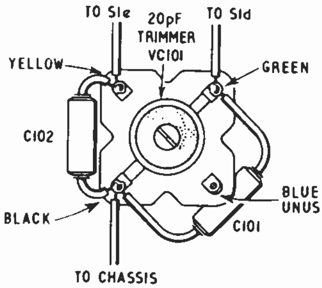


Fig. 10.120. Method used to adapt the local oscillator coils (Electroniques OS series) for use in a push-pull oscillator, using the fitted trimmer for final balance adjustment.

The active audio filter for CW reception is based on a three-section phase-shift oscillator incorporating V6A. VR3 sets the loop gain to just below the oscillating level. The bandwidth of the filter is about 200 c/s at 6dB, centred on 1 kc/s.

Audio-derived a.g.c. is used, with the detector D3 (OA202) coupled to the product detector output. The 100 kilohm resistor R1 slows down the a.g.c. rise-time slightly so that it does not respond to short noise pulses. The decay time of the a.g.c. line is about 1 second. VR4 controls the i.f. gain of the receiver.

In the synthesiser, the push-pull oscillator V101 feeds the control grid of the synthesiser mixer (V102A), and a Miller oscillator V102B is coupled to its suppression grid by means of C104. The bandswitch selects the appropriate crystal (see **Table 10.25**), the correct tuning capacitance and trimmer to resonate L102, and the appropriate local oscillator coil (L101) and capacitors (C101, C102). A 20 pF trimmer (VC101) is included with each oscillator coil to adjust for correct balance (see **Fig. 10.120**). The synthesiser i.f. amplifier V103 uses a conventional circuit with wideband couplers T101 and T102. These are Electroniques units designed to tune 5·0–5·5 Mc/s, so each 30 pF tuning capacitor is changed to 22 pF so that the range 5·88–6·38 Mc/s is covered, except in the secondary of T102, which is untuned. Ideally the phase-detector components (R101, 2, 3, 4, C105, 6, and D105, 6) should be matched to five per cent or better. The output of the d.c. amplifier TR101 varies between about 5 and 17 volts during normal operation. Diode D107 prevents the varactor control line falling below a level of 2 volts. C107 determines the sweep oscillator speed.

Construction

The receiver is built on a 16 s.w.g. aluminium chassis 14½ in. wide by 10 in. by 2½ in. deep, with a 6½ in. high front panel. Extensive screening is used under the chassis: all oscillators are completely enclosed, and the front end tuned circuits are screened from each other. The reference v.f.o. is mounted in a home-made box which derives its rigidity from a 2 in. × 3 in. dural channel ¼ in. thick and 3¾ in. long. A diecast box would also be suitable. Adjustable temperature compensation for the v.f.o. is achieved with an Oxley "Tempatrimmer," and the v.f.o. box is fixed to the main Eddystone 898 tuning mechanism rather than to the chassis, so that the v.f.o. frequency does not vary with chassis flexure.

Point-to-point wiring is used whenever possible, in order to keep leads short. High quality coaxial leads are used between the aerial socket and the front-end switch, between the local oscillator V101 and the synthesiser mixer V102A, and between the v.f.o. output socket and the phase detector. Microphone cable is used for the leads to the a.f. gain control.

New crystals are used throughout, although some saving would be possible using surplus crystals, if suitable types become available. Those crystals in the Miller oscillator which operate above 20 Mc/s were specially made for overtone operation: surplus crystals would necessitate using a more complex circuit (such as the Squier), or using the Miller oscillator in the harmonic mode.

Great caution should be exercised when substituting components for those specified. In particular, the BA111 varactor diodes should not be replaced with rectifier diodes. All the coils and i.f. transformers are available from Electroniques, and their part numbers are quoted in the Tables. The same firm can supply switches, trimmers, varactor diodes and filter crystals (see **Table 10.26**).

Alignment

Correct alignment of this receiver can only be achieved with a selection of test equipment. Particularly useful are an oscilloscope with a 7 Mc/s bandwidth, a general coverage r.f. signal generator and receiver, a wobbulator sweeping a few kHz either side of 1·62 Mc/s, and a fast pulse generator. The wobbulator is easily bread-boarded, and the other items are available in most electronics laboratories. Brief alignment details are as follows.

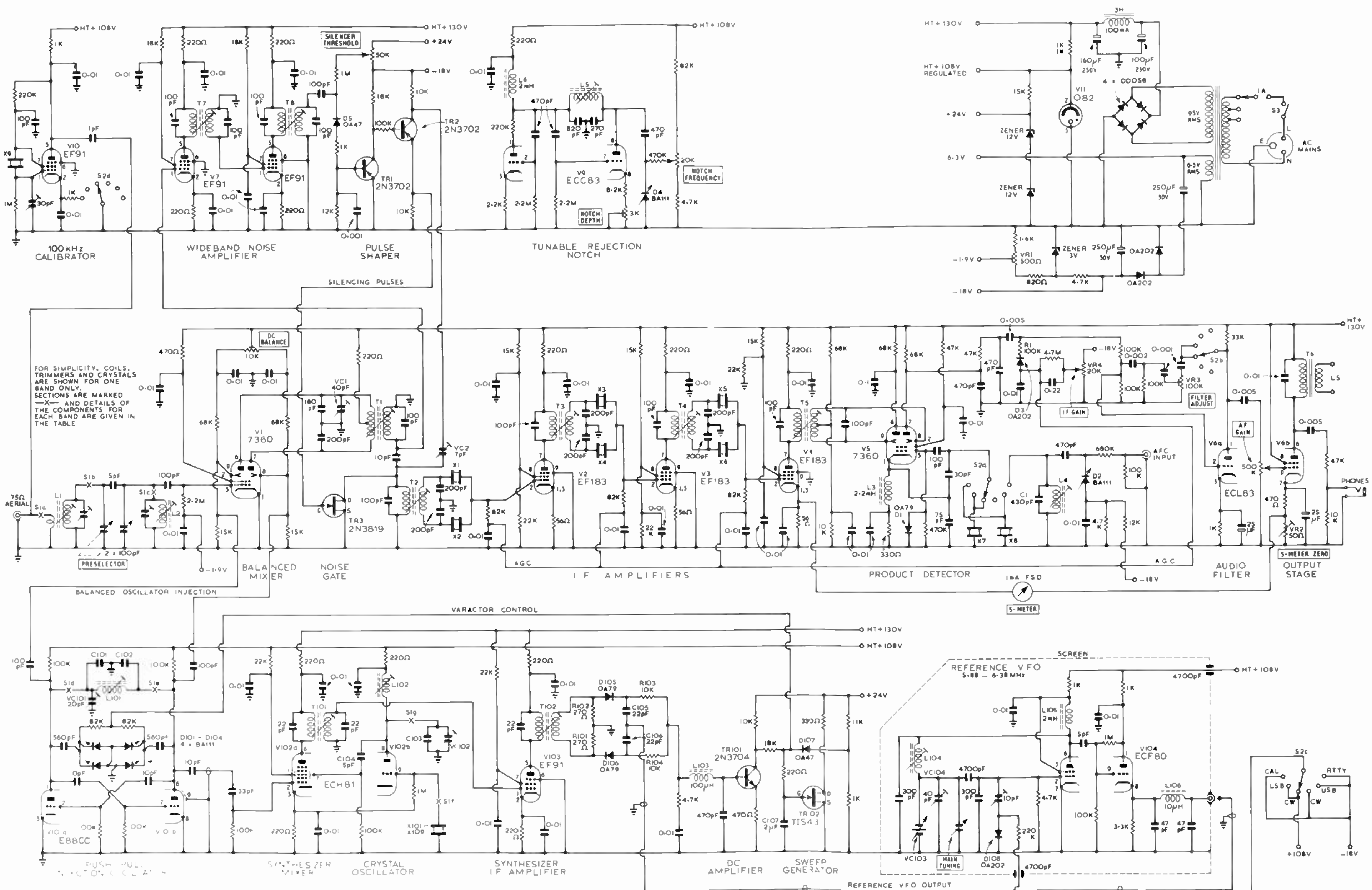


Fig. 10.119. Circuit diagram of the G3PDM receiver. To simplify the circuit, coils, trimmers and crystals are shown for one band only. Switch sections are marked X. Component details for each band are given in Table 10.25.

TABLE 10.25.

Details of components and operating frequencies for various stages of the frequency synthesizer and receiver mixer (V1). Coil references are Electroniques part numbers.

Band (m.)	Signal Frequencies (Mc/s)	Local Oscillator Range (Mc/s)	Crystal Frequency (Mc/s)	L1	L2	L101	C101 C102 (pF)	C103 (pF)	VC102 (pF)
160	1·8–2·0	3·42–3·62	9·5	LZ 1·8	BP 1·8	O5 1·8/16	47	100	40
80	3·5–3·8	5·12–5·42	11·3	LZ 3·5	BP 3·5	O5 3·5/16	0	68	40
40	7·0–7·5	8·62–9·12	15·0	LZ 7	BP 7	O5 7/16	68	22	40
20	14·0–14·5	15·62–16·12	22·0	LZ 14	BP 14	O5 14/16	68	0	20
15	21·0–21·5	22·62–23·12	29·0	LZ 21	BP 21	O5 21/16	33	0	20
10	28·0–28·5	29·62–30·12	36·0	LZ 28	BP 28	O5 28/16	0	0	20
	28·5–29·0	30·12–30·62	36·5					0	20
	29·0–29·5	30·62–31·12	37·0					0	20
	29·5–30·0	31·12–31·62	37·5					0	20

Switch on the receiver and check the various power supply rail voltages. Plug in the ECL83 a.f. valve (V6) and check for correct operation with and without the audio filter. As it is advisable to buy the c.i.o. crystals only when the crystal filter response has been accurately plotted, the i.f. amplifiers and filters should be aligned and tested with the MODE switch in the RTTY position (after L4 has been set).

Using a general coverage receiver or wideband oscilloscope peak up the crystal oscillator trimmers and check V102B for overtone operation on those bands where it is expected. Remove the reference oscillator valve V104, plug in V103, and monitor the output of T102 on an oscilloscope. Switch on the 20-metre band, and adjust the appropriate push-pull oscillator coil until the synthesiser mixer output is sweeping through the passband of the wideband couplers. The sweeping local oscillator is now wobbulating the synthesiser i.f. strip automatically: adjust the cores T101 and T102 for a square response, and use a signal generator to check that the passband includes the v.f.o. range 5·88–6·38 Mc/s. Plug in the master v.f.o. V104 and adjust L104 and VC104 for the required tuning range, using an accurately calibrated receiver or a frequency counter. The push-pull oscillator V101 should now be phase locked to the v.f.o., and as the v.f.o. frequency is varied, the local oscillator should follow.

On the remaining amateur bands it should only be necessary to adjust the appropriate local oscillator coil so that the sweep frequency range includes the injection frequencies for the band in question, in order to achieve phase-lock. Connect a high resistance voltmeter between the cathode of D107 and earth to check that the varactor control voltage is between 5 volts and 17 volts on all amateur frequencies. Once this is done, it remains to align the front-end circuits, and to adjust the Q-multiplier and crystal calibrator.

The wideband noise amplifier is best aligned by connecting a fast pulse generator to the aerial socket, feeding in 1 μs pulses with a repetition frequency of about 100 c/s, and adjusting the wideband couplers so that the output pulse resembles the input pulse. It is important that the output pulse should have the fastest possible rise-time and a flat top, even at the expense of the pulse being appreciably lengthened. The noise pulse shaper and gate require no adjustment except to set VC2 so that the gate output is free of transients when the gate is pulsed.

Simplified Versions of the Receiver

One of the design considerations for this receiver was that it should be possible to break it down into several units, some of which could be simplified or omitted completely.

TABLE 10.26

Special components in the receiver. Coil and transformer references are Electroniques part numbers.

L1, 2	see Table 10.25.
L4	1·6 Mc/s i.f. trap (RLZ 16) used as b.f.o. coil.
L5	Q-multiplier coil (QL2).
L101	see Table 10.25.
L102	BP 21.
L104	High grade v.f.o. coil (TO5/C with all capacitors removed).
VC103	Oxley "Tempatrimmer" (optional).
D2. 4 101, 2, 3, 4	BA111 variable capacity diodes.
D105, 6	Matched pair OA79.
T1	1·6 Mc/s i.f. transformer (S3D–6–1620 with modified primary capacitance).
T2, 3, 4	S3D–3–1620.
T5	S3D–6–1620.
T6	50:1 output transformer.
T7, 8	1·6 Mc/s wideband couplers (WBC 1·8 unmodified).
T101, 2	5·88–6·38 Mc/s wideband couplers (WBC 5 with 30 pF capacitors replaced by 22 pF, except secondary of T102, which is untuned).
X1, 3, 5	1618·4 kc/s filter crystals.
X2, 4, 6	1620·2 kc/s filter crystals.
X7, 8	Carrier insertion crystals (selected after plotting filter response).
X9	100 kc/s.
X101–9	see Table 10.25.

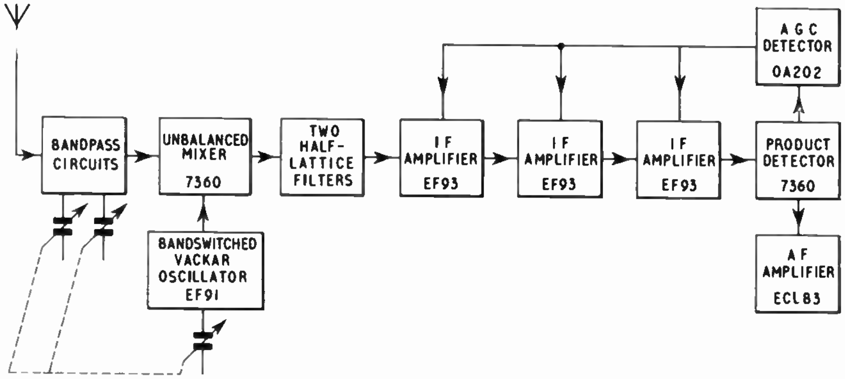


Fig. 10.121. Block diagram showing the general form of a simplified version of the G3PDM receiver.

Most of the stages contributing to the performance of the receiver are included in the six main signal-handling stages, V1 to V6. The noise silencer, consisting of V7, V8, TR1, 2, and 3, could be omitted, as could the Q-multiplier V9. The entire synthesiser could be replaced by a less stable band-switched oscillator, based on the circuitry around V101, or on an unbalanced oscillator circuit. One possible simplified design is the seven-valve receiver outlined in **Fig. 10.121.**

CHAPTER 11

RTTY

THE use of teleprinters over radio bearer circuits, known as RTTY, depends on much more than a teleprinter in good working order. Not only is frequency-shift keying almost mandatory, but a good understanding of the teleprinter synchronizing system in relation to the Murray code signals, and what the ionosphere can do to them, is essential. In this chapter a number of terms which are peculiar to telegraphy, rather than radio, will be used, and these are explained in the text.

ASPECTS OF SIGNALLING

Mark and Space

Most amateurs know that a "space" condition results in Morse code when the key is *up*, and when the key is pressed *down*, a " mark " condition results. In normal c.w. transmission of Morse, a space is simply carrier switched off, and a mark is carrier switched on. This is known as on-off keying. The terms become more obscure when applied to frequency-shift keying, for the transmitter radiates full power in both mark and space conditions. In this case the difference between mark and space is represented by a small change in carrier frequency. Frequency-shift keying is analogous to double-current keying over lines.

Single and Double-current Signalling

Machines employing the Murray code vary in the signalling voltages and currents they use. A teleprinter will normally transmit the same line conditions as it needs to receive. The British standard uses 80 volts negative at 20 mA for a mark and 80 volts positive at 20 mA for a space against earth as the reference voltage. The transmitting contact is a single-pole changeover switch, one fixed contact being connected to negative 80 volts and the other to positive 80 volts. The transmit line is connected to the moving contact, known as the " tongue." Some protective arrangement is necessary to prevent the contacts from burning in the event that they " bunch " through maladjustment. The best arrangement is to insert a lamp having a deliberately poor resistance/temperature characteristic in series with each 80 volt feed. The lamps do not glow under normal 20 mA loading, but if the contacts bunch the excessive current passed causes the lamps to glow and their resistance increases greatly, thus limiting the current to a safe value, and incidentally, warning the operator.

As is to be expected, the European Continental standard is different, but this only affects teleprinters connected by a direct line. Fortunately the signalling speed is the same, 50 bauds, so that the systems are compatible on RTTY, although the mark and space polarities may be reversed. The use of current to represent both mark and space is known as double-current signalling in the United Kingdom and as polar signalling on the American continent.

American Teletype machines use single-current signalling, called neutral signalling in North America. This is simply on-off keying. At mark, the Teletype machine sends 110 volts negative at 60 mA to line, and at space it sends nothing. The signalling speed is based on 60 words per minute, which results in an odd speed in bauds. Generally stated as 45 bauds, although fractionally greater, the speed of Teletype machines is not compatible with British and Continental machines, and they cannot work directly to each other. Amateurs are mainly concerned with surplus machines, so that British and Continental amateurs think in terms of 50 bauds machines, whilst American amateurs use 45 bauds machines.

The Baud

The baud is the shortest single signal unit in a signalling code. It is rather more precise than the term " bit " used by computer engineers, since it assumes a square shaped signal, which may be either a mark or space, at a definite repetition speed. It is thus used to specify signalling speed, and the Creed 7B teleprinter works at 50 bauds, which means that 50 signal units are transmitted every second. If telegraph signals were sent as " reversals," i.e. mark-space-mark-space alternately, one could define speed in terms of frequency, one mark followed by one space representing one complete cycle of a square waveform. The Murray code, however, has no gaps between the signal units, and the mark-space

The Creed Model 7 Page Teleprinter.
(Courtesy of Creed & Co. Ltd.)

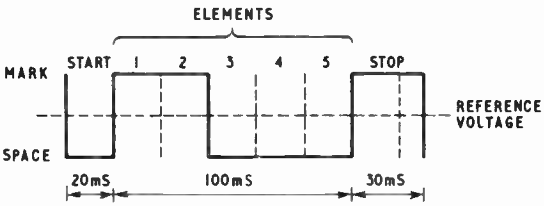


Fig. 11.1. Letter A, typical character construction.

sequence depends on the character transmitted. Letter A, for example, consists of *two* marks followed by *three* spaces in sequence, so that the waveform is not symmetrical, and cannot be described in terms of frequency without giving a totally incorrect impression of signalling speed, see **Fig. 11.1.** Nevertheless, it is true to say that the highest signalling speed that may arise in any given code is the reversals frequency, which is 25 c/s in the case of 50 bauds, and design of circuit constants must take this frequency into account.

Speed stated in bauds is precise, and the length of the baud as a fraction of a second is implicit. Speed stated in words per minute is approximate only, since an average length word must be assumed.

In Morse code the number of bauds in a character varies greatly. The Morse baud is one dot. Letter E is thus equivalent to four bauds, the dot and the space of three dots' duration following before the next character commences. Figure 0 is 22 bauds, consisting of five dashes each worth three bauds, four spaces each worth one baud between the dashes, and the space of three bauds' duration following the character.

Murray code, used with teleprinters, always has five character bauds plus one start baud and one and a half stop bauds, the latter being used for machine synchronizing. All characters thus take the same time to transmit—an obvious requirement for any form of type printing telegraph machine.

TABLE 11.1

The Murray Code.
Upper case characters may vary.

Upper Case	Lower Case	Elements 1	2	3	4	5
-	A	M	M	S	S	S
?	B	M	S	S	M	M
:	C	S	M	M	M	S
AB†	D	M	S	S	M	S
3	E	M	S	S	S	S
%	F	M	S	M	M	S
@	G	S	M	S	M	M
£	H	S	S	M	S	M
8	I	S	M	M	S	S
Bell	J	M	M	S	M	S
(	K	M	M	M	M	S
)	L	S	M	S	S	M
.	M	S	S	M	M	M
,	N	S	S	M	M	S
9	O	S	S	S	M	M
0	P	S	M	M	S	M
1	Q	M	M	M	S	M
4	R	S	M	S	M	S
!	S	M	S	M	S	S
5	T	S	S	S	S	M
7	U	M	M	M	S	S
=	V	S	M	M	M	M
2	W	M	M	S	S	M
/	X	M	S	M	M	M
6	Y	M	S	M	S	M
+	Z	M	S	S	S	M
Carriage Return		S	S	S	M	S
Line Feed		S	M	S	S	S
Word Space		S	S	M	S	S
Letter Shift		M	M	M	M	M
Figure Shift		M	M	S	M	M
All Blank		S	S	S	S	S

† AB denotes the key which triggers an answer-back mechanism which can be fitted to the distant machine.

Murray Code

This is the five-unit code internationally adopted for teleprinting machinery. The complete code is shown in **Table 11.1**, and it will be noted that there are 32 possible combinations, of which 31 are used. The exception is the character comprising all spaces. A teleprinter receiving a continuous space condition is receiving a permanent start signal, and will have its mechanism permanently rotating without printing any character. This condition is known as " spacing-in," and is commonly the first intimation that the telegraph line has become faulty.

The marks and spaces which convey intelligence are known as " elements," and are numbered 1 to 5 in sequence. When transmitted, the elements are also bauds, but the term element implicitly describes one of the bauds making up the character, and not the start and stop signals, which are for synchronism and do not contain message intelligence.

The start signal is transmitted at the commencement of each character, and consists of a single space baud. This is followed by five bauds, spaces or marks, depending on the character formation, and finally a stop signal of one and a half bauds mark minimum duration is transmitted. When sending by keyboard the actual length of the stop signal will depend on how long the operator takes between pressing one key and the next. There is a locking bar on the keyboard which prevents the operator from pressing another key before the previous character has been fully transmitted, and this ensures that the length of the stop signal is at least one and a half bauds.

At a speed of 50 bauds the length of one baud is 20 milliseconds, so that each character takes 150 milliseconds to transmit. This is approximately equal to 66 words per minute.

The Receive Magnet

The receiving line of a teleprinter circuit is connected through a current limiting resistor to an electro-magnet, whose armature movement initiates the mechanical functions of the teleprinter. Obviously, since all elements are transmitted sequentially with no pauses between them, the magnet must operate and release very quickly and follow the marks and spaces faithfully. There must be a finite time during which the armature is in motion. The machine is interested in only two conditions—mark and space which is armature fully operated or fully released. The time taken for the armature to travel from one condition to the other is known as the " transit time." If this time is an appreciable portion of the baud length, it will reduce the time available for the selecting mechanism of the teleprinter to determine the polarity of each baud. This becomes very serious when working over a radio path, because considerable distortion of baud lengths can be fortuitously introduced by ionospheric phenomena, such as multiple path propagation. Moreover, the time taken for the armature to move from mark to space must be the same as that taken to move from space to mark, or a mechanical bias of the mechanism will result.

A machine using single-current signalling has an electromagnet whose armature is returned to space by a spring. When the magnet is energized, the magnetic force acting on the armature has a square law characteristic. The nearer the armature gets to the pole face, the greater the magnetic force acting upon it. Thus, to travel from space to mark, the armature can be in continuous acceleration . When the current in the magnet coil is removed, the spring, which has a linear characteristic, is left to return the armature to space position. Its acceleration is therefore quite different. However, during armature operation by energizing the coil, the spring is acting against the magnetic force, and a fairly delicate balance of spring tension against magnet coil ampere/turns can be found which makes operate and release transit times equal, at the expense of reduction of magnet sensitivity. It should be noted that the single-current Teletype machine requires 60 mA magnet current against the 20 mA used for Creed double-current teleprinters.

The double-current operated machine employs a polarized electro-magnet which is exactly balanced mechanically. As in the case of the simple headphone, polarization much increases the sensitivity of the magnet. It operates under exactly the same force conditions in both directions, since the armature is driven by magnetic force from space to mark and from mark to space. Consequently it does not introduce any inherent mechanical bias, and the transit time is kept very short by the high sensitivity of the magnet.

Start-Stop Synchronizing

When a teleprinter is switched on, but the keyboard is at rest, it transmits a steady mark. If it is connected to a distant teleprinter receiving magnet, that magnet remains operated to mark and nothing happens. Both motors are turning, ideally at exactly the same speed, but the teleprinter mechanisms are idle because of a clutch device which isolates them from their motors. When an operator presses a character key he engages the sending mechanism, and a start signal of one space baud is transmitted, followed by the five elements of the character and a stop signal. The start signal switches the distant teleprinter magnet to space and the clutch engages, causing the receiving mechanism to rotate. A mechanical selecting device then samples the position of the receive magnet armature at fixed intervals of time following the engagement of the clutch (**Fig. 11.2**). After one rotation the clutch again latches during the period of the stop signal. Since the position of the armature depends on the character

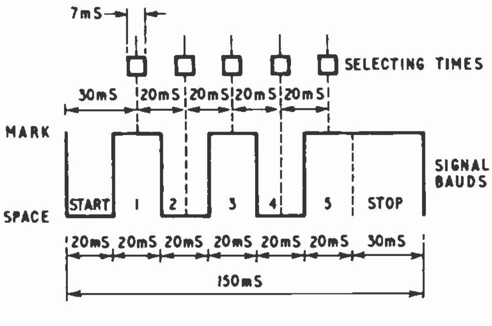


Fig. 11.2. Letter Y showing ideal selection timing.

transmitted, the selecting device is able to put the make-up of the character into a mechanical store until the transmission of the character is completed. It is then printed during the reception of the next character.

The intervals at which the selecting device looks at the armature position are determined by the speed of rotation of the mechanism. The first " look " is timed to take place about the middle of the first character element. Thereafter the " looks " are spaced at 20 millisecond intervals if the motor speed is absolutely accurate. Obviously, if either the transmit or receive teleprinter motor speed is inaccurate, the " looks " will be displaced from the middle of the elements, and this is cumulative. If the receive teleprinter motor were running 10 per cent slow, the middle of the selecting " look " normally timed to take place 30 milliseconds from the commencement of the start signal, would be delayed by three milliseconds. At the 110 millisecond point from the commencement of the start signal, properly coincident with the middle of the fifth element, the middle of the selecting " look " would be displaced by 11 milliseconds. Since the element length is only 20 milliseconds and the " look " takes a finite time (about 7 milliseconds in old type teleprinters), the need for accurate synchronism is important. In practice one teleprinter motor may be running slightly fast, and the other running slightly slow. The speed differences add, so that the maximum speed error that could possibly be allowed, assuming no distortion of baud length arising from any other source, would be equal to *half* the displacement of the fifth element selection to the start or finish of the fifth element baud. The baud length is 20 milliseconds, and the selection time is 7 milliseconds disposed equally about the baud at its 10 millisecond point. This leaves 6·5 milliseconds of the baud either side of the selection period. Only half this displacement can be allowed, and this works out to about 2·7 per cent deviation of motor speed. However, although two teleprinters, one having its motor running 2·7 per cent fast and the other 2·7 per cent slow, connected by a piece of wire introducing no signal distortion, would theoretically print correctly, in practice there are so many sources of possible baud length distortion or mechanical bias that a much more realistic figure for motor speed accuracy is plus or minus one per cent. The motor governor can achieve this accuracy.

A speed fault can be diagnosed fairly easily by analysis of the page-printed copy at the receiving end and reference to the Murray code. It will be noted that, as the teleprinters start in synchronism by reason of the start signal, and then run slowly out of synchronism because of the difference in motor speeds, it is always the fifth element that is most likely to be incorrectly recorded because of displaced selection. If therefore, the text of the message contains a number of errors all of which show incorrect sensing of the fifth elements, but which show correct sensing of the earlier elements, the trouble is almost certainly speed, but the fault could be on either the transmit or receive teleprinter.

On-Off Keying versus Frequency-Shift Keying

On-off keying of the radio carrier presents many disadvantages when applied to machine telegraphy. An operator copying c.w. on-off keying has at his disposal the best machine of all—his brain! He can pick out his wanted signal from often louder interference; he can read down into the noise; he can recognize a Morse character even though the sending is poor and erratic.

A machine is much more limited. It must be set to operate from a signal threshold and discard all below that threshold as noise or interference. Automatic gain control can only give very limited help, and an on-off signal that is quite strong may be wrecked by a much weaker interfering signal if the threshold is set too low. Since radio signals commonly fade up to 40db, a voltage ratio of 100 : 1, the setting of the threshold control is, at best, a dubious compromise. One must chance losing the signal when it fades in order to block off the weak interference and noise when the signal is strong. One mutilated baud in the Murray code makes the teleprinter print the wrong character and seven and a half bauds are lost. If the baud lost happens to be a start pulse, the following character may be displaced and lost.

It may take several characters before the teleprinters again come into synchronism. Suppose for example, letter A is transmitted. The teleprinter will send a start signal, two marks and three spaces in sequence, and then one and a half bauds stop signal. If, due to a sudden burst of interference, the start signal were not received, the receiving teleprinter would remain at rest until the first space, element 3, was received. This it would interpret as the start signal, and its mechanism would still be revolving on that character when, in fact, the next character was coming over the air. This is why the stop signal has an odd half baud in its length. If the complete character with its synchronizing signals were an exact number of bauds in duration, one dropped start signal could throw the teleprinters out of synchronism until such time as there was a pause in transmission. Even as it stands, one dropped start signal can throw the teleprinters out of synchronism for several characters, depending on their make-up, and this is the great weakness of the start-stop system.

Until the advent of frequency-shift keying it was generally recognized that start-stop Murray code signalling was not practicable over radio link circuits. Frequency-shift keying did nothing to remove the inherent weaknesses of the system, but it did provide a much more reliable radio link system, so that it became possible to use start-stop signalling on point-to-point commercial circuits and keep them traffic-worthy. Commercial operators regard the use of start-stop signalling as merely a stop-gap whilst more sophisticated systems are coming into use.

One big advantage of frequency-shift keying is that it always has a carrier present, mark or space, and provided the wanted signal is slightly stronger than an interfering signal, it will be correctly recorded. There is no touchy threshold control: the signal limiting stage preceding the discriminator fixes that, and noise cancels out in the discriminator circuit.

Two difficulties arise. First, use of automatic gain control is very limited and its time constant must be very short. This is because the mark and space frequencies, separated by less than one kilocycle, fade independently, and a strong mark which backs the receiver sensitivity right off on a.g.c. may be immediately followed by a weak space which would be lost if the receiver remained backed off by a.g.c. with a long time constant. In any case, unless the a.g.c. can actually follow the signals, it can give rise to some unpleasant telegraph distortion effects. The effective range of the a.g.c. is therefore tied to the dynamic range or margin

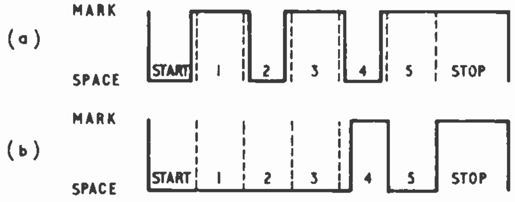


Fig. 11.3. (a) Bias distortion, letter Y with marking bias. (b) Characteristic distortion, carriage return signal with characteristic distortion of element 4.

of signal over which the limiter will operate effectively.

Secondly, radio signals are usually received on one or two paths simultaneously, and since the path lengths vary, the signals over different paths arrive at the receiving aerial at slightly different times. As an example, if 20 millisecond bauds are being received by one hop and two hop propagation, and the second hop is arriving three milliseconds after the first, each first hop baud will have three milliseconds of the previous second hop baud interfering with its first three milliseconds. On the other hand, each second hop baud will have three milliseconds of the next first hop baud interfering with its last three milliseconds. If the fact that alternate bauds may be mark and space bauds is added, and the overlaps tend to cancel out in the discriminator, it will be appreciated that six milliseconds of the 20 millisecond bauds are of very dubious value. Taken together with an otherwise tolerable speed difference between the teleprinter motors and some telegraph distortion, the chances of character mutilation become considerably greater. This effect occurs commonly enough when the radio signal appears to be very strong and would be expected to do nothing but print perfectly. The sophisticated cure is to use an aerial with a vertical reception angle that receives only one path. The unsophisticated cure is to reduce transmitter power until the weaker path fades below trouble level!

It is multiple path effects that set a term to the maximum keying speed that can be achieved in radio telegraphy. It must be borne in mind that the effect of distortion on the telegraph receiving equipment depends on the ratio of the time taken to complete the sensing process of the baud polarity and the actual length of the baud. In this respect there is scope for amateur ingenuity.

Telegraph Distortion

The term " telegraph distortion " means distortion of the length of the baud. Each character in the Murray code at 50 bauds takes 150 milliseconds to transmit, and so it follows that if any baud is artificially lengthened it can only be so at the expense of another baud in the character being shortened by the same amount.

There are three classic forms of telegraph distortion, of which bias distortion, depicted in Fig. 11.3 (a), is the most important. This gives the effect that all marks are lengthened by a fixed amount, and all spaces are shortened by a similar amount, or vice versa. If the marks are lengthened it is called a marking bias, and lengthened spaces are called a spacing bias. It is unforgivable because it is an obvious equipment fault; either it arises through mechanical adjustment of a transmitting contact or relay which is mechanically biased to one side, or because of an electronic circuit which is electrically biased to one side. The test is to put square wave signals at the correct frequency of 25 c/s through the keying circuits and check the output with an oscilloscope, adjusting the time base so that mark and space bauds interlace and checking that the transitions coincide. Relays may be individually checked by switching a current through the contacts into a centre-zero milliameter, energizing the coils with a symmetrical waveform of correct frequency, and observing that the meter pointer vibrates about zero. A bias will cause the pointer to vibrate about a position off-set from zero, and the degree of displacement from zero is a measure of the degree of bias distortion. **Fig. 11.4** shows a circuit for such a tester.

Teleprinter transmit contacts are more difficult to check, as they cannot be keyed with a repetitive frequency. One can, however, get a good idea by clipping an oscilloscope across the line output, pressing down the key for letter Y, and pushing up the keyboard pawl abutment, so that the teleprinter sends a continuous letter Y see Fig. 11.2. By running the oscilloscope at 150 milliseconds sweep time, the whole character is displayed, and one can measure mark and space bauds on the graticule to check for any distortion. It is best to make a direct connection to the *Y*-plates of the oscilloscope for these tests, as the in-built amplifier may introduce distortion and give a misleading picture.

The second form of distortion is characteristic distortion, and this is much more subtle. A circuit suffering from this finds the bias distortion tests the easiest to pass. As its name implies, the distortion varies with the character which is transmitted. It arises from the asymmetry of the waveforms involved. A carriage return character, for example, is sent as four spaces in sequence (start signal and elements 1, 2, 3) followed by one mark (element 4) and another space (element 5). If any part of the circuit has a time constant which enables the reference voltage on the grid of a keying valve to float up and down, it would charge heavily towards the space voltage during the four consecutive spaces, so that the single mark which follows would have to overcome that charge before it could register its appearance. The leading edge of the baud would take on an exponential slope according to the discharge time constant of the circuit, and after

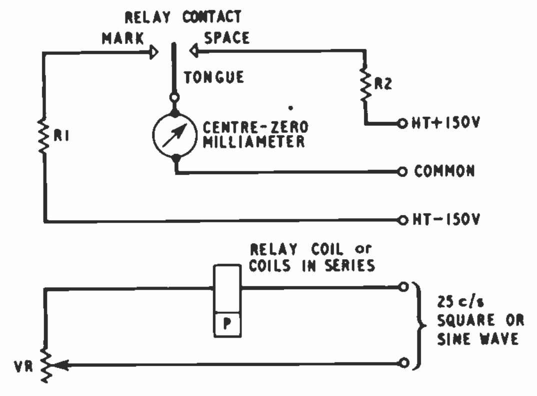


Fig. 11.4. Bias test set for polarized telegraph relays. R1 and R2 are selected to give full scale deflection in mark or space direction when the contact is operated by hand. VR is used to set the coil current.

passing through following squaring stages would appear as a short baud **Fig. 11.3(b).** On the other hand, a letter Y, which is mark-space-mark-space-mark, would knock the charge alternately as much in one direction as the other, and would have a much better chance of getting through the circuit undistorted. For this reason a.c. couplings in electronic circuitry associated with keying systems are best avoided. Amplifiers should be d.c. coupled, both at the transmitting end and following the receiver discriminator.

An interesting point arises from this in connection with the discriminator. If, as is highly desirable, it is to be d.c. coupled to its following amplifier, the output marks and spaces will need to be in the form of equal positive and negative voltage excursions against a reference voltage suited to the following stage circuit parameters. If the discriminator is tuned exactly midway between the two frequencies representing mark and space, the corresponding output voltages will be equal and opposite in polarity, referred to the reference voltage to which the output circuit is returned. If the discriminator is slightly off tune, but within the i.f. bandwidth, the mark and space voltages will still add up to the same total excursion, but will be displaced with respect to the reference voltage. This would inhibit the correct operation of the following stage, particularly if the mark to space voltage amplitude is only just enough to key it fully. In other words, tuning of the receiver must be very accurate. Commercial operators get over this by using crystal controlled receivers, or a highly accurate automatic frequency control system. This latter must be a motor tuning type, as the reactance modulated type of automatic frequency control derives its control voltage from the tuning error, and hence can never completely correct tuning without losing the control voltage which initiates the function.

Crystal control of the h.f. receiver is unsuitable for the amateur, who is not working a fixed point-to-point circuit. The automatic frequency control system is probably not an economic proposition, and would require resetting every time the amateur switched over from transmitting to receiving during a contact. The best compromise approach is to make the discriminator output voltage excursion very much in excess of that required by the following stage.

If the following stage requires a 6 volt excursion to key it fully, a discriminator output excursion of plus and minus 50 volts for mark and space would so exceed the requirement that the valve would either be biased far beyond cut-off or running far beyond grid saturation. In this case, keying accurately about the reference voltage is of less importance, because the valve can only "see" a small part of the excursion. Since telegraph circuitry has only two conditions of interest, mark or space, it is common practice to arrange all valves to work in cut-off or saturated condition as far as possible, excessive grid current being prevented by high value grid series resistors.

A discriminator working at about 455 kc/s, following a standard i.f. strip, is unlikely to produce an output excursion of more than eight volts for a frequency shift of 850 c/s. It is therefore better to change the i.f. again to something around 10 to 15 kc/s, and to operate the discriminator at this frequency. The shift frequency then becomes a much greater proportion of the base frequency and the output voltage is correspondingly much higher.

A drift of only 50 c/s using plus or minus 425 c/s deviation

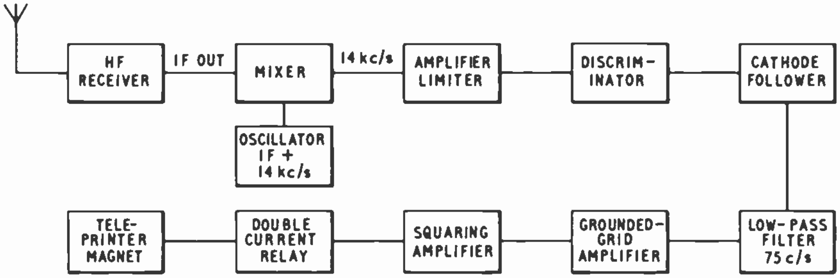


Fig. 11.5. Block diagram of a typical RTTY receiver.

from the nominal frequency represents nearly 12 per cent off tune to the discriminator. At the same time, if the signal frequency is 3·5 Mc/s, the receiver h.f. oscillator only needs to drift 0·0014 per cent to cause the 12 per cent displacement of the signal from the discriminator mid-point. The same applies to the transmitter. When receiving frequency shift telegraphy the initial tuning-in of the receiver therefore needs to be extremely accurate. The frequency stability of both transmitter and receiver is most important.

It may well be felt that the difficulties arising out of the use of d.c. coupling from the discriminator to the following stage outweigh the disadvantage of using a.c. coupling, and thereby incurring characteristic distortion effects. It is necessary therefore to point out that the time constant of the a.c. couplings must be long enough to allow a mark to pass steadily all the time the teleprinter is at rest, whilst the transmitting operator is thinking of something else to say. Unless some trick circuit is employed, the capacitors in the couplings are bound to assume substantial charges during such periods.

The third form of distortion is known as fortuitous distortion, and this is something little can be done to prevent. It is, in fact, any form of distortion that arises in the radio propagation path, such as multiple path reception already described, interference and static. The only real way to guard against multiple path reception is to design the aerial for good vertical directivity, and at 3·5 Mc/s this is obviously out of the question for most amateurs. Similarly, the best way to guard against interference is to design the aerial for good horizontal directivity, and once again this is not practicable for 3·5 Mc/s operation by amateurs. It is, however, possible to do something to alleviate the effects of fortuitous distortion in the receiving equipment the general circuit arrangement being similar to that shown in **Fig. 11.5.**

Remedial Circuitry

As has been mentioned, it is the beginning and end of the baud which is affected by multiple path reception. Looking at the i.f. signals on an oscilloscope the baud transitions can be seen preceded and followed by a portion of the bauds fading up and down quite irregularly and badly broken. The teleprinter itself accepts that the start and finish of the signal element may well be unreliable, and the selecting process is timed to take place disposed about the middle of the bauds, the duration of the process being considerably shorter than the baud length. Obviously, the shorter the selection time can be made, the less likely are telegraph distortions of any kind, and speed inaccuracies, to interfere with the selection process. Telegraph manufacturers are fully aware of this, and modern designs have speeded up the selection process. Nevertheless, it is essentially a mechanical operation, and cannot hope to compete with the selection times possible in electronic gating circuits. The amateur will almost certainly be using old type machines obtained on the surplus market, and is thus saddled with the longer selection time variety.

Noise and other interference occurs at any time during the bauds, and usually takes the form of knocking a hole of quite short duration in them. The maximum signalling frequency is equivalent to a 25 c/s square wave, and the noise pulses of short duration are equivalent to half cycles of considerably higher frequency. If the signals are passed through a low pass filter with suitably adjusted cut-off frequency, the short duration noise pulses will be unable to pass through the filter, but the wanted signals will. In order to retain the approximately square shape of the bauds, a proportion of their harmonic content must be passed, so it is reasonable to fix the filter cut-off frequency at about three times the maximum keying frequency, i.e. 75 c/s. This will deal quite happily with short breaks in the bauds occasioned by noise, and will smooth the leading and trailing transitions of the bauds, broken by multiple path reception effects, into curves of satisfactory steepness. It is important that the filter be correctly terminated in a resistive load equal to the characteristic impedance, as otherwise reactive components will be seen by the load, and the baud may become distorted.

Following the filter a squaring amplifier, with its grid running well into saturation and cut-off, is necessary to restore the square shape of the bauds. It is essential that the grid bias point of the valve is fixed in the middle of the signal voltage excursion, so that marks and spaces key it equally in both directions. Operation at a bias point displaced from the middle of the excursion will result in telegraph distortion, since the leading and trailing transitions of the bauds will have been curved by the filter action.

A regenerative repeater may be used to shorten the selection time. This is an electronic device into which the signals are passed instead of directly into the teleprinter magnet coil. The received start signals trigger off a pulse generator which sends selecting pulses of about 100 microseconds' duration, each at the correct 20 milliseconds intervals, to gate the element bauds into a flip flop. These pulses are used to sample the baud polarities and trigger the flip flop accordingly. The flip flop output then provides distortionless bauds whose length is determined by the pulse intervals, and the output baud polarities follow the signal baud polarities. The regenerated signals are fed through an amplifier to the teleprinter magnet, the selecting time having been cut from seven milliseconds to 100 microseconds. The teleprinter has to re-select the elements in the normal way, but this is now no disadvantage as the bauds are known to be distortionless. This sounds like a wonderful idea, and does provide quite an improvement in operation, but the weakness lies in the assumption that the start signal which initiates the operation of the selection pulse generator is always good. If the start signal is displaced, all the selecting pulses will be too. The regenerative repeater is thus far from being the complete answer.

The complete answer, in fact, lies with doing away with start and stop signals for synchronizing, and relying on highly stable and accurate frequency sources to control both transmitting and receiving speeds. However good the synchronizing frequency source at each end of a radio telegraphy link, the systems must eventually drift out of synchronism, so that an averaging frequency correction device is required also, and this is very complex.

TRANSMITTING FREQUENCY-SHIFT SIGNALS

Until the advent of s.s.b., frequency-shift keying was invariably accomplished by reactance modulating an oscillator stage in an exciter. The system differed from frequency modulated telephony systems only in detail, because of the specialized application. In the telephony case a.c. coupling into the modulator is normal, and the output frequency of the oscillator returns to nominal during periods of no modulation. The frequency excursion of the carrier wave is referred to as the "deviation," and this term means the amount by which the carrier departs from nominal frequency during modulation of maximum amplitude. If, for example, deviation is stated as 75 kc/s it means that the carrier will deviate up to *plus* and *minus* 75 kc/s from the nominal frequency. Although the term "deviation" is also used in f.s.k. terminology, it is much more common to refer to the *shift*, which is the full excursion of frequency from mark to space, and thus twice the deviation. The reason for this is simply that the transmitter never radiates the nominal frequency because it is always on mark or space frequency. In order to hold the transmitter to mark or space when the transmission is not being keyed, it is necessary for the control voltage into the reactance modulator to be d.c. coupled, or the oscillator would drift back to the nominal frequency as the coupling capacitor charged up. The degree of shift can be controlled by a potentiometer at the input in the same way as a potentiometer would act as a deviation control in an a.c. coupled telephony circuit.

If the oscillator frequency were changed, the amount of shift would also change, and it would be necessary to re-adjust the shift control every time a change in frequency was made. To prevent this, the modulated oscillator is used on a fixed frequency, usually 200 kc/s, and its output is mixed by the heterodyne method with a second oscillator to provide the output frequency. Frequency changes are accomplished by changing the frequency of the second oscillator, and so the shift remains constant for all frequencies. When frequency multiplication is carried out in later stages of the transmitter the shift will be multiplied by the same factor as the carrier frequency. The simple method of compensating for this is to have a switched potentiometer made up of fixed resistors in the control voltage circuit to the reactance modulator. Since the modulator slope will be linear, it is only necessary to set the potentiometer to divide the input voltage by the frequency multiplication factor to keep the shift constant for the radiated frequency. Reactance modulated type frequency shift exciters will generally have a switch which will divide the input voltage by two, three, or four, to compensate for doubling, trebling, or quadrupling of the exciter output frequency in the transmitter. Alternatively the shift control potentiometer, which is continuously variable, can be calibrated.

Since the output frequency of the exciter is the sum or difference of the modulated oscillator and second oscillator frequencies, it should be noted that sense of the keying will be reversed if one changes from the sum frequency to the difference frequency, so that marks will become spaces and vice versa. It is usual to always select the sum, or upper frequency for driving the transmitter.

The amount of shift used is not very critical provided the distant receiver is using a discriminator type detector. Reducing shift reduces the output voltage of the discriminator, but allows greater receiver selectivity to be used. If the discriminator slope is approximately linear this could be an advantage, but shift must not be so small that the operation of the circuits following the discriminator becomes marginal. The smaller the shift, the greater the frequency stability required in both transmitter and receiver. If the receiver uses an amplitude type discriminator on which the Q has been made high to give high output, the slope for small values of shift may be very flat due to non-linearity of the response, and performance will fall off rapidly with shift reduction. If the receiver uses two a.f. filters instead of a discriminator, the shift must be set exactly right for the filters.

At one time commercial operators used 850 c/s shift almost universally. Due to pressure on the h.f. frequency spectrum there is a marked tendency to reduce shift and use narrow bandwidths. With such shift and a keying speed of 50 bauds, a receiver bandwidth of around 1·2 kc/s is necessary.

Fig. 11.6 shows a typical arrangement for a frequency-shift exciter. The balanced mixer should be arranged so that the crystal oscillator frequency cancels at its output, so reducing the degree of selectivity required in the tuned amplifier which follows it. Frequency changes of a few

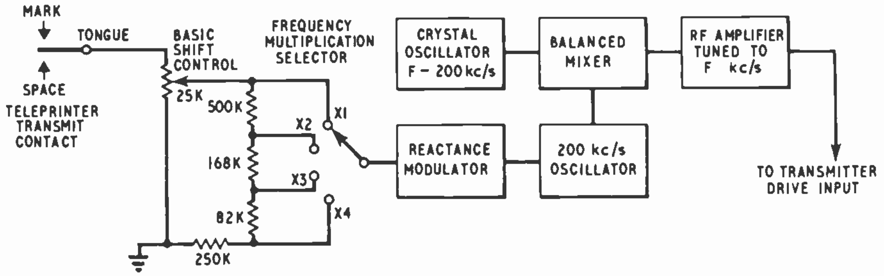


Fig. 11.6. Typical f.s.k. exciter. The reactance modulator should be arranged to cause the 200 kc/s oscillator to *increase* frequency when the teleprinter contact switches to mark.

kilocycles may be made by varying the frequency of the 200 kc/s oscillator without affecting the shift to any significant degree. Having regard to the high degree of frequency stability required in f.s.k. working, this type of exciter is particularly good. A frequency stability of 0·05 per cent in the 200 kc/s oscillator would only worsen the stability of the exciter output at 3·5 Mc/s by 0·003 per cent.

The S.s.b. Exciter as a Frequency Shift Exciter

Amateurs who possess single sideband transmitters can simulate f.s.k. very simply. The radiated frequency of an A3J emission depends on the modulation frequency referred to the suppressed carrier frequency. If two tones, one for mark and the other for space, are keyed sequentially into the audio input of an s.s.b. exciter, the transmitter output will be indistinguishable from an f.s.k. emission. Suppose the suppressed carrier frequency is 3500 kc/s, upper sideband is used, and the mark and space tones are 1350 and 500 c/s respectively. When the mark tone is keyed the radiated frequency will be 3500 + 1·350 = 3501·35 kc/s. When the space tone is keyed the radiated frequency will be 3500 + 0·5 = 3500·5 kc/s. The difference between the two radiated frequencies is 0·850 kc/s, and the s.s.b. emission is thus equivalent to an f.s.k. emission on a nominal centre frequency of 3500·925 kc/s with 850 c/s shift. Furthermore, the shift will be absolutely constant, irrespective of the nominal radiated frequency. If the exciter were to be switched to lower sideband working the effect would be to reverse mark and space frequencies.

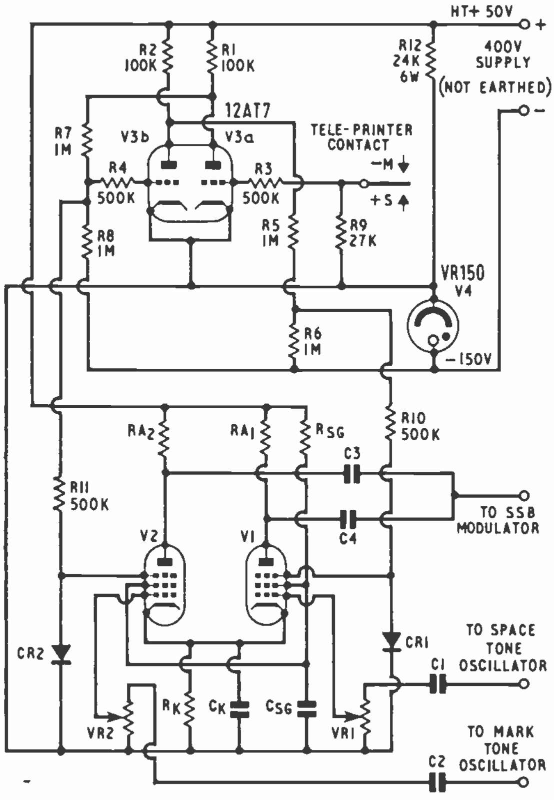


Fig. 11.7. F.s.k. signalling using an s.s.b. exciter.

It is important that the amplitude of both tones should be identical, so that the same power is radiated for both mark and space. To provide a ready means of adjusting the levels of the two tones it is best to use separate oscillators for each tone, and to feed them via gain controls into amplifiers whose output is switched by the teleprinter signals into the common s.s.b. modulator.

Fig. 11.7 shows a typical circuit for the switching arrangement. V1 and V2 act as switched amplifiers for the continuously running mark and space oscillators, the output levels being adjusted to be equal by VR1 and VR2. Switching is carried out by biasing the suppressor grids of the pentodes, and valves having a suitable characteristic for this must be used. The Mazda 6F33 is specifically designed for such service and has a sharp cut-off suppressor grid characteristic, but there is so much margin in this circuit that many small pentode types (such as the EF91, 6AM6) can be used.

Current flows to only one anode at a time, but both valves will pass screen current all the time, and the screen current of the valve whose anode is cut off will probably be higher than that of the one which is working as a normal amplifier. R_k and C_k, R_{sg} and C_{sg}, are chosen to give the normal voltages to the valve working as an amplifier in class A. The circuit of V3a and V3b simply provides a *two-wire* anti-phase output of double-current signals driven from the *single-wire* double-current output of the teleprinter transmitting contact.

When the teleprinter is sending a mark, negative 80 volts is applied to the grid of V3a, and cuts off the anode current. The anode voltage of V3a is therefore practically the h.t. voltage. R7 and R8 form a potentiometer from V3a anode down to h.t. negative 150 volts. The voltage which appears at the junction of R7 and R8 is therefore *potentially* 190 volts positive referred to h.t. negative 150 volts, and is thus 40 volts positive to earth. This junction is connected via current limiting resistors R4 and R11 to the grid of V3b and the suppressor grid of V2. Diode CR2 will conduct and clamp the voltage applied to V2 suppressor grid to earth potential, and V2 will behave as a normal amplifier and send mark tone to the s.s.b. modulator input. V3b will conduct fully, its anode current for a 100 K load line being 2 mA. The anode voltage of V3b will thus fall to positive 50 volts referred to earth, since R2 will drop 200 volts. V3b anode is connected via the potentiometer formed by R5 and R6 to h.t. negative 150 volts, so that the voltage at the junction of R5 and R6 will be 100 volts positive referred to h.t. negative 150 volts, which makes it negative 50 volts referred to earth. This voltage is applied via the current limiting resistor R10 to the suppressor grid of V1, and cuts off the anode current of this valve, blocking the space tone oscillator output from reaching the s.s.b. modulator input. Diode CR1 does not conduct.

When the teleprinter sends a space, positive 80 volts is applied to the grid of V3a, via the current limiting resistor R3. This causes V3a to conduct to 2 mA and drops the anode voltage to positive 50 volts referred to earth, which makes the voltage at the junction of R7 and R8 negative 50 volts referred to earth, so cutting off V3b and V2. The anode voltage of V3b therefore rises to practically the full h.t. of

250 volts, and this raises the voltage at the junction of R5 and R6 to *potentially* 40 volts positive to earth. Diode CR1 conducts and clamps the suppressor grid of V1 to earth, and V1 behaves as a normal amplifier for the space tone, which is applied to the s.s.b. modulator input.

Whilst the voltages appear *potentially* at the junctions of R5 and R6, R7 and R8, these voltages will be reduced when positive to earth by the flow of grid and diode current which arises. This does not matter, because it is only necessary to bring the control and suppressor grids to earth potential to cause full conduction of the valves, and the current will adjust itself through the high resistances in circuit to keep the voltages applied to these points fractionally greater than earth potential. From this it will be seen that there is a great deal of margin available in the circuit operation.

Diodes CR1 and CR2 may be almost any type of small diode with a p.i.v. rating greater than 50 volts. The resistor R9 prevents the grid of V3a becoming open-circuited during the transit time of the teleprinter contact tongue, and its ohmic value is fixed to give a reasonable contact current and provide a suitable load resistance to spark suppression filtering in the teleprinter line output circuit. Although the teleprinter keying voltages have been shown as plus and minus 80 volts (the British standard for line telegraphy), V3a will obviously key satisfactorily at much lower voltages. The limiting voltages are negative seven for mark and earth for space. However, if a local record copy of the messages being transmitted is required to be printed on the teleprinter page, the voltages used on the transmitting contact will also need to be compatible with the requirements of the local record circuitry.

A switch to enable the sense of keying to be reversed is a useful facility. It would, for example, be necessary in the s.s.b. application to f.s.k. if the transmitter used lower sideband on 3·5 Mc/s and upper sideband on 14 Mc/s. It would be inconvenient to reverse the voltages connected to the teleprinter contact, because this would prejudice local record circuitry. The switch can either be arranged to reverse the tone oscillator outputs, or the suppressor grid connections to V1 and V2. In f.s.k. it is normal practice to use the higher frequency for mark.

RECEIVING FREQUENCY-SHIFT SIGNALS

Commercial f.s.k. receiving equipment generally takes the form of units containing the circuitry required between an output taken from a normal h.f. communications receiver, and the keyed d.c. input required by the telegraph machine to be driven. They are known as f.s.k. converters or adapters. Some early models take their input direct from the i.f. strip of the h.f. receiver and discriminate at the i.f. Others take their input from the audio output of the h.f. receiver and use tuned filters to identify and isolate the mark and space signals. Modern equipments take the receiver i.f. and convert it by heterodyne method to a frequency of 10 to 15 kc/s, after which it is limited and detected by a discriminator.

The design problems encountered when applying discriminators to f.s.k. telegraphy have already been considered in the section on Telegraph Distortion on page 11.4. **Fig. 11.8** shows the response curve of a good discriminator that would be suitable for receiving f.m. telephony. It will be seen that the curve reaches maximum amplitude in positive or negative sense for a given increase or decrease of input

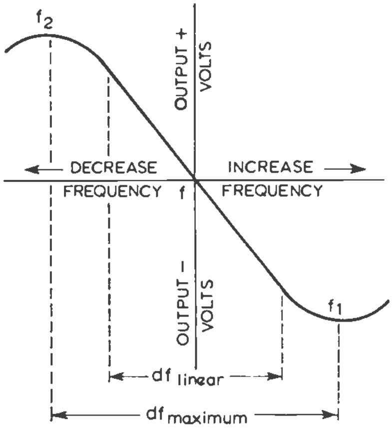


Fig. 11.8. Discriminator response graph.

frequency to the discriminator, the frequency difference (shift) between these points being indicated as df_{max}. Its telephony application, however, is limited to the extremes of the linear part of the curve, indicated as df_{linear} on the graph, because amplitude distortion must be avoided, and design of a discriminator for telephony reception is concentrated on obtaining a good linear response. For f.s.k. purposes where only the mark and space frequencies are of interest, provided the shift is known, the requirement is that the shape of the curve resulting from increasing frequency shall be similar and opposite in polarity to the shape of the curve resulting from decreasing frequency. It is preferable that the maximum response arises at the mark and space frequencies in use.

The frequencies of maximum output in a phase type discriminator are determined in a complicated manner by the relationships of the Q of the primary winding, the Q of the secondary winding, and the coefficient of coupling between them, for a given base frequency. There is thus a practical limitation related to f in the value that can be given to df_{max}.

The amplitude type of discriminator is rather different. In this case the two frequencies of maximum response are set by adjusting the two tuned circuits. If a linear response is required it is necessary to adjust the Q of the tuned circuits to give the correct shape to the individual responses, which again sets a practical limitation, but one which is not particularly worrying in the case of f.s.k. For good linear response the Q of the tuned circuits are:

$$Q_1 = \frac{f_1}{df_{max}} \text{ and } Q_2 = \frac{f_2}{df_{max}}.$$

If df_{max} is very small compared with f_1 and f_2, the Q of both circuits will be nearly equal, but if df_{max} is substantial compared with f_1 and f_2, the Q will differ accordingly.

From this reasoning it follows that when the input to the discriminator uses an audio frequency output from the h.f. receiver, the Q of the tuned circuits will either have to be extremely low if linearity is sought, or the response curve will be very peaky about f_1 and f_2. Low Q will result in a flat response curve and low output, and high Q will result in rapid deterioration of signalling if the transmitter shift is not set accurately.

The disadvantages of discriminating at the i.f. of the h.f. receiver have already been explained. Briefly they are that the low output realized in practice makes d.c. coupling into the following stage difficult, and the frequency stability problem more urgent.

For f.s.k. purposes, taking into account the practical limitations imposed by available components, the optimum frequency about which to carry out the discriminating function lies between audio and receiver i.f., and has been set in modern practice to be about 10 to 15 kc/s. This range of frequencies also permits the use of a driven flip flop or locked multivibrator as a limiter.

Many of the f.s. converters available on the surplus market are early types designed to receive a 455 kc/s i.f. directly, and to discriminate about that frequency. To overcome the difficulty of frequency stability they use an a.c. coupling out of the discriminator, and to overcome the difficulty of holding the teleprinter to mark during idle periods, they use a trick circuit known as the " mark restoring circuit." This is a simple time constant device, which charges up slowly and triggers a steady mark to the output circuit if no change of signal polarity is received over a period of several seconds. The effect is that when the transmitting keyboard is idle, the coupling capacitor between the discriminator and following stage charges up until the control voltage passed to the following stage is indeterminate. About this time the mark restoring circuit triggers and holds the teleprinter to mark. On arrival of the first space following the idle period, the mark restoring circuit is reset and signalling proceeds. However, in the meantime the coupling capacitor has become fully charged, and so the first space is mutilated. Since this will be a start signal, one can usually reckon that several characters will be mutilated. Due to the asymmetry of the signalling waveforms, the state of charge on the coupling capacitor will vary with the signalling, and characteristic distortion results.

Some of these converters use a " locked oscillator " as a limiter preceding the discriminator. The theory is that the oscillator locks to the receiver i.f., and it is the oscillator output which is fed into the discriminator, so that detection is entirely free from amplitude response. This is a very sound idea, but has its practical difficulties. The main one arises from the big amplitude variations of signal that are encountered in radio paths, and the difficulty of designing an inductance-capacitance type of oscillator which will have a dynamic range of locking voltage which may be as high as 100 : 1. To reduce the range required, the locked oscillator is usually preceded by a conventional saturation type limiting stage, and the combination works quite well. Whilst the limiter reduces all signal voltages to a low level, the locked oscillator serves as an amplifier and ensures that the signal fed to the discriminator is at reasonably high level. The peak-to-peak signal output voltage from the discriminator is usually of the order of six to eight volts.

Some commercial f.s. converters embody quite high voltage power supplies. One popular model uses an h.t. positive supply of 250 volts, and a negative bias supply of 516 volts, so that there is a potential difference of 766 volts in the equipment. Safety precautions against electric shock should be observed when putting fingers inside them!

The method of using the audio output of the h.f. receiver, by running the b.f.o. so that mark and space frequencies are heard as two audio tones, has already been mentioned. Circuits have been produced where only one filter is used tuned to one tone, on the basis that if the other tone is not coming through, the signalling condition transmitted must be that of the other tone, and the output relay should be switched accordingly. In effect, this arrangement changes the signalling system from f.s.k. to on-off keying, and whilst the argument is true enough, the effects of adopting this system make nonsense of using f.s.k. for the transmission. The operation is back to a touchy threshold control and weak interference wrecking strong wanted signals. In fact, the system has nothing to recommend it and should be avoided.

A Simple Frequency-Shift Converter

Fig. 11.9 shows a circuit for a frequency shift converter which embodies the best features of modern design, but restricts the circuit to essentials without any frills. To avoid the necessity of a chain of d.c. coupled stages to step the output voltage excursion down to a reference voltage equal to earth potential, the teleprinter magnet is returned to h.t. positive 150 volts. This saves an additional high voltage h.t. negative supply, and at least one valve stage. The circuit is intended for discrimination about 10 kc/s, and employs a flip flop limiter, amplitude type discriminator, low pass filter, squaring amplifier, and electronic double-current output relay. The transmit-receive switch can be a relay contact or one changeover contact on a multiple contact switch which also controls the transmitter and receiver. The switch enables the teleprinter to print a record of all messages transmitted, by diverting the input connection to the squaring amplifier from the filter output to the tongue contact of the teleprinter transmit contact.

T1 is a matching transformer suitable for the connection to the h.f. receiver output, and must be selected to give reasonably level response to mark and space frequencies. It is probably convenient to take the output of the h.f. receiver from the normal audio output, but attention must be paid to the frequency response of the audio stages if this is done. Consideration should also be given to the a.g.c. time constant and its degree of control in the receiver, as selective fading may invalidate the use of a.g.c.

The heterodyne mixing voltage to provide the signal centred on 10 kc/s can be injected at the receiver detector stage, but it is advisable not to use the c.w. b.f.o. for this purpose, as its level of injection is generally set very low. The point of injection should, in any case, be chosen so that the injection voltage does not block the a.g.c. line. Ideally, a balanced modulator should be used for the mixing process, so cancelling the injection voltage and preventing it reaching the a.g.c. detector.

Operation of the flip flop limiter is quite simple. When the polarity of the voltage applied to the grid of V2a is positive, that applied to the grid of V2b will be negative, as they are worked in push-pull. The positive grid of V2a causes the valve to conduct, drawing current through R10, the anode resistor. The voltage at the anode of V2a falls, and this is reflected via R9 to V2b grid. Due to the current through R7 and V2a, a positive voltage appears at the common cathode connection of V2a and V2b. V2b therefore tends to cut off,

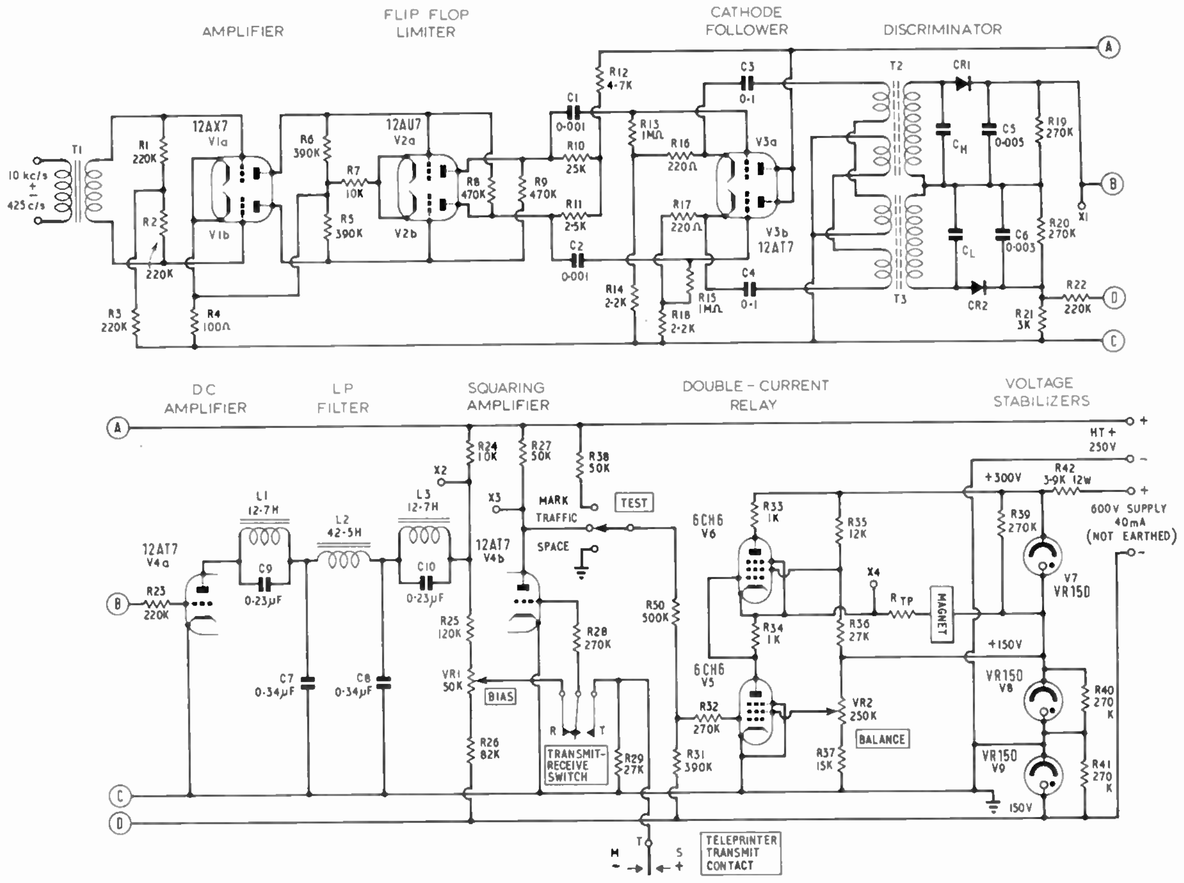


Fig. 11.9. Frequency-shift converter for reception of RTTY signals.

and its anode voltage rises. The rise is reflected to the grid of V2a via R8, so that V2a conducts more heavily, so driving the grid of V2b still more negative with respect to cathode. The effect is cumulative, and V2a triggers into saturation, whilst V2b cuts off completely. When the signal cycle reverses and appears as positive at the grid of V2b, and hence negative at the grid of V2a, the valve current conditions are reversed. The voltage developed across R7 remains constant, since one valve will always be fully conducting and the other fully cut off. Provided the signal voltage at its peaks rises above the threshold voltage necessary to commence the triggering action, the voltages appearing across R10 and R11 will always be those arising from saturation or cut-off of the valve, and so complete limiting action ensues. The output waveform is a square wave of constant amplitude at signal frequencies.

The design of the discriminator is much simplified by the use of Mullard Ferroxcube pot cores. Type LA4 is very suitable, and a maximum unloaded *Q* value of 280 at 10 kc/s can be realised. At 10 kc/s the value of inductance that will reduce the *Q* attainable by five per cent due to self capacitance of the winding is 100 mH, and this value should not be exceeded. A winding of 54 turns gives one millihenry, and as inductance increases as the square of the turns, 100 mH requires $54 \times \sqrt{100}$ turns = 540 turns. A suitable wire which will leave room for a small primary winding is 32 s.w.g. enamelled copper wire. The sense of the primary windings relative to the circuit connections is important.

Tuning of amplitude type discriminators is always difficult, since the mutual inductive coupling between the two tuned circuits, arising from the linking of the primary windings, causes pulling between the tuned circuits. The primaries of T2 and T3 are driven from a cathode follower push-pull stage to keep the circuit impedance low, and the primary windings therefore only need a few turns each, all of them being identical. The tuning of the secondaries is done by fixed mica capacitors. Calculated values for 100 mH coils are C_H = 2333 pF and C_L = 2766 pF.

The method used is to make up two stacks of capacitors totalling slightly less than these values to allow for production tolerances, and then to pad them with small mica capacitors until the tuned circuits resonate at the correct frequencies. After each adjustment, the other circuit should be rechecked and corrected until eventually both circuits are on tune. The use of close tolerance and high stability capacitors is obviously desirable.

When adjusting the discriminator, the test oscillator used must have good setting accuracy on the frequencies used, i.e. 10,425 and 9575 c/s. Calibration of the average cheap audio signal generator is unlikely to be satisfactory for this job. Amateurs who possess s.s.b. exciters can improve setting accuracy by modulating the exciter with 500 and 1350 c/s for the signal frequencies and tuning the receiver to give 10 kc/s output when the modulation frequency is set to 925 c/s. An alternative method is to build a 10 kc/s oscillator having a limited tuning range, and to calibrate it by beating its twentieth harmonic against known broadcasting stations in the long waveband.

If reception of shift frequencies much lower than 850 c/s is envisaged, some degree of linearity will be required in the discriminator response slope. This may be achieved empirically by loading the secondary windings of T2 and T3 with resistors selected to give a suitable shape to the slope. The values of C5/C6 and R19/R20, which form the filters for the signal frequency component of the discriminator output, are fairly critical. Too small a time constant will result in the signal bauds being modulated with the signal frequencies at the input to V4a. Too large a time constant will cause appreciable curvature to the baud transitions, and may result in telegraph distortion. Diodes CR1 and CR2 are not critical, and may be any type with low self-capacitance and p.i.v. rating of 100 volts or more.

The low pass filter is designed for a characteristic impedance of 10,000 ohms. There is no standing d.c. current through the inductors, since V4a is keyed fully between cut-off and saturation. Maximum current through the inductors is 12·5 mA when V4a is fully conducting. The total copper resistance of the three inductors in series should not exceed 1000 ohms. The signal bauds out of the filter will have their leading and trailing transitions curved by the filter action. It is therefore very important that V4b should be keyed symmetrically about its grid reference voltage, or telegraph bias distortion will result. VR1 enables the grid working point to be centred on the signal voltage excursion applied to it.

When the grid of V4b is made positive against earth by the signal polarity, its anode voltage will drop to positive 60 volts. R30 and R31 form a potentiometer between the anode of V4b and -150 volts, so that the voltage which appears at the junction of R30 and R31 is -58 volts with respect to earth. This cuts off V5. When the grid of V4b is made negative against earth by the signal polarity, its anode voltage will rise to practically the full h.t. (250 volts positive). The voltage that will then appear at the junction of R30 and R31 will be positive 16 volts against earth, and V5 will conduct fully. The test switch enables V5 to be keyed fully to mark or space artificially, and is useful to stop the teleprinter chattering on noise when tuning, as well as for setting up the output circuitry.

When V5 is fully conducting it draws current from h.t. positive 150 volts through the teleprinter magnet, R_{TP}, R34, towards earth. The grid of V6 is returned to the anode of V5, so that the current through R34 and the voltage drop so produced causes V6 grid to be negative with respect to the cathode of V6, and V6 is cut off. When the signal polarity changes and V5 is cut off, current ceases to flow in R34, and the grid and cathode of V6 are at the same potential, causing V6 to conduct fully. Current is therefore drawn through V6 from h.t. positive 300 volts to h.t. positive 150 volts, via R_{TP}, R33, and the teleprinter magnet. This current is flowing in the reverse direction to the previous current arising from V5 conduction, and so double-current energization of the teleprinter magnet is achieved. V6 screen grid is held to a steady potential by R35 and R36. The screen grid of V5 is held to an adjustable voltage which can be set by VR2. This control is used to balance the mark and space currents in the magnet circuit to be equal and opposite. Since V6 has its cathode operating above h.t. positive 150 volts it is as well to operate its heater from a separate supply to avoid the risk of heater/cathode insulation breakdown.

The h.t. positive 150, 300, and negative 150 supplies should all be stabilized, and VR150/30, 150C3 or OD3 regulators will serve to do this. A common 600 volts supply with the regulators connected in series across its output to provide the various voltage taps makes a neat arrangement. R_{TP} is chosen to set the magnet current to 20 mA, so the regulator valves can be worked well within their ratings.

Useful oscilloscope test points are indicated on the circuit by crosses followed by numbers. In all cases the use of an a.c. coupled amplifier in the oscilloscope should be avoided, as it will almost certainly introduce distortion of the waveforms presented. Since most of the test points are at quite high circuit resistance points, the input resistance of the oscilloscope should be several megohms to avoid shunting the circuits under observation.

Test point X1 monitors the discriminator output, and is very useful when tuning-in an f.s.k. signal. Multiple path signals and fading below limiter threshold can be clearly seen. Most operators monitor this point continuously during reception, and modern f.s. converters often have a built-in oscilloscope.

Test point X2 shows the signal shaping due to the filter, and will give a good idea of the extent to which noise is cleaned up by the filter, especially if X1 and X2 can be displayed simultaneously on a double beam oscilloscope.

Test point X3 shows the cleaned-up signals squared-up by the squaring amplifier, and forms a useful indication when adjusting the bias potentiometer VR1.

Test point X4 shows the output double-current signals into the teleprinter magnet, and can be used as a voltage balance indicator when adjusting VR2.

Attention to the polarity of mark and space voltages is necessary. In the circuit of Fig. 11.9 the output of the discriminator is positive for the higher frequency, which by convention is used for mark. A mark at the grid of V4a is therefore positive, so that the grid of V4b will have a negative voltage applied to it for mark. Since the grid of V4b also receives the transmitting signal voltages for local record purposes, it follows that the teleprinter mark output voltage needs to be negative with respect to earth, and the radio transmitter should therefore be arranged to send the higher frequency of the shift when the input to the keyed exciter is negative. A signal reversing switch which reverses the discriminator output connections may be incorporated in the converter.

Some method of checking that the h.f. receiver is correctly tuned to RTTY signals, apart from observing the page-printed copy on the teleprinter, is necessary. By far the best method is to display the discriminator output on an oscilloscope (Test Point 1), but a less effective method which is commonly used is to connect a centre-zero microammeter in series with a current limiting resistor across the output of the discriminator. On a reversals signal the pointer will vibrate about zero when the receiver is correctly tuned, but during signalling from a teleprinter will tend to show bias to one side or the other according to the mark or space weighting of the characters being received. The operator needs to become familiar with its behaviour, and to judge the tuning over a number of characters. The sensitivity of the meter needs to be high so that the series resistor is high, and does not seriously shunt the discriminator output resistance.

TELEGRAPH RELAYS

Special polarized relays such as the Carpenter type are often encountered in telegraphy practice. Like the double-current magnet they are completely balanced mechanically and side-stable. When the current is removed entirely from the coils, the contact remains switched according to the direction of the current prior to its removal, held by the permanent magnet providing the polarization. There is only one changeover contact, the fixed contacts being designated mark and space, and the moving contact known as the tongue. Generally there are two identical coils.

Circuit Applications

Fig. 11.10 shows a number of arrangements for telegraph relays. The circuit in (*a*) has coils *A* and *B* in series assisting, and requires the current in the coils to reverse direction for the contact to change over. It therefore accepts a normal double-current signal, and may be used to convert double to single-current signalling, or to convert a double-current signal working about one reference voltage to a double-current signal working about another reference voltage.

The circuit in (b) has each coil fed separately from a two-wire source. A positive current in coil *A* will operate the relay one way, and a similar current in coil *B* will change it over. Only one coil is energized at a time. It is used with circuits such as a flip flop, where one coil may be placed in series with each anode resistor. The relay then follows the switching of the flip flop, and is a means of switching relatively high current from the high resistance circuitry of the electronic device, providing a single-wire output.

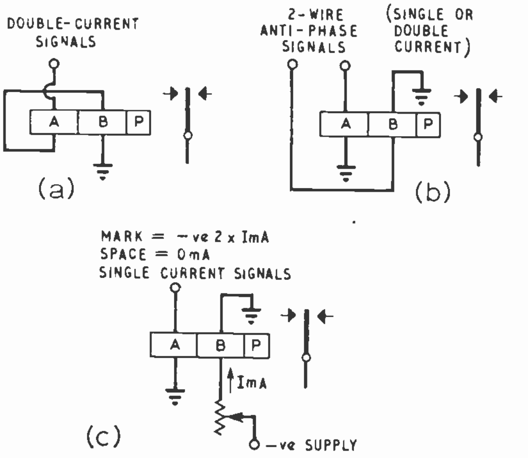


Fig. 11.10. Polarized telegraph relay connections. See text for applications.

Circuit (c) works entirely from single-current signals, and is the least sensitive of the three arrangements. A steady current is passed continuously through coil *B*, so that when no current flows in coil *A*, the relay will always switch one way. The signal current in coil *A* is set to be twice the current in coil *B*, so that the *change* in magnetic force acting upon the armature for operation from mark to space and from space to mark is always the same. Since the force arising from the continuous energization of coil *B* is always in opposition to the force arising from signal current in coil *A*, sensitivity is reduced, but the arrangement is better than using a relay with a spring return for the armature, because operation in each direction is effected by identical change of magnetic force, and the relay retains its mechanical balance. This circuit is used to change single to double-current signalling.

Adjustment

Telegraph relays are a common source of bias distortion, and their adjustment is critical. The mark and space contacts mounted on the tips of screws with fine threads are made adjustable. Ideally they should be adjusted dynamically using a test set as shown in Fig. 11.4. They can be set out of circuit approximately as follows:

(i) Back off both mark and space contacts until they are clear of the tongue contact.

(ii) Slowly advance one contact until the tongue switches away from it.

(iii) Advance the other contact until the tongue switches away from it.

(iv) Repeat the process, advancing each contact in turn to make the tongue just switch, until the tongue travel is correct for the type of relay.

(v) Fine up the adjustment with the contact screws, backing off one as the other is advanced by the same amount, until the tongue remains whichever side it is put. Use a small screwdriver tip to toggle the tongue back and forth, judging that the force required to toggle it is the same for each direction. Telegraph mechanics become highly skilled at this process, but it takes some practice.

The contact travel of a telegraph relay when correctly set should be of the order of 0·002 in., and this can be measured with a set of engineer's feeler gauges. Often the contact screws are calibrated around their knobs or heads, relative to the threads. When in good adjustment the relay should change over with half a milliampere flowing through either coil. The contacts will have some form of spring mounting to absorb the impact when the contact changes over, so that the contact will not bounce away.

Coil resistances are quite low. This is because the inductance of the coils must be low in order to allow a rapid build-up of the current through them, so keeping the contact transit time short. An important point in this respect arises from the fact that the time constant of an LR circuit decreases as R increases. To speed up the operating time of a teleprinter magnet or telegraph relay, a resistor may be inserted in series with the coil, and the applied voltage raised to keep the steady state current at its correct value. This is why the standard telegraph line voltage is

80 volts. The teleprinter magnet coil resistance is about 200 ohms, so that it is only necessary to apply 4 volts in steady state across the coil to pass 20 mA through it. By using 80 volts with a series resistor of 3800 ohms to limit the current, the magnet operation is much speeded.

Pulse Circuit Application

Since these relays are side-stable, they may be operated by short pulses at baud length intervals, and can thus be used to regenerate signal pulses from electronic gating circuits into full-length bauds. For very short pulse circuits a flip flop can be used similarly.

TELEPRINTER OPERATING MODES AND FACILITIES

Apart from the fact that there is a common motor, the teleprinter receiving and transmitting mechanisms are independent of each other. When a single teleprinter is operated as a terminal equipment it can be used in one of three ways.

Simplex Mode. The teleprinter is used either for sending or receiving only. A teleprinter used solely for receiving may have its keyboard removed. The Creed Model 8 is such a teleprinter and has no keyboard.

Semi-duplex Mode. The teleprinter is used for sending and receiving, but not at the same time. There is an automatic send-receive switch which is operated by the keyboard. Normally the receiving line is connected to the magnet circuit, but when a character key is pressed, the magnet circuit is switched to monitor the transmit line. Consequently both incoming and outgoing messages are printed on the page, so that simultaneous transmission and reception would cause mutual interference.

Duplex Mode. The send-receive switch is not in circuit, and no record of transmitted messages is made. The keyboard and receiving mechanisms are independent, and transmission and reception can be carried on simultaneously.

Amateur Operation

The use of the terms simplex, semi-duplex, and duplex, is not generally agreed in definition. Semi-duplex, described above, is commonly referred to as simplex, and duplex is defined as requiring two teleprinters at each terminal, one giving local record of messages and used for sending only, the other without keyboard and used for receiving only. The latter arrangement is, in fact, two simplex circuits working entirely independently, their only point in common being their physical proximity. The mode requires separate sending and receiving positions, and is obviously not adaptable to amateur radio practice.

The amateur RTTY operator will probably find the semi-duplex mode most suitable. Just as most c.w. operators prefer to monitor their own keying, so it is helpful to be able to read teleprinter messages as they are sent. A page teleprinter at the receiving end requires the sending operator to put in carriage return and line feed signals at the correct points. Tape type teleprinters designed for semi-duplex working have a lamp which glows when a full line of type has been transmitted, but Creed Model 3 teleprinters, designed for duplex working and printing on gummed tape, give no indication.

The amateur has to switch his radio equipment from transmit to receive, and vice versa. There is little point in making use of the automatic changeover switch in the teleprinter, which only switches to transmit for the duration of each character sent. The receiving operator would find a contact very disconcerting if the carrier were cut every time the distant operator paused in his keyboard typing for a moment. It is therefore better to make use of the radio equipment transmit-receive switch to change the teleprinter magnet circuit over from receiver output to teleprinter transmit contact in order to monitor outgoing messages. Fig. 11.9 indicates a simple method of doing this.

Extra Teleprinter Facilities

Creed Model 7 teleprinters usually embody a mechanically operated switch which cuts power to the motor if no signals are received for a period of one and a half minutes. The first space received trips the switch and starts the motor again, but the first few characters will, of course, be mutilated as the motor accelerates to correct speed. For sending, the switch may be tripped manually by pushing a button in the right-hand side of the teleprinter cover. The purpose of this switch is to reduce motor wear when teleprinters are left switched on for long periods for reception of intermittent traffic, and it may prove more of an annoyance than a help to the amateur operator. It can be short-circuited.

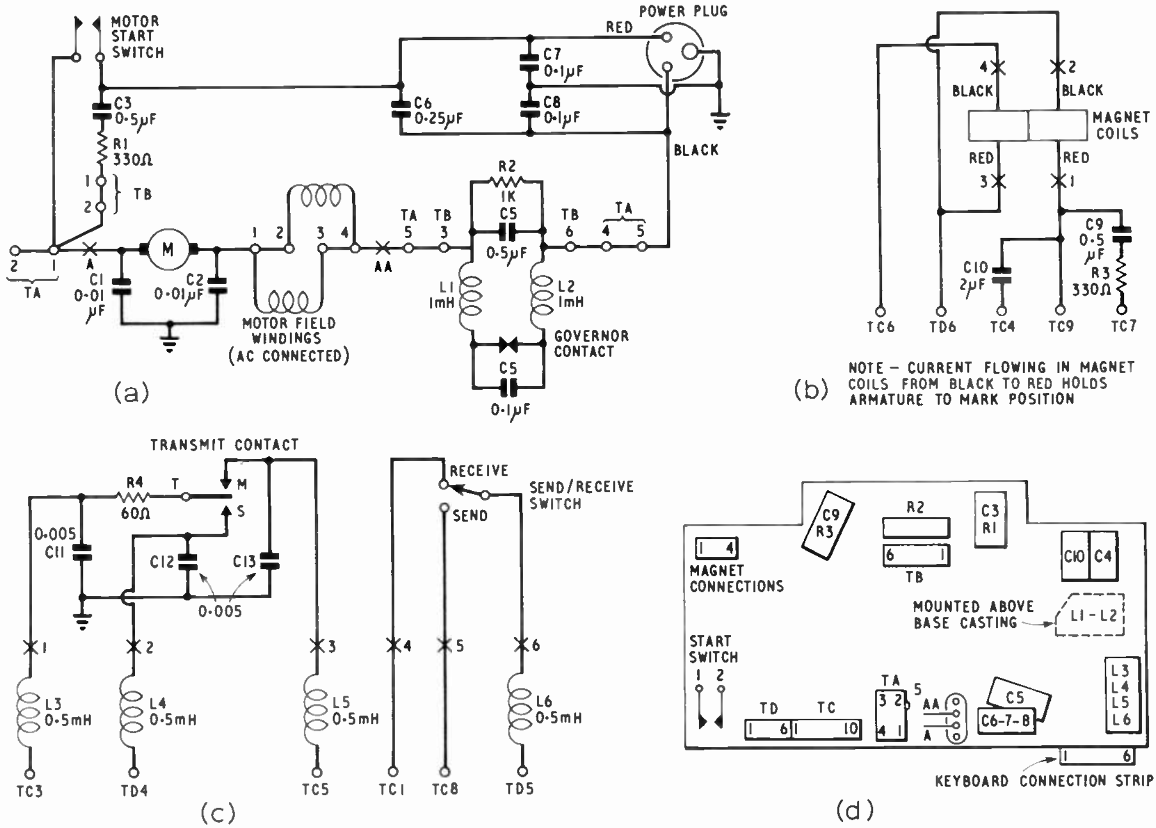
The "answer-back" facility is an ingenious device, but quite unsuited to amateur RTTY operation. If the upper case of character D is received by the teleprinter receiving mechanism, it triggers a mechanism in the keyboard which transmits a prearranged call-sign or identification code. This is most useful for Telex line working, because the operator making an automatic Telex call can check that he is through to the right subscriber if the distant teleprinter is unattended, and then send his message and be reasonably sure it will be printed. On RTTY it is a complete nuisance. Whenever a character is mutilated on reception and is interpreted as the upper case of D, the receiving teleprinter immediately switches to transmit and sends a series of signals which inhibit reception for their duration. To avoid this happening the operating linkage can be removed.

Another facility which is of little use to the amateur is the J-bell alarm. If the upper case of character J is received, a pair of contacts are closed, and these can be wired to an alarm bell circuit. This allows the operator of an intermittently attended teleprinter to be called by the distant operator, but the facility is obviously only valid if the connection between the teleprinters is continuously maintained.

TELEPRINTER CIRCUITS

External Connections

Fig. 11.11 shows the standard wiring for a Creed Model 7 teleprinter. Detachable units such as the motor and keyboard, have fixed built-in plugs and sockets which automatically pick up the circuit connections when the units are placed in position and bolted down. These are indicated by points marked X on the diagram. Older teleprinters use a nine-way plug on a flexible cord, mating with a GPO Jack 52 for the line and signalling power supplies. Modern teleprinters use a twelve-way plug of the Jones type. The motor



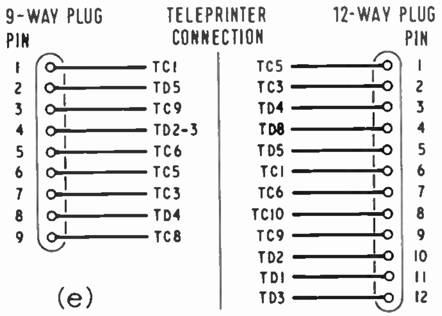


Fig. 11.11. Wiring and component layout of Creed Model 7 teleprinter.
(Courtesy of Creed & Co. Ltd.)

connection is brought out on a separate three-core cable terminated in a three-pin plug.

Motors

Creed teleprinters are normally equipped with an a.c./d.c. motor for voltages between 85 and 250 volts, the working range for any motor being plus or minus 10 per cent of the nominal voltage marked on it. Power rating is 1/25 h.p. The field coils are brought out to a connection block on the motor, and should be connected in parallel for a.c. and series for d.c. working.

Many surplus machines have been used on d.c. supplies, and the strapping of the field windings should be checked before connecting power when a surplus teleprinter is acquired. Motors rated for d.c. supplies of less than 50 volts require shunt, instead of series, governing. The armature winding and the field coils are in parallel connection, and the governor resistor is placed in series with the field windings. A special type of governor is required.

Speed Governor

The governor holds the motor to within one per cent of its rated speed by means of a centrifugally operated contact. This opens when the speed exceeds 3000 r.p.m. and closes when speed falls below this figure. A heavy duty 1000 ohm resistor, R2 in Fig. 11.11 (a), is connected across the contact so that it appears in series with the motor when the contact is opened. As a result the motor is continually hunting between just over and just under its rated 3000 r.p.m. The contact is switching a fair amount of power into a reactive load, and L1/L2 and C4/C5 form a spark suppression circuit to protect the contact and limit radio interference.

Electrical Noise Suppression Circuits

Electrical noise generated by the teleprinter is quite a problem in RTTY service. Much attention must be paid to the suppression of commutating and contact signalling circuits, or the operator will find his radio reception is being heavily jammed by his own teleprinter equipment. Even the electrolytic action arising from the rubbing together of many metallic parts in the teleprinter can cause interference if a "hot" aerial wire passes close to the teleprinter. It is essential to get the aerial well away from the telegraph apparatus, and to use coaxial cable for some distance.

A common earth wire to teleprinter and h.f. receiver can display standing waves arising from interference noise and cause a great deal of obscure trouble. It is a good idea to provide a direct capacitance earth to the teleprinter by connecting it to a large sheet of metal laid flat on the floor by the shortest possible wire. The metal sheet should also be connected to the mains earth as a safety precaution, but its capacitance to true earth will short-circuit interference voltages that may otherwise get back into the h.f. receiver input via the earthing system. Mains wiring in the vicinity of the equipment should be screened and filtered. Not all teleprinters available to the amateur have the suppression devices shown in Fig. 11.11 built-in, so a check of these is well worthwhile.

The suppression of the transmit contact by capacitors C11, C12, C13 and coils L3, L4, L5, shown in Fig. 11.11 (c), can give rise to telegraph distortion troubles. They are intended for line use with the standard 20 mA line current. If the terminating resistance of the radio transmitting and monitoring equipment is too high, their time constant may be excessive and cause too much curvature to the leading and trailing baud transitions. The obvious way out of this difficulty is to load the transmit wire with a resistor to reduce the circuit resistance, and hence the time constant. This will increase the signalling contact current and, potentially, the noise, so that the contact should not be loaded more than is strictly necessary to make the baud shaping tolerable.

In Fig. 11.11 (b), TC7 should be strapped to TC6 for double-current working. This places C9 in series with R3 across the magnet coils to limit back e.m.f. effects when signalling into the magnet.

TELEPRINTERS

By far the most common type of teleprinters used in the British Isles is the range manufactured by Creed & Co. Ltd. A fair number of American Teletype machines went into service during World War 2, and many of these have found their way onto the surplus market.

Creed teleprinters use governed motors that are independent of supply frequency, but many Teletype machines have synchronous motors which may be for 50 or 60 c/s operation. A Teletype machine fitted with a governed motor can be adjusted to run at 50 bauds, and they are very reliable and robust machines. Their governed motor runs at 2100 r.p.m. and requires speeding up to make it compatible with the British and Continental standard signalling speed of 50 bauds.

The Creed machine has a rotating typehead driven by a complicated clutch which allows the head to rotate so that the required character is presented to the type hammer. Letter feed of the paper is accomplished by a moving carriage running on a rail and carrying the complete roll of paper. With a full roll its mass is considerable, so that when it receives a carriage return signal after a full line has been printed, it strikes its end stop with quite an impact, and this is damped by an air piston device.

The Teletype machine has a type basket similar to that on a typewriter, and the whole basket moves to provide letter feed. The carriage for the paper roll is fixed.

Creed & Co. Ltd. not only produce a range of teleprinters for different applications, but during the years the various models have been improved, so that sub-assemblies on models designed for the same service may differ substantially according to date of production. The Model 7 teleprinter is most suited to amateur application amongst the older types which are likely to be available. This model can be fitted with a page carriage, or with a tape printing attachment, which provides a narrow gummed tape with the messages printed on it in a long continuous line. The purpose of the latter is to allow the messages to be stuck to pages of any size in the most convenient form. The amateur will undoubtedly find the page carriage is most convenient for his purpose.

There are over two thousand parts in a Creed Model 7 teleprinter, and it is obviously outside the scope of this chapter to give a complete and detailed description of its working and adjustments, particularly with regard to the differences between the various versions of the same model that may be encountered in practice. The manufacturers produce very good instruction books. The following information refers to the Creed Model 7 teleprinter only.

Speed Adjustment

Correct governed speed of the motor is the first requirement. Early teleprinters have a large white spot painted on the typehead spindle gear wheel, just above the keyboard and facing the operator. This rotates quite slowly. A stroboscope consisting of a shutter mounted on a reed which may be made to vibrate at its resonant frequency, so opening and closing the shutter, is used to view the spot. If it appears stationary, speed is correct but if the spot appears to move clockwise or anti-clockwise the speed is fast or slow respectively. When the speed discrepancy is large, this is a difficult and frustrating device to use. Modern teleprinters have five equal white spots painted around the circumference of the governor cover, and these are viewed through a shutter consisting of two light metal plates mounted at the tips of the arms of a 125 c/s tuning fork. Each plate has a slot in line with the fork arms, and as the fork oscillates the slots become briefly coincident at the resonant frequency of the fork.

If the amateur's machine is of the type first described, it is recommended that a strip of white paper ½ in. wide be cut to the exact length of the governor circumference, and divided into ten equal parts, five of which are coloured black. This can then be stuck around the governor with a powerful adhesive.

A cheaper and more readily obtainable indicator than the 125 c/s fork can be made by constructing an oscillator checked against 50 c/s mains frequency with an oscilloscope, and using its output to drive a small neon lamp. When the neon is held near the governor cover in semi-darkness the white divisions will appear stationary if speed is correct, or move slowly in the direction of the motor rotation if speed is fast, and vice versa.

The speed of a newly acquired machine may be quite a

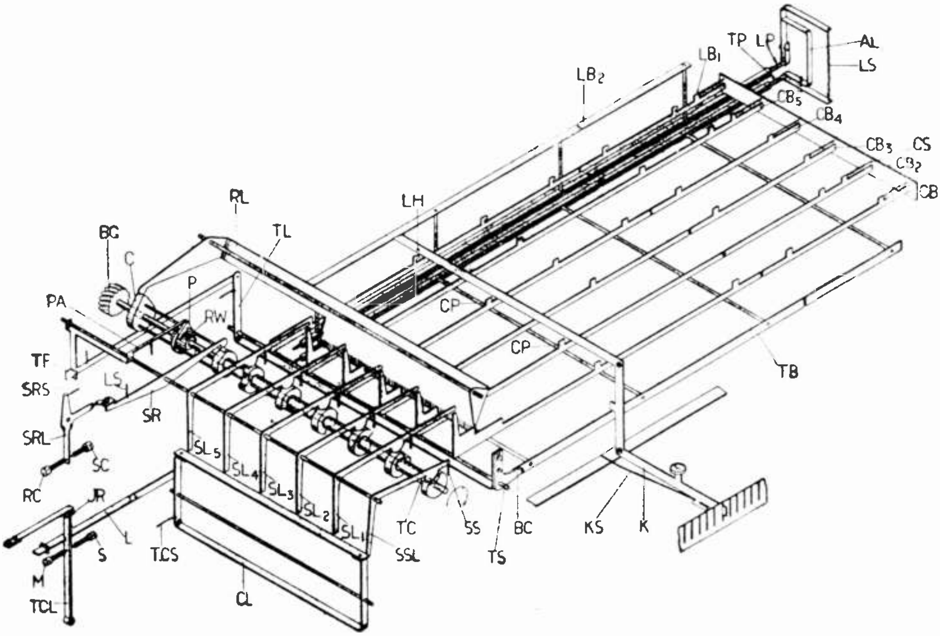


Fig. 11.12. The Model 7 Teleprinter Transmitting Mechanism. *(Courtesy of Creed & Co. L'd.)*

long way out, in which case it may be outside the range of the stroboscope. It is possible to find a false speed setting which holds the spots steady due to some harmonic relationship with the correct speed. Setting the speed approximately correct can be done by counting the number of characters printed in a measured time, with the teleprinter continuously transmitting one character. To do this, press one character key and hold it down, at the same time pushing up the keyboard clutch pawl abutment so that the clutch cannot latch. Sixty characters can be printed in a single line on a page teleprinter, and since each takes 150 milliseconds, 60 will take nine seconds.

The receiving and sending mechanisms both operate from the same motor and will therefore always be in synchronism with each other, irrespective of actual speed. A speed fault will not show up as mutilations on the local record printed.

Speed adjustment is made by removing the governor cover, which is held by a single screw. Increasing the centrifugal spring tension by the tensioning screw at its end increases speed, or slackening it decreases speed. The condition of the governor contacts is important, both from the point of view of smooth control and radio interference. If pitted, they should be rubbed flat with fine emery paper.

The Transmitting Mechanism

The principle of operation from the keyboard is illustrated in **Fig. 11.12.** There are five combination bars (CB), one for each element in a character. These bars have a series of projections along their upper edges arranged so that, if a character key is pressed, the keybar (K) associated with it will impede movement to the right of such combination bars as have projections at that point. The combination bars are each spring-loaded towards the right, but are held against the springs by the common returning lever (RL) whilst the keyboard is at rest.

The vertical member of each keybar has a projection which forces down the trip bar (TB) as soon as a key is pressed, and this causes the trip lever (TL) to move in a clockwise direction, so lifting the pawl abutment out of engagement with the one-revolution clutch pawl (P).

The one-revolution clutch is shown in **Fig. 11.13**, and its purpose is to provide one rotation of the transmitting cam shaft (TC) whenever a key is pressed. Its latching action prevents a second rotation in the event that the key is held down longer than the time taken to transmit the character. Operation of the keybar (b) forces down the trip bar (a),

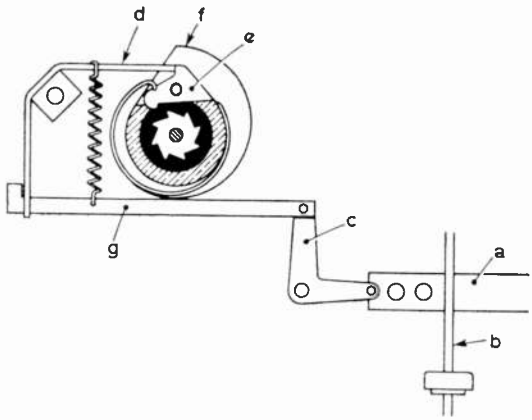


Fig. 11.13. The Model 7 one-revolution transmitter clutch.
(Courtesy of Creed & Co. Ltd.)

and the bell-crank (c) moves in a clockwise direction, causing the pawl abutment (d) to move in an anti-clockwise direction, so raising its tip out of engagement with the pawl (e). The pawl therefore drops down under the pressure of its spring and engages a tooth on the ratchet wheel, which is continuously rotating under motor drive. This causes the cam shaft to rotate. In rotating, the cam (f) forces the link (g) between the bell crank and the pawl abutment out of engagement with the pawl abutment, which therefore moves back into position to arrest and trip the pawl out of engagement with the ratchet when the single revolution has been completed. If the key is held down after this time, the pawl abutment will not again release the pawl because the link remains out of engagement, and will not re-engage the pawl abutment until the trip bar has been released by the raising of the keybar when the character key is released from pressure by the operator.

Reverting to Fig. 11.12, when the camshaft starts to rotate, the common returning lever (RL) is allowed to move in an anti-clockwise direction on its spindle by the profile of its cam (C), thus releasing the combination bars, which can then move to the right under their spring tensions. Due to the projections on the combination bars, some will move freely, whilst others will be arrested by the keybar, according to which key has been pressed. The combination bars therefore set up the character combination by their relative positions. The process which follows is that of transforming this simultaneous condition of five elements into a sequential string of signals on the single wire output from the transmit tongue contact.

As the cam shaft rotates, the shape of the front cam on the shaft forces up the start-stop lever (SSL), which allows the common operating lever (CL) to move in a clockwise direction on its spindle under the tension of its spring (TCS). The link (L) moves to the right and pulls over the transmit contact tongue (TCL) to space, thus sending a start signal. The tongue has a jockey roller (JR) giving a toggle action to the contact, which improves its transit time and provides a firm contact pressure.

Each of the element cams has a dimple in its profile, and these dimples are staggered in relation with each other. When the dimple on the first cam becomes coincident with the projection on the first selector lever (SL1), this lever will drop under its spring tension, provided the combination bar has moved to the right so that its tip does not impede the movement of the selecting lever. Dropping of the selector lever will force the common operating lever to move in an anti-clockwise direction on its spindle, and its link will switch the tongue contact to mark. If the selector lever is prevented from dropping by its combination bar, the contact will remain at space.

This process is repeated for each element in turn, as their individual cams offer movement to the selector levers. After the combination has been transmitted, the front cam profile again allows the start-stop lever to drop, so forcing the common operating lever to mark for the stop signal. The clutch latches at the completion of one revolution, and the cam at the rear end of shaft causes the common returning lever (RL) to reset the combination bars.

There are two locking bars associated with the keyboard. The first (LB1) moves and locks the keybar as soon as a character key has been pressed, so that the key will remain operated for the duration of the character transmission, irrespective of when the operator removes his finger from the key. The second (LB2) moves forward when a key is pressed and takes up position between the operated keybar and all the others, thus preventing operation of a second key whilst a character is being transmitted. A fast typist can feel the action of this locking bar holding back the speed of typing.

The modern " N " series keyboards differ from the type described above. They are available with three or four rows of character keys. In the latter case figures are on a separate row, just like a typewriter keyboard, and an interlocking arrangement prevents a key in either upper or lower shift case being pressed unless the appropriate shift key has been pressed previously.

The Selection and Printing Mechanism

Fig. 11.14 illustrates the selection and printing mechanism. The movement of the armature extension of the receiving magnet (RA) causes the trip shaft (TS) to move on its spindle, clockwise for a space and anti-clockwise for a mark. The trip shaft is linked to the pawl abutment (D) of the receiving cam clutch, which works in rather a similar manner to the transmitter clutch. The first space received trips the pawl and the cam makes one revolution.

The striker blade (SB), which is a thin steel spring leaf, is arranged to oscillate in the horizontal plane by the movement of the finger setting lever (STL), running in cam slot T1. At the same time a vertical oscillatory movement is imparted to the striker blade by the movement of the trip shaft, which is under control of the magnet, and whose movement thus follows the marks and spaces of the character combination being received. The striker pin (SP) is held in position by the traversing link (L), and the latter is driven by the traversing lever (TL), running in cam slot T3, which causes the link to move parallel to the trip shaft, carrying the striker pin with it. The timing of the traversing link, relative to the striker blade, is such that the striker pin is in line with one of the selector fingers (F1 to F5) every time the striker blade moves towards it. If the striker blade is in the upper position of its vertical plane movement, occasioned by a mark, the edge of the blade will strike the striker pin and move it forward, thus pushing back the coincident selector finger. When a space is being received, the striker blade will be in its lower position and will pass beneath the striker pin, so that the coincident selector finger will not move. In this manner the combination of the character is recorded on the positions taken up by the selector fingers.

So far the operation of the mechanism has reached the stage of converting the sequential movements of the magnet armature into five steady state conditions of the selector fingers. The next step is to convert these five separate conditions into one of the 32 possible conditions which together they provide. This is done by the translating mechanism. Five combination discs (C1 to C5) are mounted in an assembly, each able to move radially. Around their peripheries are a series of slots, and around the assembly are grouped 64 bellcranks (B), each tensioned by its own spring (BS) towards the assembly. The combination disc setting lever (CSL) is driven from slot T5 in the cam, and raises the spindle on which the five selector fingers pivot, after all the fingers have been set to the combination. Simultaneously the bellcrank lifting lever (BL), operating from cam slot T4, pushes the sliding sleeve (BC) up against the bottom ends of the bellcranks, lifting them and so freeing the combination discs from their pressure. The combination discs are spring-loaded and will rotate if their extension arms are not impeded by the selector fingers. Those selector fingers which have been pushed back by the striker pin into the mark position will obstruct the combination disc movement, whilst those which are in space position will be clear of the combination disc extensions, and those discs will move. For any combination there will always be one channel formed by a slot in each combination disc in line with all the others, allowing one of a pair of adjacent bellcranks (upper and lower case of the character) to drop into the channel so formed. A disc mounted at one end of the combination disc assembly and slotted regularly about its periphery (SC), is set in one of two positions by either the letter-shift bellcrank (LSB) or the figure-shift bellcrank (FSB), and the positioning of its slots determines which one of the character bellcranks will drop into the channel formed by the combination discs.

At the time the bellcranks are lifted by the sleeve (BC), the typehead clutch is engaged. This is a friction device which drives the typehead round until it is stopped by the position of the selected bellcrank. In this position, the character set up by the combination is presented to the type hammer (HH) ready for printing. By this time, however, the cam has almost completed its full revolution, and the actual striking of the type by the hammer, initiated by cam slot T2, takes place during the next selection process just before the bellcranks are lifted for setting of the combination discs.

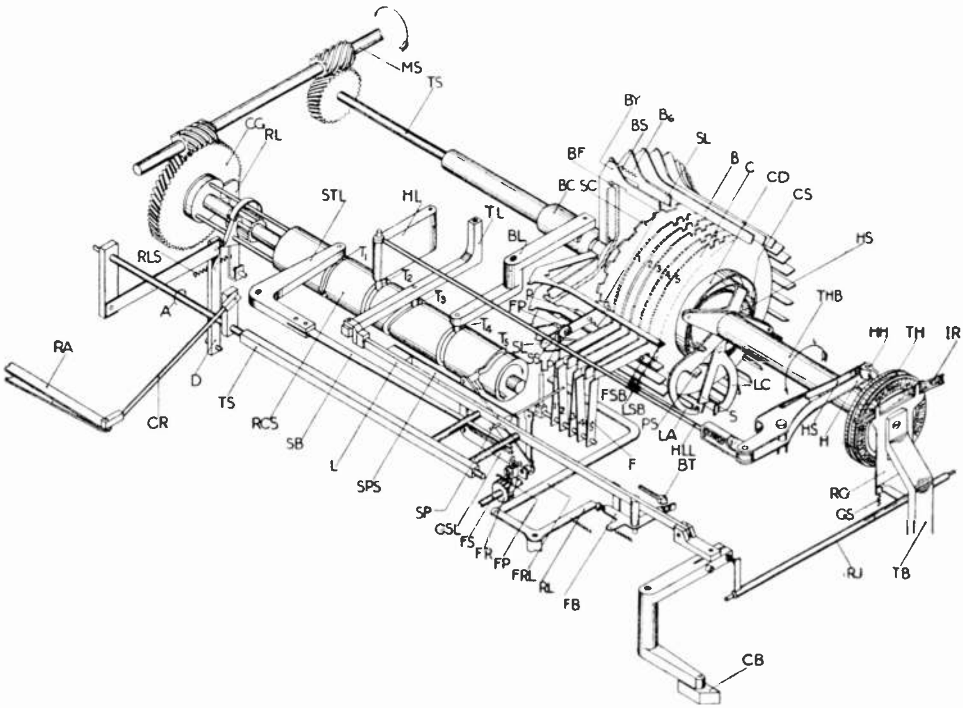


Fig. 11.14. The Model 7 teleprinter selection and printing mechanism.
(Courtesy of Creed & Co. Ltd.)

Consequently the printing is always one character behind the transmission.

Letter feed of the page carriage is accomplished by the action of the feed lever (CB), driven by the traversing link (L). Combinations resulting from letter-shift, figure-shift, carriage-return, upper case of J (bell) or all spaces, must not cause the carriage to feed. The bellcranks proper to these functions operate control levers, all of which operate a common non-feed lever, which inhibits operation of the feed mechanism for such non-printing combinations.

A pawl (FP), mounted at one end of the combination disc setting lever (CSL), steps a ratchet wheel (FR) to provide the feeding action of the inked ribbon.

Teleprinter Maintenance

A few drops of medium lubricating oil should be applied to all oil cups and holes, and to the selecting fingers. Before oiling, dirt and surplus oil should be cleaned away. It is important not to over-oil the machine, as surplus oil will pick up dust and become abrasive. The carriage platen spindle should be kept clean and lubricated sparingly with thin oil occasionally, but the air piston should not be oiled. Frequency of oiling should be about every hundred hours of running time. The latch cam of the typehead clutch should be greased at the same time. Ball races should be lubricated every 1000 hours.

TABLE 11.2

Tools for Teleprinter Maintenance

Set of instrument screwdrivers.
Screwdrivers with $\frac{1}{8}$, $\frac{3}{16}$, $\frac{1}{4}$ in. blades. (Keep blades in good condition.)
Steel tweezers and wiring pliers.
Sets of flat and box spanners, 0 to 8 BA.
Dentist's mirror and watchmaker's eyeglass.
Spring-type clutch gauges, 0 to 4, 0 to 16 oz.
Spring-type tension gauges, 4 to 24, 10 to 80 gm.
Set of engineer's feeler gauges, 0·0015 to 0·025 in.
Non-fluffy duster and chamois leather.
Camel hair brushes for oiling and cleaning.
Toothbrush, for cleaning type out.

Operation of the striker blade (sometimes called the finger setting blade) can be checked by turning the motor by hand and observing the blade in relation to the striker pin, whilst operating the magnet manually. This will show if there is any tendency for the blade to miss the pin on marks. The blade should be straight if it is to strike the pin squarely. In time the striking edge of the blade wears, and it has to be changed. Application of a little Oildag (a graphite preparation in oil) to the striking edge will prolong its life, but it should be applied very sparingly.

The magnet may be rotated slightly on its mounting by means of an adjusting screw at its rear. With the operating link disconnected from the armature extension, the magnet should be positioned so that equal force is required to push the armature extension from mark to space and from space to mark. If a spring balance type of clutch gauge is available, it will provide an easy means of measuring the force required, and applied to the tip of the armature extension, enables a very accurate setting to be found. With the operating link reconnected, the force required to operate the magnet armature should still be balanced, but will be rather less because of the action of the striker blade. If the force required is unbalanced, the striker blade should be *set* from the point of its attachment (*not bent or twisted*) to correct the bias.

The travel of the transmitting contact tongue should be 0·006 in. When the tongue is operated to the right, there should be a clearance of from 0·004 to 0·010 in. between it and the right-hand edge of the gap in the operating link (L in Fig. 11.12). The jockey roller may be positioned by slackening the fixing screws securing its bracket, and adjusting the eccentric screw just to the right of the jockey roller. The adjustment is correct when equal force is required to operate the tongue from mark to space and from space to mark. In the centre of its travel the tongue should be truly vertical.

It will be appreciated that the accurate response of the machine, both sending and receiving, depends to a great extent on the accuracy of the cam profiles. The only cure for worn cams is to change them. End-play or slop of the cam shafts will adversely affect performance, as will wear on all contacting surfaces where mark and space sensing is carried out. The majority of the moving parts are spring-loaded, so it is essential that the parts move freely under the tension of their springs when called upon to do so. Old and gummy lubricant, dust and fluff, are the worst enemies of light mechanisms, and owners are advised to cover their machines when not in use, and to keep them in warm, dry rooms at all times. Pure carbon tetrachloride applied with a good quality camel hair brush will wash away old lubricant, and will itself evaporate quickly. It is a good idea to apply fresh lubricant with a camel hair brush too.

The carriage is a separate unit with a hinge at one end and a latch at the other. It may be swung open on its hinge for access, and can be removed from the teleprinter simply by lifting it up off its hinge. When closing the carriage to latch it to the teleprinter, it should not be slammed shut. Care should be exercised to see that the various functional levers linking the teleprinter mechanism to the carriage engage correctly.

Teleprinters are well made machines which deserve careful treatment. Adjustments must be accurately carried out, and owners should be absolutely sure that an adjustment is really necessary before attempting to do it. All settings are quite logical and take into account the functions of the parts concerned. Too little pressure or tension will result in erratic operation, and too much pressure or tension will increase wear and shorten life of the parts. Failure of a part to move correctly is more likely to be due to frictional obstruction, arising from dirt or lack of lubrication, than lack of spring tension.

CHAPTER 12

PROPAGATION

THE establishment of radio contact between widely separated stations depends not only upon the purposeful activities of the operators but also upon the vagaries of the propagation characteristics in the intervening space. One of the most intriguing and exciting aspects of amateur communication is the variability of this medium, for although much research has been conducted in the study of radio-wave propagation much is still unknown and there is a wide field that remains open to further discovery.

Strictly speaking, the study of propagation must cover everything that happens to a radio wave from the moment when it leaves the transmitting aerial to the moment when it arrives at the receiving aerial, and this would include such phenomena as the reflection from buildings and hillsides and the absorption in forests, but the subject is so vast that only the major effects can be discussed here. These are related mainly to the electrical characteristics of the atmosphere which in common with other meteorological conditions are constantly changing, and because they have such a dominant influence on radio communication the interested amateur will be amply repaid by his efforts to understand them.

Radiation

The radiation from an aerial may be regarded as a succession of concentric spheres of electric force or strain of ever-increasing radius, moving outward from the aerial at a constant speed. This distribution of electric force surrounding the aerial and extending over what are sometimes astonishingly great distances is known as the *electric field*. For many purposes it is convenient to suppose that an electric field is made up of " lines " of force, and the direction of these lines will indicate the direction of the electric force in any particular part of the field. The intensity of the field can be represented by the concentration of the supposed lines of force.

In the spherical field produced by a transmitting aerial the lines of force will have a form similar to the lines of longitude on a globe, the aerial being collinear with the polar axis. At increasing distances from the aerial the lines of force will approximate to parallels over sufficiently large areas of space that to a practical receiving aerial they will present a uniform distribution of electric field over the region occupied by the capture area of the aerial.

At right angles to the electric field and inseparable from it, is the *magnetic field*, which is likewise conveniently supposed to be made up of lines of magnetic force. These two forces form a *wave* which travels outward at approximately the speed of light, i.e. 186,000 miles per second (300×10^6 metres per second). The exact velocity depends on the dielectric constant of the medium through which the wave passes.

Since the current in the aerial is alternating in direction at a high frequency the electric and magnetic fields are similarly alternating at the same high frequency, and the distance in space between successive peaks of field intensity will vary according to the rate of alternation of current in the aerial. The distance along the direction of propagation between two points at which the instantaneous field intensity is equal in phase and practically equal in amplitude, is called the wavelength λ. This is related to the frequency of alternation f by the simple equation—

$$\lambda \text{ (metres)} = \frac{\textit{velocity of propagation} \text{ (m/sec.)}}{f \text{ (cycles per second)}}$$

Taking the velocity as 300×10^6 metres per second, this relationship becomes—

$$\lambda \text{ (metres)} = \frac{300}{f \text{ (Megacycles per second)}}$$

The amplitude of the field intensity at the two points separated by one wavelength would be exactly equal but for the gradual fall in intensity as the wavefront moves further from the transmitting aerial and the power per unit area decreases.

A sinusoidal voltage in the transmitting aerial is continuously varying through a peak value in one direction, falling to zero, rising to a peak in the reverse direction and falling to zero again in each cycle, and in an exactly analogous manner the field strength at any given point remote from the aerial will undergo a cyclic change. How this induces a corresponding sinusoidal current or voltage in a receiving aerial is illustrated in **Fig. 12.1**(a). Spherical waves represented by the curved lines are radiated from the transmitter *T* and reach the point occupied by the receiver *R*. The sketch gives an instantaneous view of the field of electric force at the moment when the wave-front corresponding to a voltage peak at the transmitter has just reached the receiving aerial, and at this instant the electric and magnetic fields in the small section of the spherical wave-front in which the receiving aerial can be represented by **Fig. 12.1**(b). A quarter of a cycle later the front corresponding to zero intensity will reach the receiver, and after another quarter-cycle the wave-front corresponding to the opposite peak will arrive: at this moment the electric and magnetic fields will be reversed as shown in **Fig. 12.1**(c). As the succession of wave-fronts passes the receiving aerial so will a sinusoidal current be induced in it.

In each of these imaginary spheres the phase of the electric

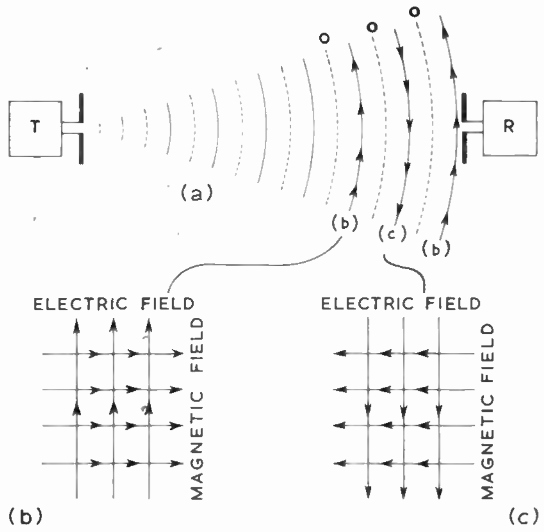


Fig. 12.1. The electric field of force radiated from a transmitting aerial. In (a) the spherical waves emanating from the transmitter T are shown arriving at a receiver R. The patterns shown in (b) and (c) represent the electric and magnetic fields (which are mutually at right angles) in a small area of the spherical wave-front arriving at R. The reversal of sign corresponds to the reversal of the direction of current in the transmitting aerial. If the patterns are taken to represent peak values of electric (and magnetic) force there is a time-lapse of one half-cycle between (b) and (c). It should be noted that (b) and (c) are in a plane at right angles to the plane of diagram (a).

(or magnetic) force is uniform even though the wave-front may extend several miles up into the sky, but of course this does not mean that the intensity is uniform. The intensity, or *field strength*, at any part of the wave-front can be diminished by absorption, or *attenuation*.

Ideally, where the wave is propagated in " free space," i.e. where it is unaffected by the proximity of the earth and the various influences of its atmosphere, it would be a simple matter to calculate the field strength at a remote point since in these circumstances the intensity of the electric field is inversely proportional to the distance from the transmitter (and the power or energy density is inversely proportional to the square of the distance).

Near the earth's surface, however, the situation is very complex, and it may be necessary to take into account a large number of factors such as the frequency, the time of day, the season, the location of the station and the nature of the terrain. In fact there are so many variables, some of them unknown or incapable of measurement, that the prediction of the received field strength often becomes a matter of estimation based on statistics.

Polarization

The waves are said to be *polarized* in the direction of (or parallel to) the electric lines of force. With wire or rod dipoles the polarization is parallel to the length of the aerial and, since for convenience all dipole aerials are either vertical or horizontal, the radiation is in all practical cases either vertically or horizontally polarized.

When waves from such aerials propagate in *free space* their sense of polarization remains constant and they will appear either vertically or horizontally polarized to the receiving aerial, which must then be orientated to the same polarization in order to receive maximum signal. In practice, free space conditions are not often achieved, and the mechanism of propagation of the wave depends upon reflections and bending at intermediate points between the transmitting and receiving aerials. This is particularly so at v.h.f. under abnormal weather conditions, and at h.f. when reflection from the ionosphere is involved. In such cases, the resultant wave arriving at the receiving aerial is in fact the sum of a number of different linearly polarized waves which arrive with varying amplitudes and relative phases. Under such conditions the resultant wave may be linearly polarized but in a sense somewhere between vertical and horizontal, or more often *elliptically polarized*, that is to say the magnitude and phase of the resultant electric field vector is constantly changing such that, to the receiving aerial, the resultant vector appears to trace out an ellipse. In terms of the direction of polarization only, it can also be said that to the receiving aerial the polarization of the incoming wavefront is linear but has constantly rotating sense.

A special case of elliptical polarization occurs when the resultant wave is made up from two waves of equal magnitude which are constantly 90° out of phase. In this case the magnitude of the resultant electric field remains constant but the polarization rotates at a constant speed equal to the speed of propagation of the waves: such a wave is then said to be *circularly polarized*. It can be artificially generated by such aerials as the helix and axial-fire crossed Yagis (see Chapter 14—*V.H.F./U.H.F. Aerials*), and should be received in such cases by a circularly polarized aerial of the appropriate sense of rotation. The use of a circularly polarized aerial for general transmission and reception at h.f. is to be recommended where possible, since such an aerial offers the best

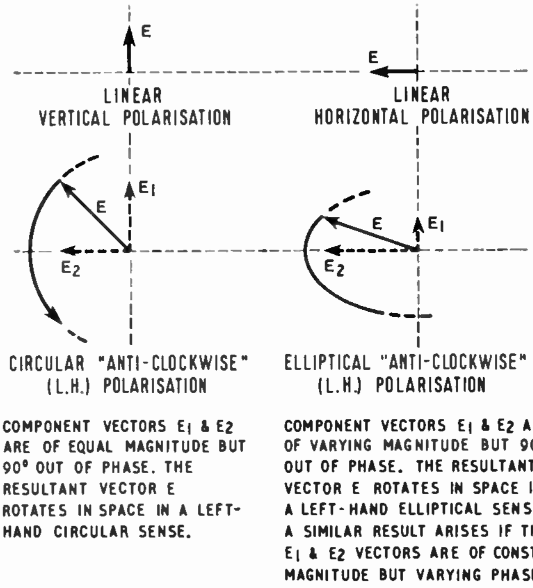


Fig. 12.2. Spatial diagrams of different senses of polarization of the electric field. The plane of the approaching wavefront is in the page.

compromise solution to the randomly varying elliptical polarization experienced by waves undergoing ionospheric propagation and reflection. This arrangement is sometimes known as " polarization diversity " and is described in detail in Reference [2].

The different modes of polarization are illustrated in **Fig. 12.2** which is drawn in terms of the direction in space of the electric field vector in the wavefront which is approaching " out of the page," i.e. as seen by the receiving aerial.

The mechanism by which waves passing through the ionosphere are split up into different components to give the result of random polarization on the downcoming wave is a function of the interaction between the earth's magnetic field and the ionosphere. A detailed explanation is beyond the scope of this *Handbook*, but readers requiring more information on the subject are recommended to study the behaviour of *ordinary* and *extraordinary* waves in Reference [1].

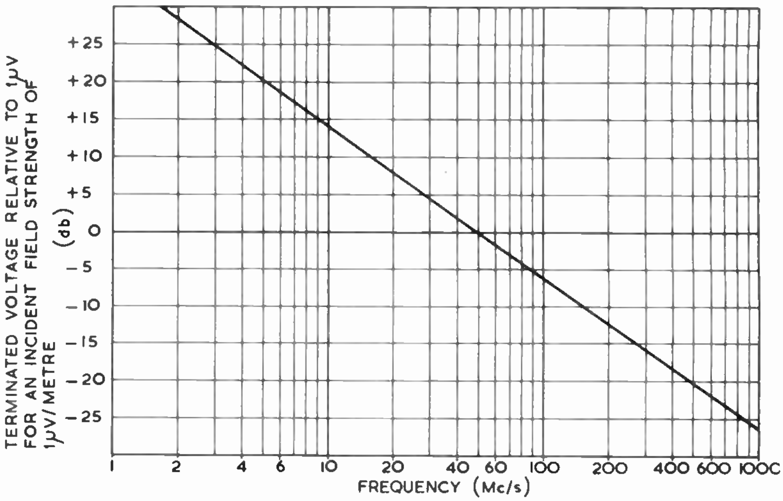


Fig. 12.3. The relationship for varying frequencies between the terminated voltage at the receiver and the incident field strength on a halfwave dipole fed with a perfect feeder.

Field Strength

The intensity of a wave is generally stated in terms of *volts per metre* of space. This will be the strength of the electric lines of force at the receiving aerial (i.e. at right angles to the direction of propagation). Field strengths normally range from low values of one microvolt per metre to high values of 10-100 millivolts per metre.

At first sight it would appear that the greater the length of the receiving aerial the greater the current induced by the field. This is true only up to the point when the wire is approaching a half-wavelength in electrical length. The reason is that the induced current-flow in the aerial moves at a speed rather less than that of the wave in space. Therefore the " distance " covered along the wire cannot be more than a half-wavelength before the electric field in the surrounding space has reversed and caused the current-flow to reverse. Thus in a wire several wavelengths long the current will, in effect, be divided into several short paths of a half-wavelength. Since the field strength is stated in terms of volts per metre it follows that the current induced in an aerial for a given field strength will be proportional to the wavelength.

For a half-wavelength dipole aerial, the open circuit e.m.f. induced at its centre is given by the expression:

$$e = E \times \frac{\lambda}{\pi} \qquad \ldots \text{(i)}$$

where e = e.m.f. at centre of dipole in volts
E = field strength in volts per metre
λ = wavelength in metres.

If this aerial is connected by a perfect feeder to a receiver which correctly terminates the feeder, the " terminated " voltage available at the input to the receiver will be $e/2$, appearing across the input resistance R of the receiver (typically 75 ohms for a coaxial line input, 300 ohms for a balanced input). To the expression (i) may be added any increase in signal obtained by using an aerial exhibiting some gain over the standard half-wavelength dipole; any losses in the practical feeder used must be subtracted.

It is customary in practice to define field strength in terms of a level in decibels relative to a field strength of 1 microvolt per metre. A field strength of 1 millivolt per metre would then be defined as +60db. Using such a notation, it is a simple matter to calculate the voltage at the receiver termination in terms of the received field and the receiving aerial parameters. **Fig. 12.3** gives the relationship of expression (i) plotted for different frequencies, as ratios in db, between the field strength and the " terminated " voltage at the receiver. Using this graph, it is possible to assess the overall performance of a receiving aerial system very quickly. If the ratio from Fig. 12.3 is X db, the aerial gain relative to a half-wavelength dipole Y db, the feeder loss Z db, and the field strength S db, then the actual voltage at the receiver terminals is given by:

$$V(db) = S + X + Y - Z \qquad \ldots \text{(ii)}$$

Example. At 70 Mc/s the field strength is stated to be 100 microvolts/metre. A three element Yagi aerial is employed with a gain of 5db over a half-wave dipole; the 100 ft. of coaxial feeder has a loss of 2db. What is the input voltage at the receiver, measured across 75 ohms?

S = +40db relative to 1 μV/m.
X = −3·5db
Y = +5·0db
Z = 2·0db

Then V = 40 − 3·5 + 5·0 − 2·0 = 39·5db relative to 1 μV. The receiver input voltage is therefore 94 microvolts, measured across the 75 ohm input to the receiver.

Height Gain

On high frequencies particularly, in addition to the waves received by the direct path between transmitter and receiver, there will also be incoming waves which have been reflected from the ground or other relatively large objects. Since these

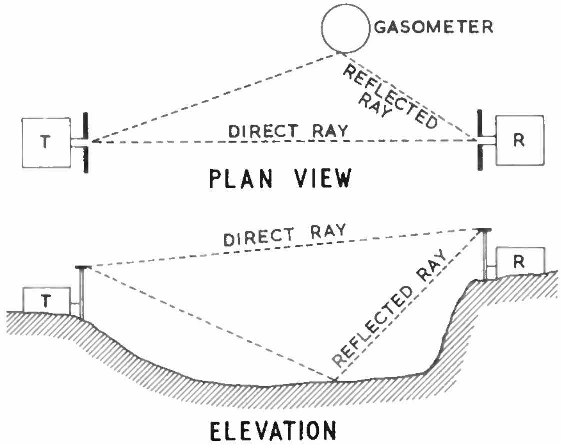


Fig. 12.4. Interference between direct and indirect (reflected) rays. The strength of the received signal depends on the phase difference between the two rays when they arrive at the receiver and on their respective intensities.

reflected waves travel over a rather longer distance, their time of arrival is slightly later than that of the direct ray. Thus they produce interference effects with the direct waves and cause the received signal to increase or decrease according to local phase relationships: see Fig. 12.4.

The interference pattern produced between the direct ray and the ground reflected ray of Fig. 12.4, when the height of one of the aerials is varied, gives rise to the phenomenon

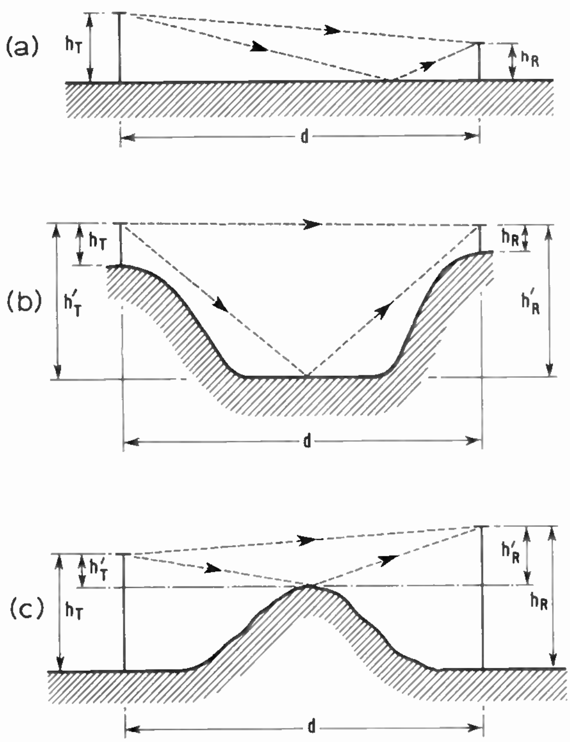


Fig. 12.5. The effect of ground profile on direct and indirect rays. (a) Linear height gain. (b) Marginal height gain. (c) Rapid height gain. (The Vertical scale has been exaggerated in each case.)

of height gain. It explains in precise terms exactly why the signal received by an aerial by a nominally direct path, i.e. with no ionospheric reflection, increases when the aerial height above ground is increased. The explanation does not hold for waves propagated through the ionosphere, or by anomalous tropospheric means, and is therefore applicable only to "normal" propagation at frequencies above 30 Mc/s, i.e. in the v.h.f. bands.

The three basic arrangements for receiving and transmitting aerials, and the intervening ground, are illustrated in **Fig. 12.5**. The classic plane-earth case is shown in **Fig. 12.5(a)**, and under such conditions the signal received at the distant aerial is given by the relationship:

$$e = constant \times \frac{h_T\,h_R}{\lambda d^2} \qquad \ldots \text{(iii)}$$

where h_T = height of transmitting aerial
h_R = height of receiving aerial
λ = wavelength
d = distance between aerials.

In this expression, the derivation of which can be found in Reference [1], h_T, h_R and d must all be in the same units, and d must be much larger than either h_T or h_R (by a factor of at least 10): this is usually the case in practice.

From expression (iii) it is clear that an increase in either h_T or h_R will result in a corresponding increase in e, and doubling the height will give an increase of 6db. This is the 6db "height gain" rule. It is worthwhile noting that for each 6db improvement the height must be doubled, so there soon arises in practical aerial systems a physical limit beyond which the added complexity of raising the aerial does not pay sufficient dividends to make it attractive.

The "classic" case of Fig. 12.5(a) may be considered that of the average v.h.f. station. In the same terms the case of **Fig. 12.5(b)** may be considered that of the "portable" v.h.f. operator who has selected a good site. Here, the aerial height above immediate ground is relatively small compared with the effective height above the ground level at the point where the indirect ray is reflected from the intervening ground. Expression (ii) now becomes:

$$e = \text{constant} \times \frac{h'_T \cdot h'_R}{d^2} \qquad \ldots \text{(iv)}$$

where h'_T and h'_R are the *effective* heights of the two aerials. There is still height gain to be achieved by increasing aerial height locally, but not at the same rate as in the first example. To obtain a gain of 6db it is necessary to double h'_R, and this will require a many-fold increase in h_R. In the limit it clearly becomes not worthwhile seeking any great aerial height: this is often the case for portable stations on hilltops, when the increased loss in the feeders is less than offset by the small additional signal to be obtained by raising the aerial.

In the third case, that of the poorly sited station whose aerial is just able to see over the surrounding higher ground, the reverse of case (b) applies. The effective height h'_R is much *less* than h_R, and a small increase in the height of the aerial is required to bring massive improvements in signal level.

In most cases, stations are situated in locations which provide examples of each condition in different directions, and the individual operator must make his own assessment of the likely all-round improvement to be obtained by increasing the height of this aerial system.

It is also possible under certain topographical conditions to observe the phenomenon of "inverse height gain." This occurs usually in areas of rugged and steep slopes, where the indirect ray is reflected from an inclined ground plane, and arrives in such a phase as to cancel the direct ray at certain receiving aerial heights. In such cases an increase of height may move the aerial into such a region, with a resulting fall in signal level. This is usually found to occur over a limited range of aerial height in any one location, and can be avoided by careful selection of the optimum aerial height by experiment.

MODES OF PROPAGATION

The mechanism by which radio waves are propagated from point A to point B on the earth's surface varies according to the frequency employed. Although the divisions are not precise, it is possible broadly to divide the spectrum into three parts.

Frequencies below 2 Mc/s tend to propagate as surface waves over the earth and are said to be subject to *ground wave* propagation. In the range 1-70 Mc/s, long distance propagation is entirely by means of reflections in the ionized layers high above and surrounding the earth: this is termed *ionospheric* propagation. For frequencies above 50 Mc/s the effects of the ionosphere are less uniform and regular and both normal and anomalous propagation is governed largely by the meterological conditions in the troposphere. such propagation is therefore termed *tropospheric*.

In addition to the normal propagational mechanisms, it is also possible to exploit certain specialized techniques for long range communication. These are dealt with later in this chapter.

Ground-wave Propagation below 2 Mc/s

The lowest frequency band used by amateurs is at the upper limit where useful and consistent distances may be covered by the ground wave, both by day and by night.

In general a receiving aerial will respond to radiated fields originating as waves arriving from different sources. It can receive waves which descend from the ionosphere after reflection, waves which descend from the troposphere by refraction and waves which arrive directly from the transmitting aerial. The first two classes are unaffected by the presence of the earth (except as discussed later), and it is the third group of nominally direct waves which are subject to modification by the earth's presence, and are normally known as the *ground wave*.

The first part is the field resulting from the sum of the wave arriving directly at the receiving aerial, and that arriving after reflection at the ground at an intermediate point. This part is generally called the *space wave* (Fig. 12.6). The other part of the ground wave is in fact a *surface wave* which is guided along the boundary formed by the earth's surface in much the same way that a transmission line guides an

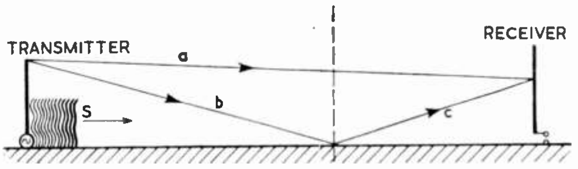


Fig. 12.6. Space waves and surface wave, together making up the ground wave.

S = surface wave (wavefront is normal to plane of paper).
a = direct component; b = incident component; c = reflected component — a, b, c: space wave

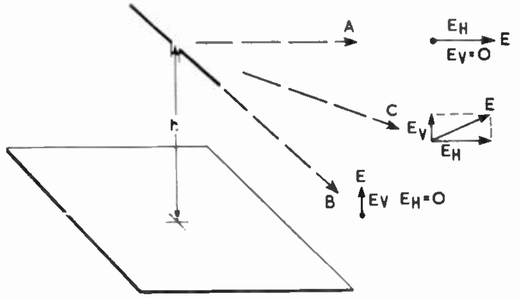


Fig. 12.7. Variation of polarization of the ground wave field with angle of azimuth around a horizontally polarized dipole.
h = height of horizontal dipole above reflecting plane
In direction A, field is entirely horizontally polarized
In direction B, field is entirely vertically polarized.
In direction C, field is elliptically polarized.

electromagnetic wave which propagates along it. As this wavefront passes over the earth's surface, energy is extracted from it to supply losses in the ground and it, therefore, suffers attenuation which is a direct function of the physical constants of the ground over which it is propagated. For aerials which are electrically close to, or, in the case of verticals, at the earth's surface, the path geometry is such that the direct and indirect components of the space wave virtually cancel out at the receiving aerial, and the major contribution to the received field is from the surface wave. The cancellation of the space wave arises from the 180° phase shift which the indirect component receives at the point of ground reflection—this is considered in more detail later.

The expressions for the value of the space and surface waves are very complex, even in the simplified form derived by Norton from the original Summerfeld equations. A detailed explanation is given in Section 16.03 of Reference [1] but it will suffice here to say that the surface wave expression differs for the case of vertically and horizontally polarized aerials. In the case of vertically polarized aerials, the surface wave is entirely vertically polarized in all directions of azimuth, and is reduced by an attenuation factor F, which is a function of distance, and of the ground conductivity, and generally increases with decreasing ground conductivity. At near-in points, F varies exponentially with distance but at greater distances varies inversely as the square of the distance. In the case of nominally horizontally polarized aerials, close to the earth's surface, the surface wave varies in polarization with the direction in azimuth, being horizontally polarized in the directions at right-angles to the aerial, elliptically polarized at intermediate directions, and completely vertically polarized off the ends of the aerial. The actual magnitude of the wave will, of course, vary with the horizontal radiation pattern of the aerial, and in general will be a maximum in the direction in which it is horizontally polarized, and decrease to a low value in the direction in which it is entirely vertically polarized.

These variations of polarization arise from the fact that the surface wave from the horizontal aerial is made up of two components, one vertically polarized and one horizontally polarized, and varying in value differentially with the angle of azimuth, the one being zero at right-angles to the aerial, and the other zero in the line of the aerial (Fig. 12.7).

The vertically polarized term is subject to the same attenuation factor F as before, but the horizontally polarized

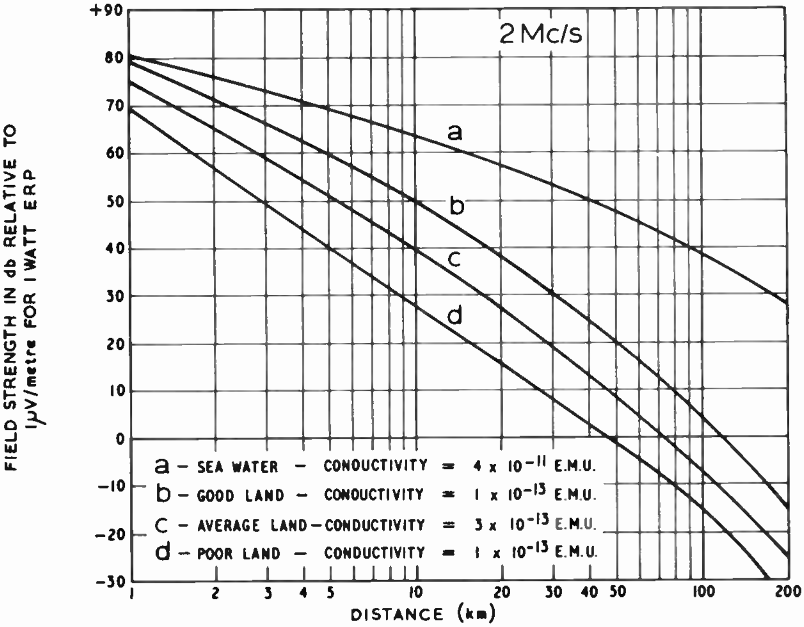


Fig. 12.8. Variation of field strength with distance, and with ground conductivity, for 2 Mc/s vertically polarized transmission. The curves refer to land of poor, moderate, and good conductivity, and also to propagation over the sea. The field strength is that laid down by a perfect short vertical aerial radiating a power of 1 watt.

term is subject to an attenuation factor G which is related to F by the expression:

$$G = \frac{F}{(\varepsilon_r + jx)^2} \qquad \ldots (v)$$

where εr = permittivity of earth in e.s.u.

$x = \frac{18 \times 10^3 \sigma}{f}$

σ = conductivity of earth in e.m.u.

f = frequency in Mc/s.

Thus for lower frequencies, or for greater conductivities, G will be very much smaller than F, with the result that the horizontally polarized surface wave will attenuate much more rapidly with distance than the vertically polarized one. This is generally the case at frequencies below 2 Mc/s and normal ground conductivities. On frequencies over 2 Mc/s the ground-wave attenuation becomes greater and ionospheric reflection becomes more important.

The effectiveness of ground-wave propagation depends largely on the conductivity and dielectric constant of the soil over which the transmission path extends, being at its best over sea water, and at its worst over old rock layers. For a given path and radiated power the signal will be stronger at lower frequencies.

Field strengths for ground-wave propagation have been measured over a wide range of conditions, and within certain limits it is possible to predict performance reasonably accurately. Thus **Fig. 12.9** shows the effect of the attenuation over average soil of 1 Mc/s and 2 Mc/s waves, vertically and horizontally polarized, compared with propagation in free space. At night particularly, extended range will be obtained on these frequencies by ionospheric reflection and considerably higher field strengths obtained. The variation of field strength with distance is also illustrated in **Fig. 12.8** which is drawn for propagation at 2 Mc/s over land of various conductivities and over the sea. The field strengths shown apply to a perfect short vertical aerial radiating a power of 1 watt over a perfect ground return and are thus applicable only to vertical polarization. The corresponding field strengths for horizontally polarized signals will be proportionately lower as shown in Fig. 12.9. In practice signals from 160 metre amateur stations may be as much as 10db lower than those quoted owing to the inefficiency of the aerial and ground systems employed.

Ground wave propagation is a very complicated subject, and even the foregoing explanations only skim the surface. A more detailed explanation of ground wave propagation can be found in References [1] and [8].

Ground Effects above 2 Mc/s

Although the propagation of ground waves, or more accurately the ground surface-wave, at frequencies much above 2 Mc/s becomes of decreasing importance due to the disproportionate increases in attenuation with frequency, the ground still plays a part in communication by means of

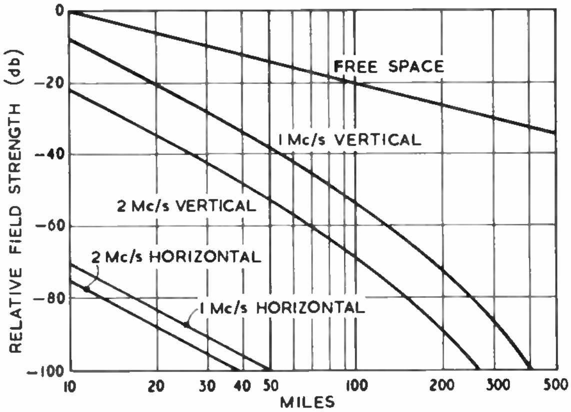


Fig. 12.9. Attenuation curves for 1 Mc/s and 2 Mc/s ground waves, relative to the "free space" field strength at a distance of 10 miles from the transmitter.

ionospheric reflections. The radiation launched upwards towards the ionosphere at the required angle to achieve the intended distance is made up of the direct and ground reflected rays leaving the aerial system which must be erected over the earth. The way in which these cancel and reinforce according to the angles involved to give rise to the vertical radiation pattern of the aerial is covered in Chapter 13—*H.F. Aerials*.

The assumption is usually made that the ground directly in front of the aerial for several wavelengths is a perfect reflector and that the ray is reflected from its surface without attenuation or random phase shift. This is not the case in practice and the indirect ray is affected by the physical constants of the ground to some degree depending upon the polarization. Rays which are horizontally polarized are reflected with some small degree of attenuation and with approximately 180° phase shift, irrespective of the angle of incidence upon the ground, but rays which are vertically polarized suffer considerable attenuation and rapid variations of phase shift between 5° and 180°, at certain angles of incidence within a critical range. These are low angles approaching grazing incidence and consequently the very angles at which maximum radiation is required for long range ionospheric propagation. The phenomenon of polarization discrimination described is known as the *pseudo-Brewster effect*, and a full treatment may be found in Reference [8].

Ionospheric Propagation: 1-70 Mc/s

The frequency range of 1-70 Mc/s covers the lower and upper limits where ionospheric reflection is the controlling factor in the propagation of waves over distances which may encircle the globe.

Propagation of radio waves is mainly the result of single or multiple reflections from ionized regions in the upper atmosphere. These ionized regions are generally found at heights of 100-400 km. (60-240 miles) and are known collectively as the *ionosphere* (**Fig. 12.10**). At these heights the rarified air is partially ionized by the sun's radiations, i.e. some of the molecules are converted into ions and free electrons, and after sunset there is a gradual recombination of the ions with the electrons.

The degree of ionization is not constant with height and there are " layers " of more intense ionization. The actual heights of these layers vary from day to night and with the seasons, and to a lesser degree continuously. There are, however, three regions of chief importance. These are the *E*-layer at about 120 km., the *F1*-layer at 200 km. and the *F2*-layer at 300-400 km. At night and in mid-winter the *F1* and *F2* layers combine to form a single layer at 250 km. Below the *E*-layer there is a *D*-layer or region (at 50-90 km.) which generally is more important as an absorber than as a reflector of radio waves since the attenuation at this altitude is somewhat greater. The approximate heights of the various layers are shown in Fig. 12.10.

The theory of ionospheric reflection is a complex subject and will not be dealt with here. In brief, a wave can be said to undergo reflection when on reaching a suitably ionized region it sets in motion the free electrons which then move in a similar manner to the electrons in the transmitting aerial and " re-radiate " (i.e. reflect) the wave in a changed direction.

As the frequency is increased, a greater degree of ionization is necessary to cause reflection. The *F2*-layer normally has the greatest degree of ionization and therefore reflects the highest frequencies. Patches of intense ionization in the form of clouds can, however, occur at *E*-layer height, especially in summer, and on certain occasions they will reflect radiation at frequencies well above the limit for normal *F2*-layer propagation. The phenomenon is widely experienced on 28 Mc/s, and is not unknown at 145 Mc/s. By reason of their nature these local layers of intense ionization are termed *sporadic-E*.

Sunspots

The solar radiation responsible for ionizing the atmosphere is continuously varying. A relationship has been found to exist between the number of sunspots appearing at any one time and the degree of ionization. This sunspot activity

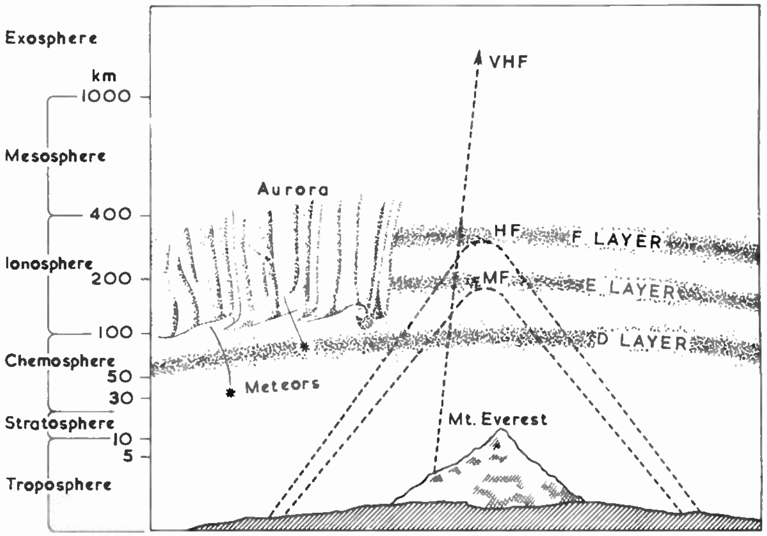


Fig. 12.10. The earth is surrounded at various heights by so-called layers of ionized gas, and the region in which these exist is known as the ionosphere. Other regions of the atmosphere are given various names according to their dominant properties and characteristics. The vertical scale is shown graded in such a way as to illustrate more clearly the detail at the lower altitudes.

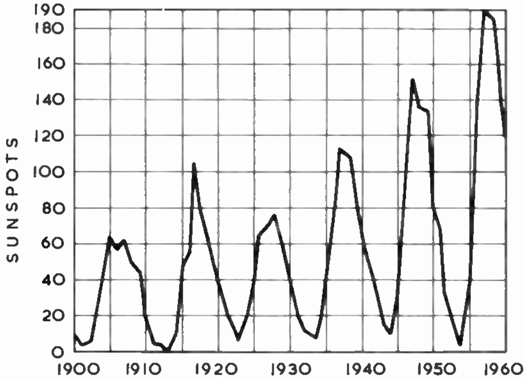


Fig. 12.11. Sunspot activity from 1900 to 1960.

occurs cyclically in a complex manner. There is a basic cycle reaching peaks of maximum activity once every 9-13 years, but these peaks (and the intervening minima) are themselves subject to considerable variation from cycle to cycle, typical values of maxima and minima being shown in Fig. 12.11. These variations in sunspot activity are faithfully followed by similar variations in the density of atmospheric ionization.

A shorter term variation in ionospheric conditions is related to the period of rotation of the sun about its own axis, being approximately 27 days. The effect is most pronounced during the appearance of a very large spot on the sun's surface, and gives rise to the ionospheric storms described below, which have a distinct tendency to follow the 27 day cycle. The seasonal effect due to the earth's movement round the sun and the tilt of its axis is also a major factor in that it causes considerable variations in the layers to occur in different parts of the world at a given time.

In addition to the ionizing radiations, the sun occasionally emits particles which arise from vast eruptions on its surface. On arrival at the earth, particularly in the polar regions, these particles tend to obliterate the ionized layers, the effect being most marked in the *F*-region. Thus communication may be interrupted for periods of a day or two over certain paths. This condition is known as an *ionospheric* (or *magnetic*) *storm*. Auroral displays are prominent at these times, and radio waves in the frequency range 10-200 Mc/s will be reflected back from the auroral zone modulated with a characteristic flutter which precludes all but c.w. as a means of communication. The degree of reflection is greater at the lower frequencies, being relatively common at 28 Mc/s, and somewhat infrequent at 145 Mc/s even under sunspot maximum conditions. It is interesting to note that the existence of a visible auroral display by no means signifies the presence of auroral reflection of radio waves. Although excited by the same solar condition, they tend to occur at different and not necessarily related times. A full analysis of auroral propagation is contained in References [5], [6] which describe the mechanism in detail, together with results of a long term experimental survey commenced in 1957 by the RSGB as part of its programme of contributions to the International Geophysical Year. Further reference is made later in this chapter under the section dealing with modes of "scatter."

Ionospheric storms are often preceded (by approximately two days) by what are called *sudden ionospheric disturbances* (*s.i.d.*). Under these conditions "squirts" of abnormally strong solar ultra-violet radiation cause intense ionization in the low *D*-layer and waves are absorbed before reaching the *F*-region or are reflected at short distances. Thus long-distance communication is subject to fade-outs of an hour or two.

In addition to the modes of propagation described in the foregoing paragraphs, there are certain second-order effects. These are *ionospheric scatter* and *meteor-trail scatter*, and a brief account of them is given later in this chapter.

Maximum Usable Frequency

The maximum frequency which is reflected by the ionosphere over any particular path is known as the *maximum usable frequency* (*m.u.f.*). It has been found to depend on

(*a*) time of day,
(*b*) season,
(*c*) latitude,
(*d*) period of sunspot cycle.

The actual value of m.u.f. will vary according to the layer which is responsible for the reflection from the ionosphere. For each layer, the highest m.u.f. is obtained when the ray path leaves the earth tangentially so that the ray approaches the appropriate layer at as oblique an angle as possible. This corresponds typically to an overall ground-to-ground distance of about 4000 km. (2500 miles) for *F2*-layer propagation; or

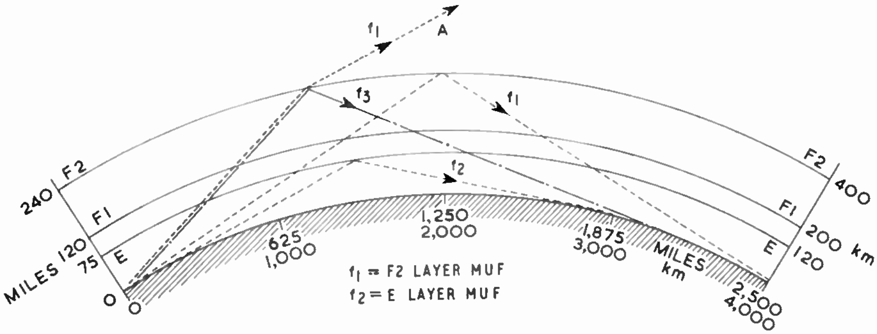


Fig. 12.12. Paths of waves leaving tangentially to the earth's surface producing oblique or grazing incidence at the reflecting layers. Such waves reach the most distant points possible and their frequency is the maximum usable frequency (m.u.f.) for this distance. For radiation at steeper angles to the earth's surface, the m.u.f. is somewhat lower.

2500 km. (1500 miles) for *E*-layer. At this highest m.u.f. any rays leaving the earth at a higher angle of elevation will penetrate the layer and not be reflected. To use such ray angles, with consequently shorter path, it is necessary to reduce the operating frequency. This is illustrated in Fig. 12.12.

Since the m.u.f. for each layer is different, that for the *E*-layer being typically lower than for the *F2*-layer, it is possible to propagate two paths to a distant receiver by using a frequency just below that of the *E*-layer m.u.f. Due to the transient nature of the ionosphere, the path lengths of the two modes will continually be varying as the layer heights change and the varying phase difference between the two signals arriving at the receiver will give rise to deep fading. This is illustrated in Fig. 12.17. The restricted range of amateur band frequencies makes the choice of a suitable frequency difficult, and the only control that can be exercised is over the projection angle (Fig. 12.13) of the aerial—a function mainly of its height above ground. This can be adjusted to favour one mode and discriminate against the other to reduce the fading at the far end. In general the strongest signals, i.e. least attenuation, will occur using frequencies just below the m.u.f. for the particular path distance and layer involved.

When a wave is sent vertically upwards (i.e. angle of radiation 90° or " normal incidence ") the highest frequency for which reflection by any particular layer will occur is termed the *critical frequency*, f_0. This frequency is much lower than the m.u.f. for oblique incidence, being related approximately by *m.u.f.* $= f_0/\cos\phi$, where ϕ is the angle of incidence of the ray to the layer. At frequencies higher than f_0 the radiation will penetrate the layer and be lost in outer space and will continue to do so as the angle of radiation is progressively lowered until an angle is reached at which reflection begins to occur. This limiting angle is called the *critical wave angle*: see Fig. 12.14. Signals will then be received at a distant point *R2*. Radiation at lower angles will obviously be reflected to greater distances such as *R3* provided that it is not cut off by reflection from the upper surface of the intermediate layer as it descends towards *R3*. In such cases it may still be possible to reach *R3* by using two-hop propagation at a lower frequency, and with angle of incidence such that the frequency used is below the m.u.f. for the upper layer, but sufficiently high to pass through the lower layer at the steeper angle. This point is discussed later in the chapter and illustrated in Fig. 12.15. At points nearer to the transmitter such as *R1* no signals will be received by ionospheric reflection, but when *R1* is sufficiently close to come within the range of the ground wave the signals will again be heard. In between there is an area of no reception and this is known as the *skip zone*. The distance from the transmitter to the nearest point at which the radiation is reflected back to earth is known as the *skip distance*. Since the density of ionization of the layer will constantly change and its height fluctuate, so will the skip distance change for any particular frequency. Clearly there will be no skip at all when the frequency in use is less than the critical frequency f_0 for the layer in question. This situation is often found on the 3·5 and 7·0 Mc/s bands.

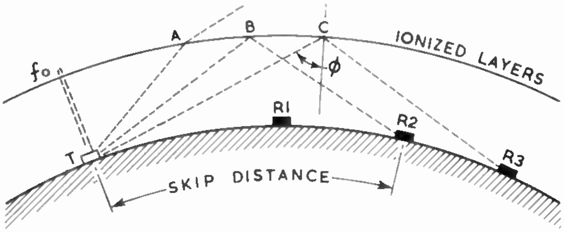


Fig. 12.14. The highest frequency at which radiation directed vertically upwards from a transmitter T is reflected back to earth by an ionized layer is termed the critical frequency (f_0). At higher frequencies the radiation penetrates the layer, and for such higher frequencies reflection can occur only if the angle of radiation is lowered to a sufficiently small angle: see A and B. The nearest point at which the radiation can be reflected back to earth (R2) is therefore at some distance from the transmitter. This is known as the skip distance. Inside this distance (e.g. at R1) no signals will be received from T by reflection. The limiting angle for reflection depends on the frequency, being steeper as the frequency is lowered towards the limiting value f_0. The skip is correspondingly shortened

The critical wave angle for any particular ionospheric layer depends upon the frequency of the radiation and becomes smaller as the frequency increases. Therefore the skip distance increases as the frequency increases, and the m.u.f. is the limit which must not be exceeded if the receiver is to remain in the area of reception just outside the skip zone. In other words, the more closely the operating frequency approaches the m.u.f. the more nearly does the skip distance extend towards the receiver. The absorption attenuation of the reflecting layer also decreases markedly as the operating frequency approaches the m.u.f., so that for any particular circuit between transmitter and receiver, the optimum working frequency lies just below the m.u.f. for that particular circuit. Any rise in operating frequency or fall in

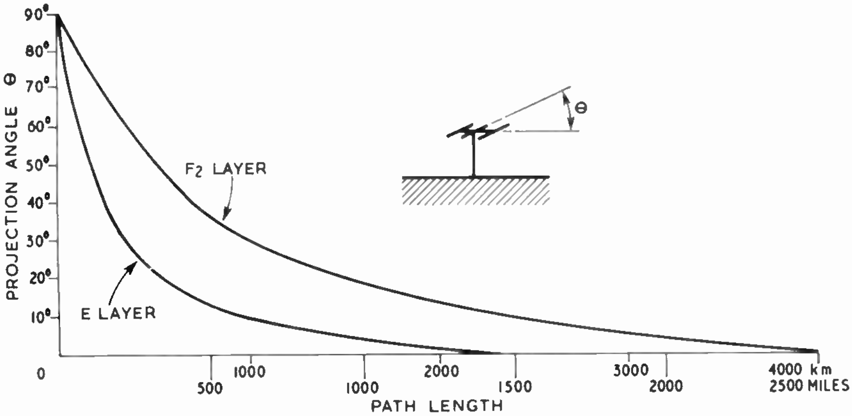


Fig. 12.13. Graph of projection angle versus ground distance for *E* and *F2* layers.

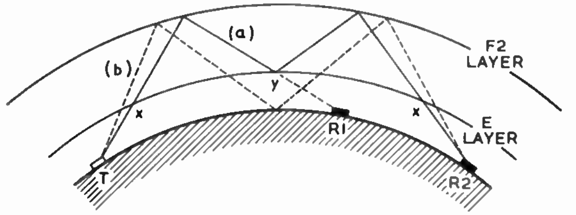


Fig. 12.15. Topside reflection from the *E*-layer. In case (b), the frequency used is above the *E*-layer m.u.f. at all times. In case (a), the frequency used is above the *E*-layer m.u.f. at points X but below it and hence reflected at point Y.

m.u.f. will result in a sudden drop-out of received signals as the skip zone extends to include the reception point.

By reflection from the ground a wave may be returned to the layer two or more times and thus arrive at a more distant receiver by " multi-hop " propagation. World-wide distances may be covered in this way under suitable ionospheric conditions.

The critical frequencies for the *E*, *F1* and *F2* layers are measured continuously at many scientific stations in various parts of the world, notably the Radio and Space Research Station at Slough, England, and the Central Radio Propagation Laboratory in Boulder, Colorado, USA. Over a period of years it has been possible from these measurements to build up world maps of the distribution of critical frequency at various seasons and stages of the sunspot cycle. This is of value because there is generally a simple but approximate relationship between the critical frequency f_0 and the m.u.f. for a particular layer. Thus for the *F2*-layer the m.u.f. for a distance of 4000 km (2500 miles) is approximately three times the *F2* critical frequency, and for the *E*-layer the m.u.f. for a distance of 2000 km (1250 miles) is approximately five times the *E*-layer critical frequency.

From these factors the f_0-distribution maps can be converted to show the m.u.f. distribution. One such typical map is shown in **Fig. 12.16**. These maps form a very useful basis for forecasting, under normal conditions, the optimum operating frequency for any particular path.

It is important to understand that the contours marked on the map refer to the state of the ionosphere overhead at each point. It is permissible for both receiver and transmitter to lie outside the region bounded, for example by the 28 Mc/s contour, and for communication on that band to occur between them, provided that the point of reflection in the ionospheric *F2*-layer lies within the region bounded by the contour. For long paths, the circuit should be divided up into a number of equal length hops, each corresponding to the maximum for *F2*-layer propagation (approximately 2500 miles). This involves a certain simplification of the problem since the effective height of the layer varies with latitude being significantly greater in the equatorial regions.

In the example given in Fig. 12.16, it would be possible for a station located on latitude 10 South to communicate on

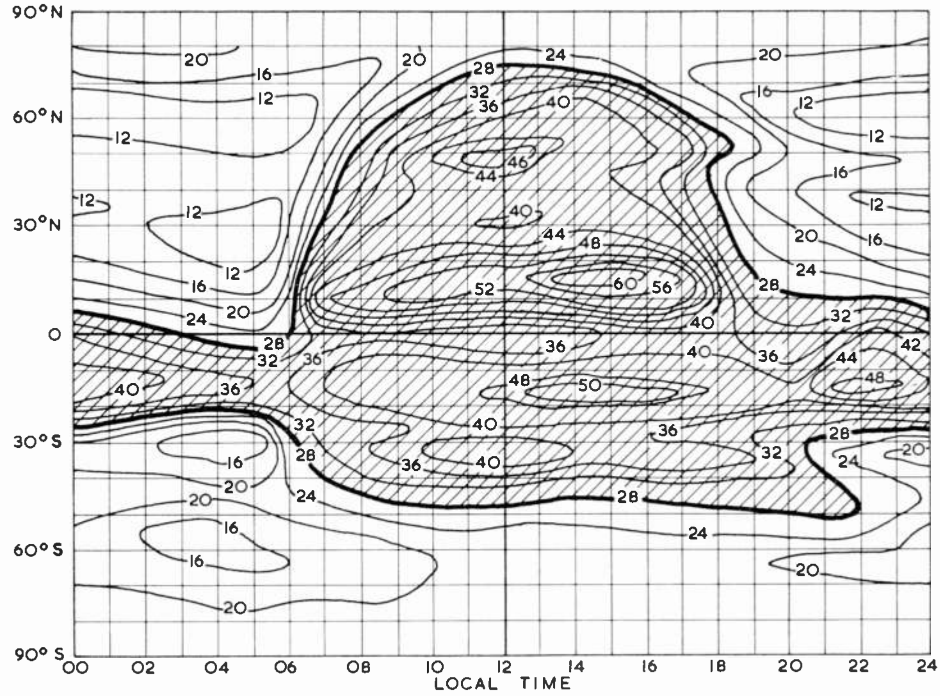


Fig. 12.16. A typical distribution map showing the m.u.f. for transmission from a given site over distances of 4000 km. (2500 miles) using reflection from the *F2*-layer varies during a 24-hour period (local time). The map changes throughout the year (the example shown here being the distribution in November 1947), but it can be applied to transmitter sites on any latitude. The region within which *F2*-layer reflections at 28 Mc/s are possible is shaded.

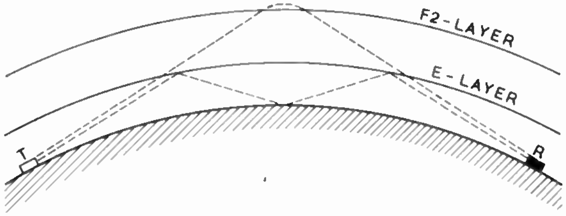


Fig. 12.17. A relatively low-frequency wave, which is reflected from a low-altitude layer (such as the E-layer) may reach a distant receiver by successive reflection between the earth and the layer. Attenuation of the wave is relatively greater since reflection occurs at a low altitude. A higher-frequency wave is reflected at a higher altitude and covers the distance in one hop. If the same frequency will propagate by reflection from both layers, fading will occur at the distant receiver.

28 Mc/s with any station on the same latitude at any time of night or day provided the distance involved was near that of maximum *F2*-layer propagation (or multiples thereof). With the very high m.u.f. occurring in certain regions along this latitude, the skip distances will be relatively short on 28 Mc/s and the receiving station can lie within quite wide limits about each point of nominal *F2*-layer return. For day time transequatorial propagation, as for more temperate latitudes, the scope is less and the distances more critical, while for transequatorial propagation at night, the limit will be single hop propagation, with a fair chance of deep fading as the skip changes.

As the frequency is reduced the reflection tends to occur in the lower ionospheric layers, and, particularly in the *D*-layer, the density of the atmosphere begins to have an appreciable effect in the absorption of the energy in the wave. The increased absorption is due to the higher rate of collision between the electrons which are set in motion by the wave and the air molecules. The effect is greater as the frequency is lowered, and the limit for any particular path is reached at what is termed the *lowest usable high frequency* (*l.u.h.f.*).

The m.u.f. is defined quite precisely by the limiting frequency at which the radiation will reflect from, rather than penetrate, the appropriate layer, and is therefore independent of the power involved or the aerial type. If the radiation is not returned to earth, no amount of increase in power or aerial gain, or receiver sensitivity can make it receivable. However, the l.u.h.f. does depend very much on effective radiated power, receiver sensitivity and signal-to-noise ratio, since it is a balance between the attenuation experienced along the path and the minimum readable signal detectable at the receiver. It is therefore usually defined for an overall practical circuit taking into account the equipment parameters at each end. As explained earlier it is also possible at point *R*1 in **Fig. 12.15** to lose the downcoming wave from, say, the *F2*-layer, by reflection from the topside of the *E*-layer, if the frequency used is too low, i.e. below the m.u.f. for the lower layer, taking into account the angle of incidence of the downward wave on those layers. This occurs when the m.u.f. of the *E*-layer at point X is very much lower than that point Y, permitting the ascending radiation to pass through at X, but reflecting it at Y. It is then possible to get effective two-hop *F2*-layer propagation to point *R*2 without an intermediate ground reflection. Since only three effective passages of the *E*-layer are then involved (as against four for the normal two-hop *F2*-layer circuit) as in Fig. 12.15, the overall path attenuation is less and an enhanced signal is received at point *R*2.

When the frequency used is below the m.u.f. of the *E*-layer at all points a normal *F2*-layer single hop circuit may turn into a two hop *E*-layer circuit, and the additional attenuation due to one further reflection at the ground and in the ionosphere will result in a weaker signal at the receiver. The use of a frequency higher than the *E*-layer m.u.f. will ensure a single hop circuit with a stronger received signal. This is illustrated in Fig. 12.17.

It is clear from the foregoing that the practical l.u.h.f. may under certain conditions be determined by the ionization intensity of the intervening lower layers rather than the basic overall circuit loss. This is more likely to be the case when high powers and sensitive receivers are employed, which could normally overcome quite high absorption attenuations in the ionosphere.

Ionospheric absorption is much less at night than during the day and therefore the attenuation of the lower-frequency signals (of the order of 1-5 Mc/s) at night is very little different from that of higher frequencies in the 20 Mc/s region. Since the m.u.f. at night over a particular path will generally be less than half that of the daytime figure, this means that for night-time long-distance communication it is possible to use considerably lower frequencies and still maintain good signal strength. Broadly speaking, the m.u.f. for a particular path is higher during the winter months than it is during the summer months, but during periods of ionospheric storms the m.u.f. may become much lower for transmission in certain directions but higher in other directions.

The variation of the average daytime m.u.f. over six successive years is shown in **Fig. 12.18.** The first year represented by these records (1937) coincided with a sunspot maximum, and it will be noticed that the m.u.f. is exceptionally high throughout this period, whereas during the sunspot minimum (1941-42) the m.u.f. was consistently low.

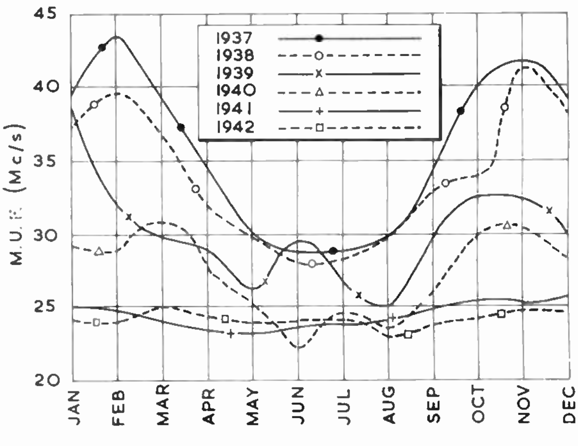


Fig. 12.18. The curves show how the average daytime m.u.f. for all directions (recorded at G6DH) varied over a period of six years (1937-42). Note the high winter peaks in the sunspot-maximum year 1937 and the small variation as between winter and summer in the sunspot-minimum period 1941-42.

In planning optimum working frequencies for any particular time, distance and direction, it is therefore necessary to take all these variations into account. Readers who may wish to go deeper into the subject can obtain the CRPL monthly *Basic Radio Propagation Predictions* on subscription from the National Bureau of Standards, Washington 25, DC, USA. Maps of the type shown in Fig. 12.16 are included in this publication.

From a knowledge of the published *F2*-4000 km and *E*-2000 km m.u.f. it is possible to make an estimate of the m.u.f. applicable to shorter path-lengths. **Figs. 12.19** and **12.20** have been prepared for the *F2* and *E* layers respectively. They are idealized to the extent that they are based upon uniform ionization intensities along the propagation path and are derived from an approximate relationship between the maximum m.u.f. and critical frequency. They are however generally applicable to one-hop transmission from the respective layers. The following examples illustrate the use of the diagrams.

Example I. If the *E*-2000 km m.u.f. is 20 Mc/s, then the 14 Mc/s skip distance is almost exactly 1000 km.

Example II. If the *F2*-4000 km m.u.f. is 20 Mc/s then either the 3·5, 7 or 14 Mc/s bands may be used to propagate over a 2000 km path. In such a case the 14 Mc/s band would be preferable, being the nearest to the path m.u.f. (approximately 17 Mc/s by interpolation) and subject to the least attenuation.

Explanations of the present technique for ionospheric forecasting and planning can be found in *Handbook 90*, also published by the CRPL and available from the National Bureau of Standards, Washington 25, DC, USA.

Over very long distances the distribution of sunlight may be such that part of the transmission path is in darkness. This will of course apply particularly to east-west routes and be of less importance on north-south routes. Over a mixed light-and-dark path the m.u.f. may be much lower than that of the daylight zone and will usually be determined by the m.u.f. for the dark region of the required path.

For a multi-hop *F2*-layer circuit, the procedure involves consideration of the *F2*-layer m.u.f. at the intermediate ionospheric reflection points, such points being crudely established by dividing the path up into a number of equal length *F2* layer single hops, not exceeding 4000 km each, and using for each the value of *F2* m.u.f. appropriate to that length of hop and for the relative position along the whole path with respect to the varying ionospheric intensity of ionization. This was explained in connection with Fig. 12.16.

In the summer months the determining factor in propagation will frequently be reflection by the *sporadic-E* layers. The cause of this intense ionization is not yet fully known but appears to be due to a combination of the usual sun's ionizing radiation and another factor local to the earth's atmosphere. The area of the reflecting " clouds " may be quite small and is seldom more than a few square miles. Directional checks have shown that these reflecting zones move at variable, and at times, quite high speeds of the order of 200-400 km. per hour. The m.u.f. for this *Es* reflection may at times go up to 80 Mc/s and is not unknown to affect the 145 Mc/s band, but as a rule it seldom exceeds 70 Mc/s in Europe. Optimum distances are generally 600-1000 miles in the 25-60 Mc/s range and multi-hop transmission has hardly ever been known to occur, presumably due to the localized nature of the E_s " clouds." It may, however, enable a key reflection to be achieved at a critical point in a predominantly *F2*-layer multi-hop circuit. The frequency of occurrence is sufficient to cause many interference problems in Europe and USA between television stations situated 800-1000 miles apart.

The formation of sporadic-*E* layers does not seem to follow a closely defined pattern, although there is a tendency

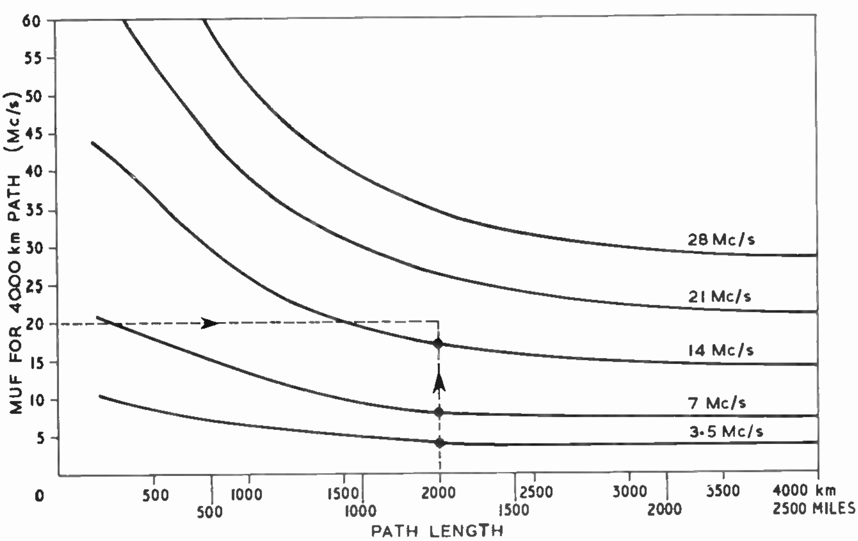


Fig. 12.19. Graph of single hop *F2*-layer propagation. For any desired distance, all bands lying below the *F2*-4000 m.u.f. line will propagate successfully. The best one will be that closest to the *F2*-4000 line.

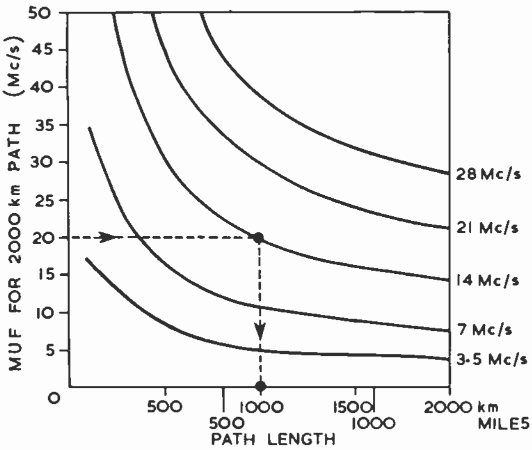


Fig. 12.20. Graph of single hop *E*-layer propagation. For any particular band, the skip distance is that at which the *E*-2000 m.u.f. line intersects the " band " curve. As illustrated, the 14 Mc/s skip will be 1000 km when the *E*-2000 m.u.f. is 20 Mc/s.

towards a 27-day recurrence. The m.u.f. tends to be higher at a sunspot maximum but not so markedly as in the *F*-region.

Fade-outs and Fading

During periods of high sunspot activity two opposing effects occur in the ionosphere. As the earth comes under the influence of an active sunspot (which may recur at approximately 27-day intervals owing to the rotation of the sun), the m.u.f. and the critical frequencies increase and eventually reach peak values. At this stage two forms of radio fade-out then often occur on frequencies, for which reflection in the ionosphere normally takes place.

The first of these is the *sudden ionospheric disturbance* (*s.i.d.*) which is characterised by the abrupt disappearance of signals for periods of a few minutes up to an hour or two, but only during daylight. These disturbances have been found to be related to chromospheric eruptions on the sun. Radiation on the lower frequencies down to about 2 Mc/s is affected to a greater extent than very high frequencies over 30 Mc/s, but very low frequencies of 10-100 kc/s undergo enhanced reflection by the *D*-layer.

The second type of fade-out, which is of a more serious nature, occurs under *ionospheric* storm conditions. During such a storm the ionization in the *F*-layer is considerably reduced and its height is subject to great variation particularly over parts of the world near the polar regions. Transequatorial signal paths are seldom affected; in fact the m.u.f. across the equator is frequently at its highest under these conditions.

The ionospheric storm originates on the sun at the same time as the faster-moving radiations which cause the sudden ionospheric disturbances but in this case the effect is due to the arrival at the polar regions of slower-moving charged particles which take some two days to reach the earth. These ionospheric storms are accompanied by magnetic storms, and the earth's magnetic field then suffers considerable variation and auroral reflection occurs. The onset of the ionospheric storm is generally slow, taking several hours to reach a maximum, and fade-outs last with diminishing intensity for two or three days. The higher frequencies (up to the normal m.u.f.) are most severely affected.

The *solar radiation* which is causing an s.i.d. may actually be heard on radio receivers, particularly on frequencies above 20 Mc/s, in the form of loud bursts of noise or hiss. At times this radiation is so intense that it blots out normal signals.

Because of the sun's rotation period, the peak values of the m.u.f., the sudden ionospheric disturbances and the ionospheric storms all tend to show a 27-day repetition cycle. However, many other variations are liable to occur between subsequent 27-day cycles and this period only serves as a useful guide.

Fading. When propagation takes place through the ionosphere or the troposphere, the received waves are seldom constant in intensity but are continually varying due to the changing conditions in the atmosphere. By " fading," as compared with " fade-outs " described above is meant the relatively rapid variations which last for perhaps a few minutes or seconds or even short time periods.

At any instant the fading characteristics may be different over a relatively narrow frequency band. For example, in telephony or television reception the sidebands may fade at different rates and depth relative to the carrier and to one another, thus producing bad distortion. This is known as *selective fading.*

Fading may generally be classified under one or more of the following headings:

(*a*) Interference fading,
(*b*) Polarization fading,
(*c*) Absorption fading,
(*d*) Skip fading.

Interference fading results from a condition similar to that where phase interference occurs when two or more waves from the same transmitter arrive at the receiver over paths differing slightly from one another in length. Due to fluctuations in the ionosphere or troposphere—according to the frequency in use—the paths covered by the waves will be continually changing in length. Because of the irregularities of the transmission path, the wave finally received is really the summation of a number of waves of relatively low intensity and of random phase.

Such fading also occurs when different order multiple hop modes are propagating simultaneously, e.g. one and two hop as in Fig. 12.17.

Bad interference fading occurs when the ground waves and sky waves are of comparable amplitudes: this is, of course, an effect generally restricted to the lower frequency range. Very rapid fading, known as *flutter*, which is often experienced in long-distance reception and also frequently in local television reception, can likewise be attributed to interference effects.

Polarization fading, as its description implies, results from a continuous change in the polarization of the wave reflected from the ionosphere. These changes are in the main brought about by the effect of the earth's magnetic field on the ionosphere during the reflection process.

Absorption fading is caused by variations in the degree of absorption as the wave passes through the ionosphere or troposphere, and is generally of a longer period than either interference or polarization fading. A sudden ionospheric disturbance (s.i.d.) is an extreme case of this form of energy loss.

Owing to the fact that the m.u.f. for a given path is constantly changing over short periods of time, a receiver just within the skip distance may experience *skip fading*. This will occur when the m.u.f. temporarily drops below that for the particular path and frequency in use so that the skip distance lengthens and brings the receiver within the skip zone. Such fading can be very deep and abrupt to the extent that for a time the signal completely disappears. Obviously this kind of fading is more likely to occur when the frequency in use is near the m.u.f.

Tropospheric Propagation: 50 Mc/s-30,000 Mc/s

Considering a wave that is being radiated from an aerial relatively close to the ground, as this wave travels outward it will be attenuated by an amount which depends on the nature of the soil or water and the vegetation and the buildings over which it passes. The attenuation will be less over water. At frequencies greater than 400 Mc/s, there will be an increasing tendency towards absorption attenuation due to the presence of water vapour droplets in the atmosphere. This effect is most marked in the microwave bands.

As the wave reaches the " optical " horizon, which will be decided of course by the respective heights of the receiver and transmitter aerials, the earth's curvature will cause a shadow effect. Under some conditions the wave will be rapidly attenuated beyond this point. At this stage, however, further and rather variable factors become important. The first of these is *diffraction* which, as with light rays, is a bending of the ray path around an opaque obstacle. According to prevailing weather conditions the wave will also be *refracted* around the earth's curvature to a greater or lesser degree. *Refraction* can be defined as a gradual change in the direction of propagation caused by progressive changes in the dielectric constant (k) of the medium, which in this case is the earth's atmosphere. This effect is important close to the earth's surface, up to heights of the order of 1 kilometre or so. In order that refraction shall cause the wave to follow the curvature of the earth, it is necessary for the dielectric constant to decrease with increasing height. This is the normal condition, and hence there is always some refraction taking place, bending the waves in a favourable direction. The presence of refraction under normal conditions is usually taken into account in propagation calculations by assuming plane geometrical propagation in straight lines over an earth having a radius 4/3 that of the actual earth.

Changes in k will depend mainly on air temperature and humidity, and are therefore directly related to the weather. The normal value of k is around 1·0003, with a decrease with height of 4×10^{-5} parts per kilometre. This is an inconvenient number to use in practice, and the more familiar unit is N, where N is given by $(k-1) \times 10^6$. The normal surface value of N is then 300, and the lapse rate 40 per kilometre. The surface value of N can vary between 270-350, from cyclonic to anticyclonic weather conditions. The lower part of the atmosphere where these effects are important is known as the *troposphere*: see Fig. 12.10.

If, as occurs under certain weather conditions, there is a sudden diminution in k with height, *reflection* of the wave will take place at the boundary between the two air masses concerned.

The passage of a wave over a relatively long path (50-500 miles) can afford several opportunities for reflection and refraction at differing regions of the troposphere, due to the continually changing values of k along the path. Even under relatively stable weather patterns there will exist isolated regions of air of differing dielectric constant, from which the signal will scatter in random fashion, the various components adding up to give enhancement of the signal at the distant receiver. Under these conditions the path length (giving the strongest received signal) is continually fluctuating as the reflection sources vary. Often the change in path length is small compared with a wavelength, but it can become appreciable, leading to the rapid fading often associated with tropospheric propagation under average conditions.

Under the appropriate weather conditions it is possible for a region to be established in the troposphere in the form of a narrow layer of air of a certain k value trapped between two regions of lower k. This is referred to as a *tropospheric duct* and can occur over quite long distances in a stable form, and at greatly varying heights, from almost ground level up to several hundreds of metres. A wave entering this duct will propagate by multiple reflection along it with very low attenuation, being " guided " in a manner somewhat analagous to microwaves in a waveguide. This gives rise to very considerable and stable enhancement of received signals over periods of time up to several hours, before the duct disperses. This effect is most common when hot days are followed by rapid cooling at night under a prevailing slow moving anti-cyclone.

These tropospheric effects are detected on all frequencies but are important chiefly in v.h.f. communication because only by this form of propagation is the range extended beyond the optical horizon on frequencies above 70 Mc/s except in very abnormal and infrequent conditions.

It has already been explained that the propagation of radio waves round the curvature of the earth is made possible by the fact that they undergo reflection and/or refraction as the result of the small decrease in the dielectric constant (k) and hence in the refractive index with height. The actual dielectric constant of un-ionized air is very little more than unity and the changes of k with height are extremely small. The presence of water vapour increases the value of k and is indeed the main factor in determining the degree of refraction and/or reflection which takes place. Water in droplet form, such as fog or rain, has little effect on metric wavelengths (1-10m), though it causes absorption and scattering of centimetric waves (below 1m). It is the invisible water-vapour content that is important in affecting the value of k. The maximum amount of water vapour that can be held by air increases with the temperature and tropospheric propagation effects are therefore more marked

TABLE 12.1

RSGB BEACON STATIONS

Details of beacon stations operated by, or on behalf of, the Radio Society of Great Britain. Some of the services may be subject to variation and reference should be made to the current issue of the RSGB *Bulletin*.

Call-sign	Location	Nominal Frequency	Emission	Aerial Direction
GB3ANG	Craigowl Hill, Dundee	145·985 Mc/s	A1	S
GB3CTC	Redruth, Cornwall	144·10 Mc/s	A1	North-East
GB3GEC	Hammersmith, London	431·5 Mc/s	A1	
GB3GI	Strabane, N.I	145·990 Mc/s	A1	
GB3LER	Lerwick	145·995 Mc/s	A1	S
GB3LER	Lerwick	70·305 Mc/s	A1	N/S
GB3LER	Lerwick	29·005 Mc/s	A1	N/S
GB3VHF	Wrotham, Kent	144·50 Mc/s	A1	North-West

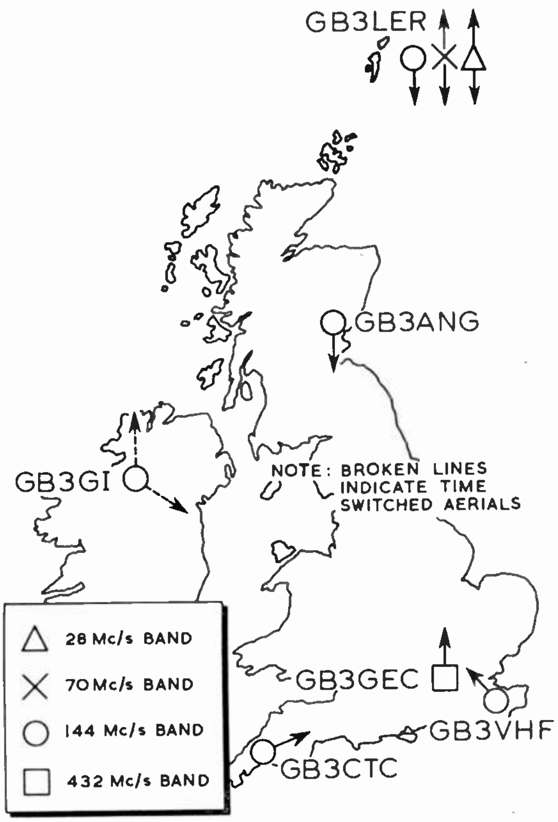


Fig. 12.21. V.h.f. and u.h.f. beacon stations in the British Isles. The broken arrows indicate time-switched aerials.

in warm moist air than in cold dry air. Because the temperature of the atmosphere lapses with height at some 6·5 C. per kilometre (i.e. 2°C per 1000 ft.), the moisture content, or humidity, becomes very low as the altitude increases, and therefore tropospheric propagation is mainly confined to the lower atmosphere at heights of up to 2 km (about 6000 ft.).

A knowledge of meteorology will be found useful in the forecasting of tropospheric propagation conditions and the interested reader is advised to study the various publications which are available including References [9], [10]. As a general guide, the main effects of weather may be stated as follows:

(*A*) For maximum extended range there should be either a condition of high humidity at ground or low level followed by a rapid decrease in humidity with height or an *increase* in temperature with height (temperature inversion). Generally the two conditions occur simultaneously but not necessarily.

(*B*) For normal coverage, i.e. no extension of range by abnormal tropospheric effects, a standard rate of lapse of temperature and humidity is required. This standard rate was defined on page 12.14. Occasionally, during conditions giving rise to a high temperature lapse rate and an associated low humidity lapse rate with altitude, propagation becomes sub-normal in as much as unfavourable refraction occurs, reducing slightly the working range. This is fairly common but the reduction in range is not very marked in practice.

Condition (*A*) is satisfied by fine, warm anticyclonic, settled weather, and condition (*B*) by rough, cold cyclonic weather. Between these two extremes there will, of course, be intermediate weather conditions where from simple observation it may be difficult to forecast radio conditions.

Anticyclonic conditions, when established, often remain much longer than the faster-moving depressions, and patches of favourable tropospheric conditions may therefore last for a week or two. In summer and warmer weather generally, tropospheric propagation will be more reliable than in cold winter months. In stable weather, conditions are most suitable over *land* for the formation of temperature inversions after sunset. As the ground cools off, the air in contact with it is likewise cooled while the air above remains almost at daytime temperature. During the day, there will be less turbulence caused by uneven heating of different ground surfaces. Over the sea there is a much smaller diurnal variation and conditions are influenced more by the weather system prevailing and by the sea temperature.

In the frequency range under consideration, 50 Mc/s–3000 Mc/s, the effect of frequency on observed tropospheric conditions is not very marked. If, for instance, good conditions are observed on 70 Mc/s, conditions on 144 Mc/s and 420 Mc/s will be enhanced although not always to the same degree. This is quite different from the case of ionospheric propagation on lower frequencies.

It is often useful to observe the reception of television signals, particularly those in Band III (175–215 Mc/s). Any suggestion of co-channel interference in the form of patterning or even lockable pictures is a sure indication of the presence of anomalous tropospheric propagation. The u.h.f. television services will provide a marker to conditions on the 432 Mc/s band, but may not themselves exhibit severe interference during band openings due to the greater frequency differences (the upper end of Band V extends into the 800 Mc/s region).

The v.h.f. sound services in Band II (88–108 Mc/s) and television in Band I (40–65 Mc/s) are less reliable as an indication of *tropospheric* conditions since they are both frequently subject to co-channel interference from sporadic *E ionospheric* propagation. They are of course useful as an indication of the presence of sporadic *E*, but such interference requires to be very severe before there is any prospect of the E_s m.u.f. reaching the 144 Mc/s band.

In addition to the use of commercial broadcasting stations as a guide to propagation conditions, RSGB operates a number of beacon transmitters, mainly at v.h.f., designed and positioned to provide markers of anomalous propagation over several of the more likely paths in the British Isles. Some beacons use aerials beaming directly towards the target areas to indicate the presence of long range tropospheric propagation, while others are beamed North to mark the onset of auroral radio reflections. A full list of the Society beacons is shown in **Table 12.1** and their positions and directions shown in **Fig. 12.21**. In addition certain European countries also have beacons in operation, and details for each country may be obtained from the respective National societies or extracted from their journals.

The service from Wrotham using the call-sign GB3VHF may also be used as a relative frequency standard, being maintained and checked normally to better than 250 cycles per second.

Distances covered by tropospheric propagation will,

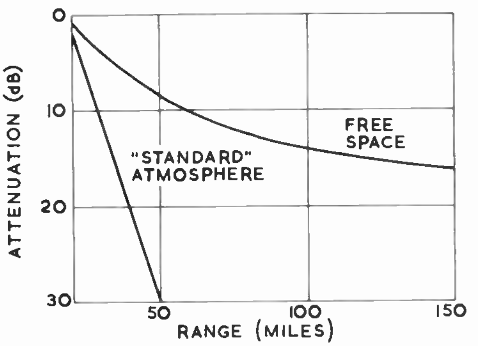


Fig. 12.22. Attenuation of the received signal by a so-called standard atmosphere beyond a horizon of 20 miles.

of course, depend to some extent on the heights of the transmitting and receiving aerials and on power, aerial gains and receiver noise factor. Beyond the optical range and under optimum conditions, the field strength decreases at approximately the free-space rate (i.e. inversely with distance). At times peaks up to 6db above the free-space values may be obtained. Typical range extensions of four or more times may be obtained, i.e. if under poor or " standard " atmospheric conditions the range is 50 miles similar signal strengths may under good conditions be obtained up to 200 miles or more. **Fig. 12.22** illustrates the attenuation under " standard " and free-space conditions beyond a horizon of 20 miles.

As a general rule, on very high frequencies results are not materially different according to whether the aerial polarization is horizontal or vertical. It is of course necessary for both receiving and transmitting stations to employ the same polarization for optimum signal transfer. This is less significant however when one of the stations is a mobile and in motion relative to the other.

Normal Tropospheric Ranges

By making certain assumptions, it is possible to predict the field strengths to be expected at v.h.f. for given ranges and aerial heights. The curves of **Fig. 12.23** (70 Mc/s) and **Fig. 12.24** (144 Mc/s) show the field strength laid down at near-ground level for various distances and for three effective aerial heights *over a plane earth* for an e.r.p. of 1 watt. In order to calculate the field strength at the actual height of the receiving aerial, a correction factor, expressed in db and obtained from the lower curve, is added to the figure obtained from the upper curve to obtain the field strength expressed in db relative to 1 microvolt per metre.

Example At 144 Mc/s, what signal is received at a height of 30 ft. at a point 10 miles from a transmitting aerial 150 ft. high with an effective radiated power of 100 watts?

From Fig. 12.24 the ground level field strength is +10db and a further +20db must be added for the transmitter e.r.p. The actual field strength is therefore 10 + 22 + 20 = +52db relative to 1 μV/m = 400 μV/m.

Inspection of the height gain curves shows a discontinuity in the slope of each at the low receiving aerial heights. Above the region of this change, the height gain is the standard 6db for each doubling of height or linear law explained earlier in

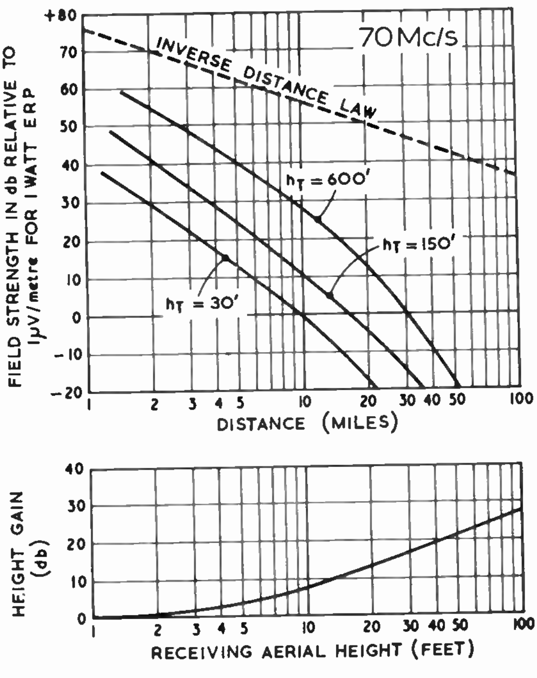


Fig. 12.23. Variation of field strength with distance and aerial height for propagation over a plane earth at 70 Mc/s. Use of the curves is explained in the text. Transmitting aerial height = h_T.

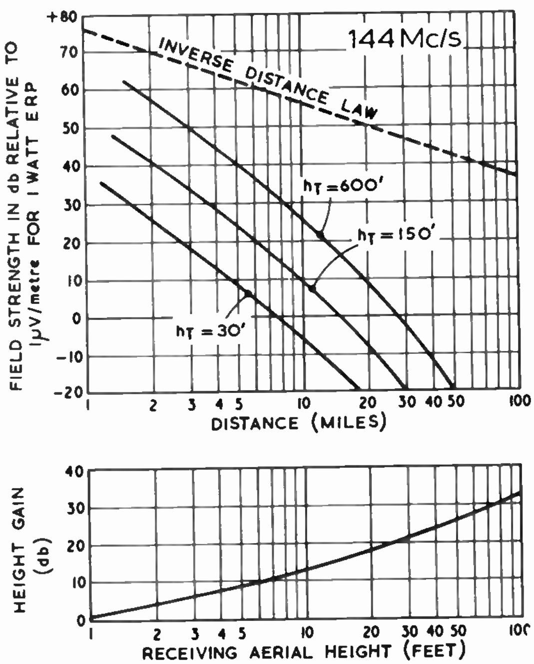


Fig. 12.24. Variation of field strength with distance and aerial height for propagation over a plane earth at 144 Mc/s.

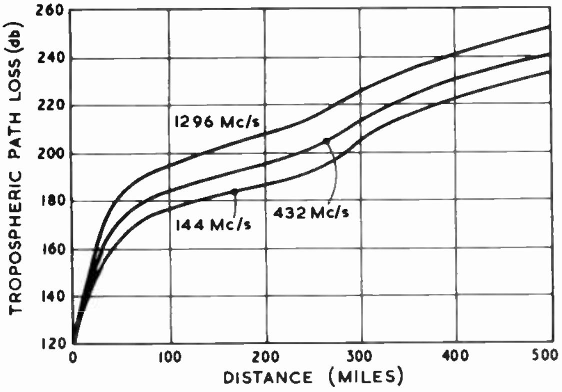


Fig. 12.25. Path loss for varying distances for v.h.f. and u.h.f. amateur bands. Use of this loss figure will give the expected field strength for 50 per cent of the time, and is typical of amateur operation.
(With acknowledgments to the American Radio Relay League Inc.)

this chapter (see page 12.4) for propagation over a plane earth. Below this region, the height gain is reduced because the received wavefront becomes increasingly subject to attenuation by the ground in the same manner as a ground wave propagated at medium frequencies.

The heights of the transmitting aerial shown are strictly those above the mean surrounding terrain and so the curves may be applied to aerials on elevated sites with the aerial height calculated relative to the surrounding levels (h'_T in Fig. 12.5(b)).

It is also possible to use the technique of insertion loss to estimate the level of received signals at v.h.f. This involves expressing the attenuation of the path between the aerials as a series loss in decibels which may subtract from the effective gain of the station at each end of the circuit. The method is described more fully in the section dealing with space communications and is covered in detail in Reference [11]. The information for tropospheric path loss over a plane earth shown in Figs. 12.25 and 12.26 is taken from that reference. The value of N (path loss in db) may be used in the formula stated later in this chapter.

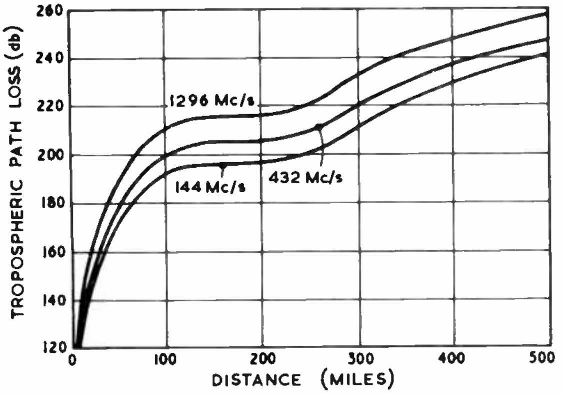


Fig. 12.26. As Fig. 12.48, but for 90 per cent of the time. These curves should be used for reliable schedules with other stations.
(With acknowledgments to the American Radio Relay League Inc.)

Propagation by Scatter

There are times when the more orthodox modes of propagation fail to provide a satisfactory means of communication and one or other of the various forms of propagation by scatter may then offer a useful alternative: these include *tropospheric scatter*, *ionospheric scatter* (which may be either *forward* or *backward*), *auroral reflection* and *meteor-trail scatter*.

Tropospheric Scatter. This is similar in character to ordinary tropospheric propagation except that advantage is taken of small local variations in the atmosphere, such as clouds, which afford partial reflection and refraction at a number of isolated points along the transmission path. However, the attenuation is much greater than by the normal tropospheric mode and it becomes necessary to use relatively high power and aerials with high gains. The distances which can be covered in this way may be as much as 500 miles.

Ionospheric Scatter. In this form of propagation ionospheric " clouds " are used to provide a reflecting surface for low-angle radiation of relatively high frequency. The reflection may take place also from the ground so that by a series of hops the wave is propagated over a more or less straight line to a distant point: this is known as *forward scatter*. It is a comparatively regular phenomenon and is being increasingly used for communication purposes but like tropospheric scatter it requires high power because of the considerable attenuation: often the power required exceeds that normally permitted to amateurs. The frequencies employed are of the order of the m.u.f. for the *F2*-layer or somewhat higher, and will usually lie in the range 35-60 Mc/s. The exact frequency will of course depend on the period in the sunspot cycle.

Partial reflection may occur at acute angles from a remote ionospheric cloud and from the ground, and the reflected radiation may be picked up by a receiver located relatively near to the transmitter: this is known as *back scatter*. It is of particular interest to amateurs since it can be used for communication over distances *within* the skip zone,

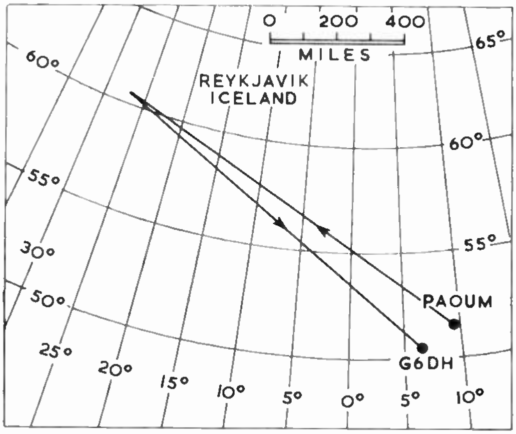


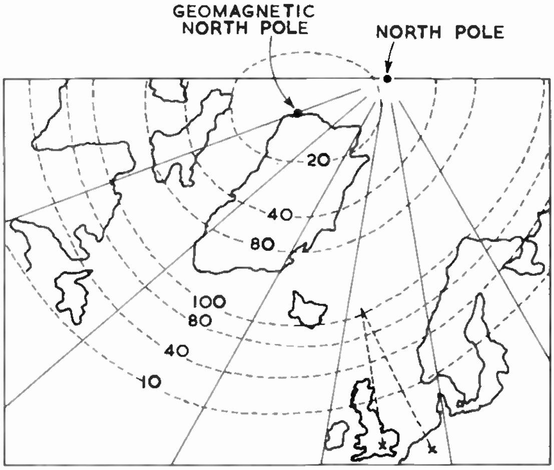
Fig. 12.27. Back-scatter propagation. A skeleton map showing the probable path by which signals radiated from PAOUM were received at G6DH. The time was 14.00 GMT on a winter day. Both aerial beams were directed approximately north-west.

especially in the 21 Mc/s and 28 Mc/s bands. For example, the skip distance may be 1200 miles and stations within this distance may not be able to establish communication when the receiving and transmitting aerials are directed towards each other over the shortest path of, say, 100 miles. However, if the aerials are *both* directed to a distant point—of the order of 1500 miles from either station—it will be found that weak but consistent reflection occurs: see **Fig. 12.27.** According to relatively local ionospheric conditions the best direction for this form of partial reflection (or back scatter) will change with the time of day and also from day to day. Therefore, both the transmitting beam aerial and the receiving beam aerial should be rotated simultaneously around the compass in order to establish the best distant reflecting area in the ionosphere. Generally in the temperate latitudes of the northern hemisphere the best directions will be found between north-east through south to north-west and seldom to the north: see **Fig. 12.28.** This is because the regions of higher ionization generally lie to the south in these latitudes. Low-angle radiation is required for this purpose and the normal Yagi beams are quite suitable.

Auroral Reflection. Similar to ionospheric back scatter but less frequent is *auroral reflection* (or *auroral scatter*) from the auroral zones around the polar regions. This occurs during the ionospheric or magnetic storms when there is marked auroral activity, and the frequencies at which this kind of reflection has been known to take place extend up to about 150 Mc/s. The reflected waves are generally characterized by a very rapid fluctuation (10-50 cycles per second) which makes telephony difficult but has no harmful effect on telegraphy. To make use of auroral reflection, directive aerials are essential and they should be pointed in a northerly direction in the northern hemisphere (southerly in the southern hemisphere). In this way communication can sometimes be established between stations located some 500 miles or more apart: see **Fig. 12.29.** Since auroral activity disrupts the ordinary forms of ionospheric propagation, a transmission path which crosses a polar region suffers a complete fade-out during such periods, and auroral reflection may then offer a means of maintaining contact.

Meteor-trail Scatter. This phenomenon is of very short duration, possibly less than a minute. When a meteor passes through the upper atmosphere it produces a trail of ionization which will, for a short period lasting from less than a second to two minutes, cause partial reflection of v.h.f. waves. Most trails are in the region of the *E*-layer (i.e. about 100 km) and consequently the distances covered are similar to those of sporadic-*E* reflection, i.e. up to 1200 miles for forward scatter. The aerial should of course be directive and should be oriented in a similar manner to that described for ionospheric or auroral back scatter but generally to a point rather less distant—say 500 miles. The best directions vary with the time of day and the rate of meteor occurrence. Any frequency band between 21 Mc/s and 144 Mc/s is suitable, the 70 Mc/s band being a particularly good one. Since bursts of reflection are of short duration and low intensity, high power and good aerials and receivers are essential. It is also necessary to pre-tune the receiver accurately to the transmitter frequency so that when a burst occurs contact is established immediately.

The incidence of meteor activity is subject to a seasonal variation which is illustrated in **Fig. 12.30.** In addition there are specific showers of extra high intensity which are given names. These are shown in Fig. 12.30 and it is during these



12.29. Distribution of visible aurora in northern hemisphere. The contours represent the percentage of days on which aurora were seen. The propagation path for communication by auroral reflection between England and Holland is indicated. Although the rate of incidence of visible aurora is related to the incidence of radio aurora, it is important to note that the presence of the one at any particular time does not necessarily mean that the other may also be observed.
(Courtesy US National Bureau of Standards)

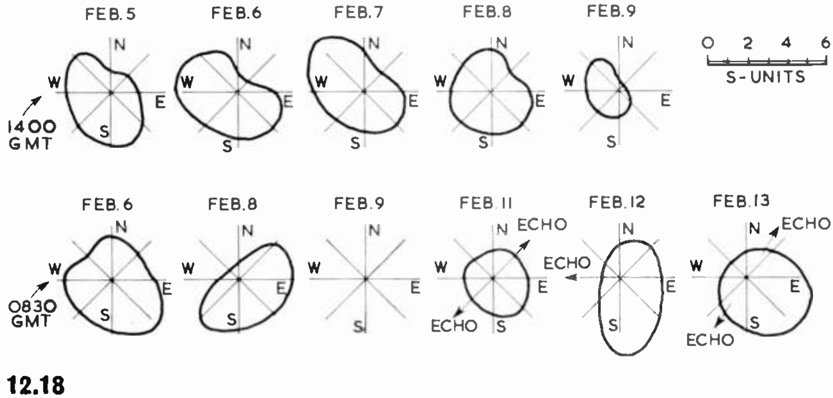


Fig. 12.28. Typical day-to-day variation in back-scatter directional characteristics. The absence of a pattern on February 9 at 08.30 GMT means that no signals could be heard. This was due to ionosphere disturbance.

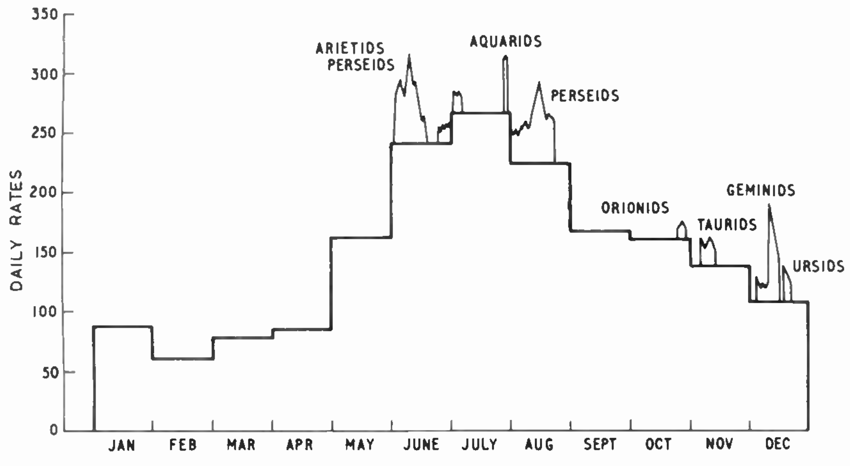


Fig. 12.30. Seasonal variation of meteor activity. The recognized showers are also shown.

showers that normal amateur meteor scatter contacts are made. Such contacts are scheduled in advance and are by means of high speed c.w. using a special reporting code and repetitive groups during each period of sending, governed by a time sequence previously agreed with the other station with which the schedule is being kept.

LONG DISTANCE COMMUNICATION USING OBJECTS IN SPACE

The modes of propagation discussed so far in this chapter have been those associated with quasi-terrestrial phenomena, e.g. the troposphere, the ionosphere, ground waves, etc. They all come about as a result of natural phenomena associated with the earth and its immediate surroundings.

There also exist certain extra-terrestrial phenomena, both natural and artificial, which can be exploited by specialised techniques to give rise to long range contacts over the earth's surface, but using actual path lengths often substantially longer than those associated even with ionospheric reflection from the upper layers. In the extreme cases, the distance covered in miles over the earth's surface between the transmitting and receiving stations is negligible compared with the distance the signal has to travel in space.

Although communication by these methods is determined very much more by the parameters of the transmitting and receiving stations than by what happens to the radiated waves passing between them, nevertheless they can be included within the generic term propagation.

The techniques to be described involve either the reflection from natural or artificial bodies in space of radio waves emitted from the earth, or the interrogation of some active artificial device in space, which itself then re-transmits the original signal intelligence back to earth using some different frequency, and possibly different systems of modulation.

Passive Artificial Satellites

Various space agencies throughout the world have launched, and are continuing to launch into space, different types of passive reflecting satellites. An example of these is the United States *ECHO* series, which involve placing in orbit around the earth various balloons which are coated on their exterior with a material of high conductivity and hence efficient as reflectors of radio waves. Successive experiments on a commercial scale have been carried out using very high transmitting powers and ultra-sensitive receivers, and signals have been "bounced" back to earth from these devices. They are, however, very small objects in relation to the distances involved, and it is not thought likely, at the moment, that amateur equipment is sufficient to overcome the high overall path loss incurred due to the very small amount of signal scattered back from the satellite. The actual mechanism of transmission is identical to that of "moonbounce" to be described later. The main difference lies in the faster tracking speeds needed for the ground aerials to follow the relatively fast moving satellite.

Passive Natural Satellites

In addition to the various artificial earth satellites launched from time to time, there exist in the sky two natural bodies of sufficiently large size in relation to the distance from earth to entertain the possibility of bouncing radio signals from their surfaces back to earth. Unfortunately one of these, the Sun, is such a powerful source of radio noise in its own right that there is no possibility for amateurs of distinguishing

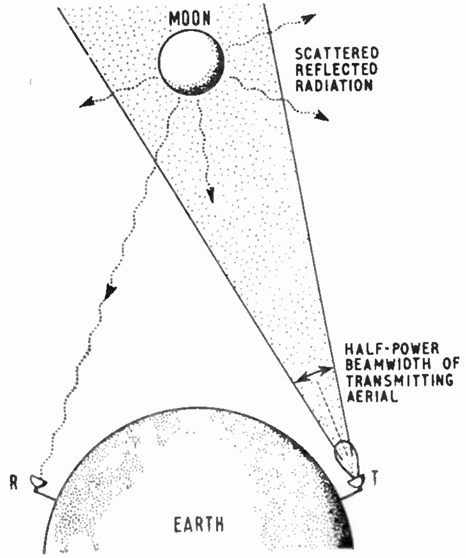


Fig. 12.31. Basic arrangement of moonbounce communication.

signals among the solar radiation which is continuously emitted. The other is the Moon which is a natural satellite of the Earth, and is in comparison with the Sun, an extremely quiet radio source. It is therefore eminently suitable for use as a reflector of radio waves and has given rise to a technique known as "moonbounce."

The basic arrangement for moonbounce is shown in **Fig. 12.31**. The transmitting aerial *T* is arranged to direct as much signal as possible at the moon. Because the moon subtends at the earth an angle minutely small compared with the beam width of any practical aerial, only a small amount of the radiating energy does in fact illuminate the moon's surface. A certain amount of this is absorbed, and the remainder scattered into space in random fashion. Some of this scattered radiation will return to earth, and will be received with equal facility by any place on the earth's surface which can see the moon. It is interesting to note that the strength of the received signal does not depend upon the earth distance between the transmitting and receiving stations, and in consequence moonbounce communication is possible in the limit between stations on opposite sides of the earth.

The frequency used for moonbounce communication must be sufficiently high to pass through the ionosphere with little or no absorption on the outward and return journeys. It must also be high so that the effective scattering area of the moon's surface is large in terms of wavelengths, and to permit the use of very high gain aerials of reasonable physical size at each earth terminal. This is essential if sufficient transmitted power and receiver sensitivity is to be achieved to overcome the enormous losses incurred due to the extreme length of the effective signal path.

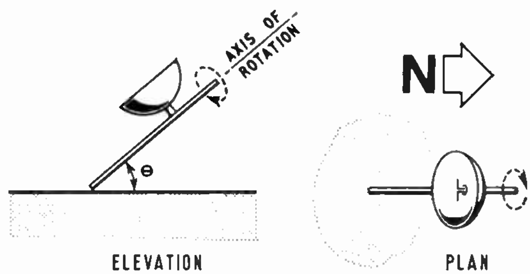


Fig. 12.33. Polar mount for moonbounce tracking aerial. The angle θ is the complement of the angle of latitude of the station.

The optimum frequency for moonbounce, or indeed any earth-space communication is in practice determined by the need to achieve best signal-to-noise ratio at the receiver in the face of the noise generated in space and incident upon the receiving aerial. This is looking up at the Moon and the aerial, because of its relatively wide beamwidth, " sees " the regions of space all around and beyond the moon itself. The variation of the effective noise temperature of space with frequency is shown in **Fig. 12.32** which shows that frequencies above 400 Mc/s are to be preferred. The overall moonbounce path-loss is proportional to the square of the frequency used, but for a fixed physical size of aerial, the aerial gain *at each end* increases as the square of the frequency. The net overall improvement to be obtained in signal-to-noise ratio is 6db for every increase in frequency by a factor of two. This does not of course take into account the increased feeder losses, poorer receiver noise figure, and greater difficulty in generating adequate power at the increased frequency. Allowing for these, the present highest frequencies used for amateur moonbounce work are in the 432 Mc/s or 1296 Mc/s bands.

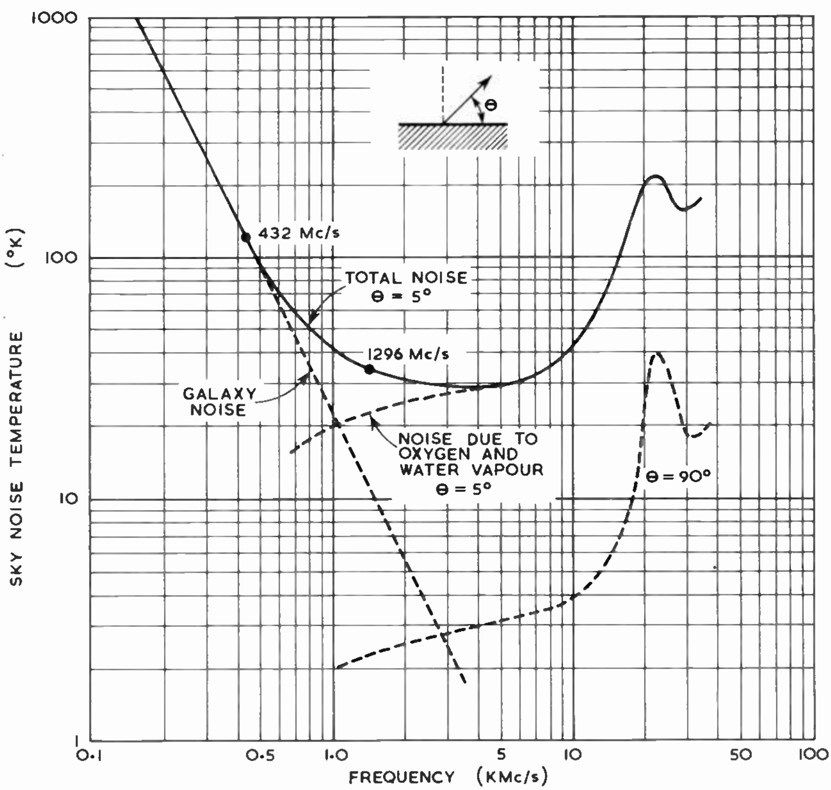
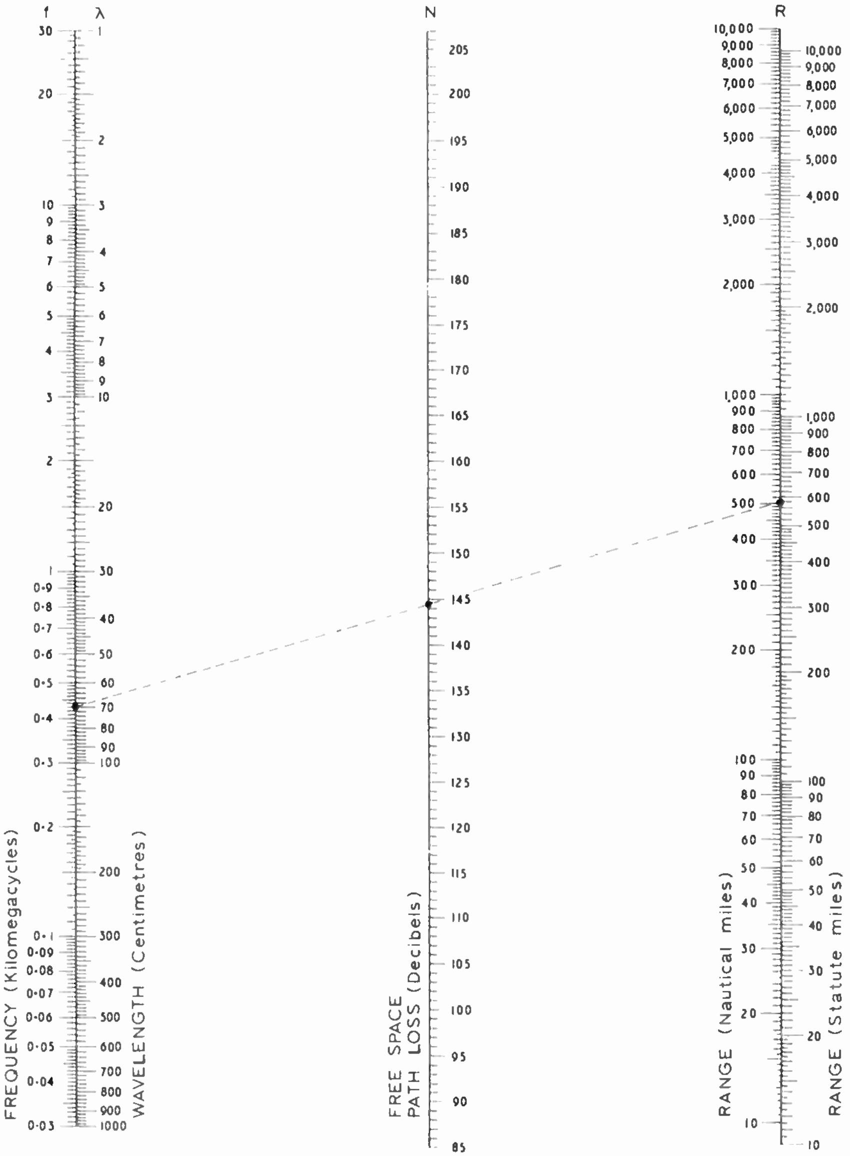


Fig. 12.32. Noise temperature of space as a function of operating frequency. The angle is the elevation of the main lobe of the receiving aerial above the earth horizon.

Typical ground station equipment for moonbounce requires a transmitter output power exceeding 100 watts, net aerial gains of the order of 15–20db, and receivers of 500 c/s bandwidth and a noise figure better than 2–3db. Full details of the required station performance can be found in the various moonbounce references given at the end of this chapter. Apart from the need for really high e.r.p. and low noise receivers, the other requirement is an ability to aim the aerial system at the moon, and to follow its motion in the sky. This progression is not excessively rapid, being some 15° angular displacement per hour, and can readily be followed by some simple manual system, reset from time to time as the moon sweeps across the main beam of the aerial, which is rarely much less than ±5° to the 3db points, even for high gain systems. The aerial should be rotated about a "polar mount," i.e. an axis parallel to that of the earth. For a latitude of 52° N the axis of rotation of the aerial should lie in a N-S plane, and be inclined at an agle of 38° to the ground. This is shown in **Fig. 12.33.**

As with other communication systems involving very low received signal levels, maximum use of c.w. is desirable together with an ability to tune exactly to the known frequency of the other station.

Moonbounce communication is almost entirely achieved by means of pre-arranged schedules with other stations, because of the specialised technique involved. It is most unlikely that normal operating techniques involving random calling will produce the desired contacts. Further information on this subject can be found in Reference [14, 15, 16].



Free-Space Path Loss

All space communication involves calculation of the *free space path loss* between the ground station and the satellite (active or passive). This loss may be usefully described as the attenuation experienced when signals are sent and received on two half-wave dipoles spaced apart by some given distance. If this loss is expressed in decibels it can be incorporated into a single formula to determine the power level received at the far end for a given transmitter power, aerial gains, feeder losses and at a given frequency of operation. Power levels are usually expressed as decibels relative to 1 watt and written as dbw. The received power, i.e. the power available in the receiver input resistance, is given by the expression:

$$L_R = L_T + (G_T - F_T) + (G_R - F_R) - N$$

where

L_R = received power level in dbw

L_T = transmitted power level in dbw

G_T = transmitting aerial gain relative to a halfwave dipole in db

F_T = transmitting aerial feeder loss in db

G_R = receiving aerial gain relative to a halfwave dipole in db

G_F = receiving aerial feeder loss in db

N = path loss in db

Fig. 12.34. Free space path loss as a function of frequency and range. If the nomograph does not cover the desired ranges, multiply the *f* or R scale by 10x and add 20x db to the path loss.

(*With acknowledgments to Scientific Atlanta Inc.*).

Active Artificial Satellites

An " active " artificial satellite is one which does not act (intentionally) as a reflector of incident radio waves. It is a receiving-transmitting station in its own right, powered either by internal batteries, by derivation from a bank of solar cells which collect energy from the sun's radiated heat and light, or by a combination of both methods, the solar power being used to re-charge the batteries, which then operate during the times when the satellite is on the " dark " side of the earth.

The equipment on board the satellite is used to receive signals transmitted from the earth, and to re-radiate on a different frequency, often well removed from the incoming one, to be received again on the earth. The advantage of the active satellite over its passive counterpart lies in the effective power *gain* which obtains at the device using its own transmitter compared with the tremendous signal *loss* incurred when relying solely on the reflecting and scattering properties. For this reason the ground station equipment may be far less complex, requiring only modest powers and aerials, and average receivers for use with narrowband low altitude vehicles. This should not be confused with the elaborate ground station requirements for commercial active satellites, such as *Early Bird*, which are wide-band systems, and require to be placed at extreme altitudes to achieve synchronous orbit with the *earth*, i.e. to maintain the same position in space relative to the ground stations at all times.

As in the case of moon-bounce, the actual propagation paths involved, from ground to vehicle, and from vehicle to ground, can be considered true free-space propagation, assuming that the frequencies used are sufficiently high to be unaffected by the ionosphere. The path losses involved can then be obtained from standard nomographs, knowing the vehicle height, and the ground station parameters determined using the known receiver sensitivity and transmitter output power of the space vehicle. Unlike the case of the passive satellite where the calculation of the ground-space path is tied up with the return space-ground path, related directly by the scattering loss of the reflecting body, the ground-space and space-ground path calculations for the active satellite are often independent. This is achieved by the use of a.g.c. in the satellite translator system, ensuring that full output from the satellite transmitter is obtained using a relatively low level of received signal from the ground station. Any subsequent increase of power of the transmitting ground station will not achieve a corresponding increase in output power from the vehicle or an increase in signal received at the far-end ground station. The latter must be sufficiently sensitive to receive the translator signal when working at its full output power.

As in the case of moonbounce, the frequencies used are usually in the 144 Mc/s or 432 Mc/s bands, although it would be technically possible to use the 70 Mc/s or 1296 Mc/s bands as well, and even 28 Mc/s during periods of minimum sunspot activity, when the level of ionization intensity in the ionosphere is insufficient to affect such frequencies to any appreciable degree.

It is desirable to employ some form of tracking on the ground station aerials, although this is less important than in moonbounce, since much lower e.r.p. and aerial gains are necessary, resulting in broader lobes of radiation and reception. Unlike moonbounce, the target object is moving quite rapidly in the sky because of the relatively low orbit altitudes achieved, and tracking is therefore somewhat more of a problem, requiring almost continual adjustment for optimum results. At the time of preparing this edition, all amateur satellites in this class have been launched in near-Polar orbits, having orbit periods (once around the earth) in the region of 80-100 minutes. It is therefore possible for a ground station to be within range of several successive passes, the first being low in the sky to the east, and of very short duration. The middle passes are nearly overhead, sometimes directly so, and therefore of longer duration from horizon to horizon, typically 20 minutes. The last pass is then low in the sky to the west and again of short duration. For the more distant passes which achieve only low angles of elevation above the horizon, a conventional beam aerial rotating in a horizontal plane only is adequate. For the overhead passes it is desirable to be able to rotate the aerial in a vertical plane right through 180° of arc, from horizon to vertically upward and through to the other horizon.

Because of the relatively short " pass " times, and the limited number of channels available within the restricted passband of the translator, typically 50 kc/s within the 145 Mc/s band, operation through active satellites is usually restricted to c.w. and the most successful contacts have been made by stations accustomed to the high speed c.w. technique usually employed for meteor scatter work (q.v.).

All active satellites for amateur use launched so far have been under the control of the Project OSCAR Association of the USA, and the most successful have been translators with both input and output passband lying within the limit of the 144-146 Mc/s amateur band. A considerable amount of general and scientific information about the *OSCAR* series is to be found in Amateur Radio journals of many countries for the years 1963-1966.

Particularly useful information will be found in References [18] and [19]. These cover in detail the important calculations involved in determining the timing, direction, and position of successive passes of the satellite, knowing the launch conditions.

GREAT CIRCLES AND MAPS FOR RADIO PROPAGATION

When a radio wave travels between two stations on the surface of the earth, it follows what is known as a *great circle* path. This is the projection of the wave's path on the surface of the earth along the edge of a circle, the centre of which coincides with the centre of the earth, and which

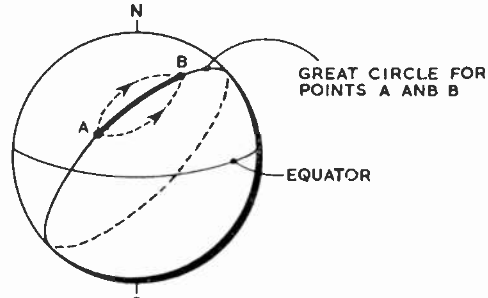


Fig. 12.35. The shortest distance over the earth's surface between points A and B lies along the great circle which includes them.

passes through the two stations A and B. This is illustrated in Fig. 12.35, which shows that in fact this path is the shortest distance between A and B *over the earth's surface.* It should be noted that different modes of propagation (ionospheric, tropospheric etc.) will produce slight divergencies from this idealized path but, within the beamwidth of practical aerials, such changes of direction may be ignored.

In order to reach a particular station at a distance, the transmitting aerial must be orientated to point along the great circle path or bearing, and of course the receiving aerial must be similarly orientated on a reverse bearing. It is usual to find maps of regions, or of the world, drawn in terms of a projection of the curved surface of the earth upon a plane surface: a common projection found in many atlases is that of the developed cylinder, ascribed to Mercator, in which lines of latitude and longitude are all straight, and the latitude is displayed as unequally spaced parallels (Fig. 12.36(a)). On such a map great circle bearings and distances between places become increasingly inaccurate as the places lie further apart, and for the purposes of assessing aerial headings, such maps are all but useless.

The correct map for use in radio work is the great circle map or azimuthal equi-distant projection, which is drawn in the form of a circle centred on the reference city. (Fig. 12.36(b).) On this type of projection, the bearing and distance *from the reference origin at the centre* to all other points on the map are both correct in great circle terms and may be read off using a straight edge laid on the map. The disadvantage of the great circle map is that it must be specifically drawn for any particular origin. For example a map centred on London will give correct bearings and distance from London to all points on the globe. A map centred on Washington, D.C., will give correct bearings and distance from there, but will look very different from the London map.

In practice a map based on London may be used with very little error for all points in the British Isles, and maps based on other cities may likewise be assumed correct for the

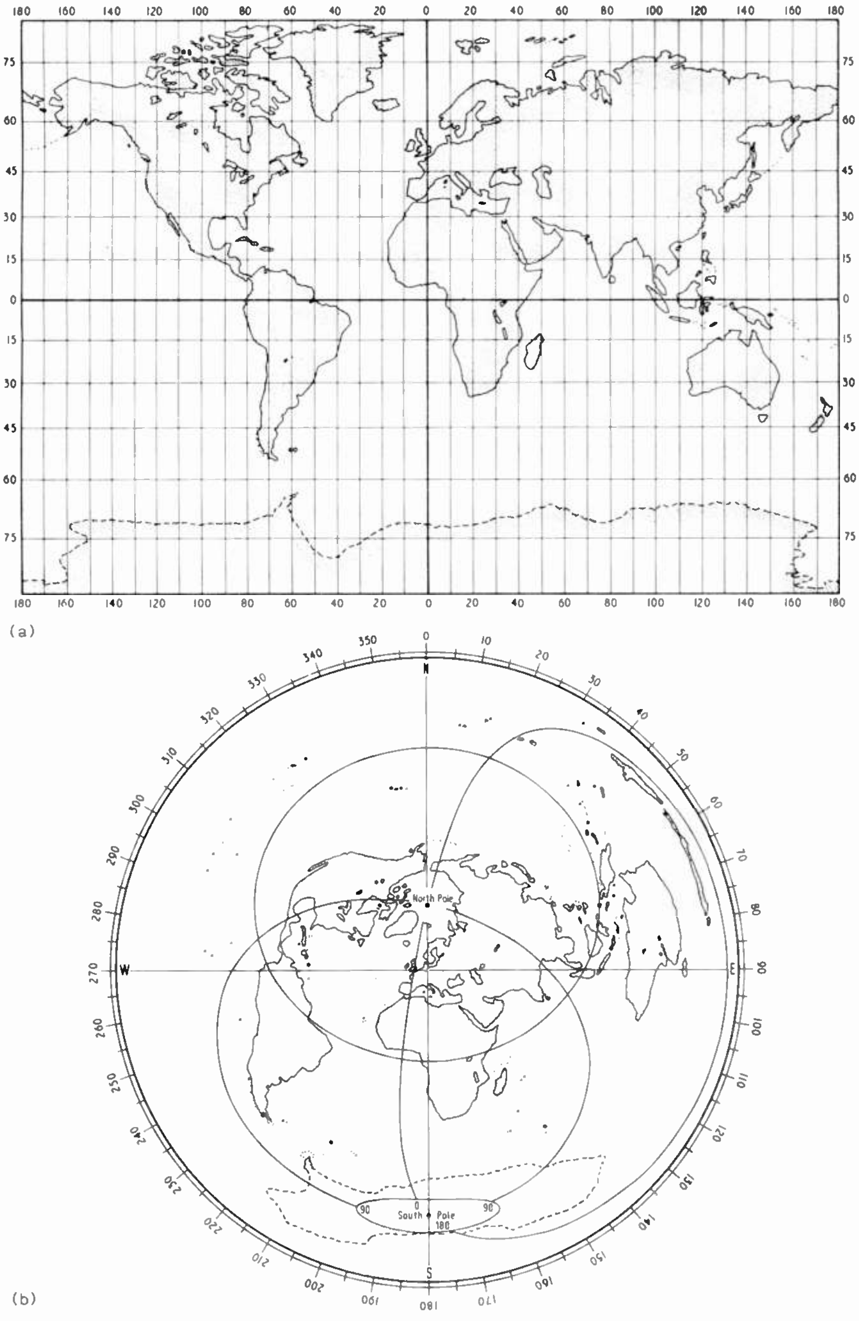


Fig. 12.36(a). Mercator's projection map of the world. (b) Great circle map of the world.

surrounding areas. A suitable great circle map for stations located in the British Isles is published by the Admiralty Hydrographic Office, and can be obtained from HM Stationery Office, or from RSGB Headquarters. It is an azimuthal equi-distant projection based on London, but as explained can be used reliably by other stations.

Overseas readers should enquire of their Government Printing agencies for the existence of suitable Great Circle Maps based on cities in their own areas. Such maps are generally available through different sources for most parts of the world.

It is possible, from a knowledge of the latitude and longitude of the two stations, to calculate the forward and reverse bearings of the great circle path between them, and also the distance involved.

Let A and B be two places on the earth's surface, as shown in Fig. 12.35, the angles α and β at A and B of the great circle passing through the two places and the distance D between A and B along the great circle can be calculated as follows.

Let B be the place of greater latitude (nearer the pole).

L_A is the latitude of A.

L_B is the latitude of B.

L_D is the longitude difference between A and B.

$$\text{Then, } \tan\frac{\beta-\alpha}{2} = \cot\frac{L_D}{2}\,\frac{\sin\frac{1}{2}(L_B-L_A)}{\cos\frac{1}{2}(L_B+L_A)}$$

$$\text{and } \tan\frac{\beta+\alpha}{2} = \cot\frac{L_D}{2}\,\frac{\cos\frac{1}{2}(L_B-L_A)}{\sin\frac{1}{2}(L_B+L_A)}$$

give the values of $\frac{\beta-\alpha}{2}$ and $\frac{\beta+\alpha}{2}$

$$\text{from which } \frac{\beta+\alpha}{2}+\frac{\beta-\alpha}{2}=\beta$$

$$\text{and } \frac{\beta+\alpha}{2}-\frac{\beta-\alpha}{2}=\alpha.$$

In the above it is convenient to take northern latitudes as positive and southern as negative.

If both places are in the southern hemisphere, $L_B - L_A$ will be negative and it is simpler to refer the calculation to the South pole making suitable conversion with respect to North later if necessary.

The distance D (in degrees) along the great circle between A and B is given by:

$$\tan\frac{D}{2} = \tan\frac{L_B-L_A}{2}\cdot\frac{\sin\frac{1}{2}(\beta+\alpha)}{\cos\frac{1}{2}(\beta-\alpha)}$$

Then to convert the angular distance D (in degrees) to linear distance:

D in degrees $\times$ 69·057 = miles

D in degrees $\times$ 111·136 = kilometres

Note it is more convenient to use decimals for the minutes and seconds of degrees.

For v.h.f., where the distances involved over the earth's surface are relatively small compared with the radius of curvature, it suffices to use the normal map of the British Isles, or of Western Europe, which is usually drawn on zenithal equi-distant projection. On such a map both bearing and distance up to 1000 km may be taken as sufficiently accurate for normal amateur requirements.

Further information on map projections may be found in Reference [7].

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H.F. AERIALS

IN setting up a link for radio communication between two stations, certain specific items of equipment must be provided at each end of the circuit. At the sending end there must be a transmitter which imposes the signal intelligence upon a carrier wave at radio frequency and amplifies it to the required power level. At the other end a receiver is required which will again amplify the weak incoming signal, and then decode from it the original intelligence.

The signal passes from one station to the other as a wave propagating in the atmosphere, but in order to achieve this it is necessary to have at the sending end something which will take the power from the transmitter and launch it as a wave, and at the other end extract energy from the wave to feed the receiver. This is an *aerial* and, because the fundamental action of an aerial is reversible, the same aerial suffices at each end. The aerial then is a means of converting power flowing in wires to energy flowing in a wave in space, or is simply considered as a coupling transformer between the wires and free space.

FUNDAMENTAL PROPERTIES

Many of the fundamental properties of aerials are common to their use in any part of the radio frequency spectrum and in free space two aerials which are electrically the same at two widely different frequencies will behave in exactly the same manner. A piece of wire which is one half-wavelength long will, in free space, have the same directional radiation characteristics and appear as exactly the same load to a transmitter whether it be a half-wavelength at 1 Mc/s or at 1000 Mc/s. All that matters is that the wire should maintain a constant relationship between its physical length and the wavelength used. Strictly speaking its diameter should also be scaled with the wavelength but this does not alter the basic principle. The classification into aerials for h.f. and aerials for v.h.f. in this Handbook is one based on the differing practical requirements for aerials for use in the two frequency ranges, taking account of physical size limitations and also the widely differing mechanisms of propagation which dictate different forms of aerial. Much of what is to be said in this chapter about the fundamental characteristics of aerials is equally applicable to both h.f. and v.h.f., and it is only later in the chapter that consideration is given to aerials specifically designed for the h.f. bands. At that point the reader with v.h.f. interests should pick up the story of specialized v.h.f. aerials in Chapter 14—*V.H.F./U.H.F. Aerials*.

Wave Motion

An understanding of aerial behaviour is very closely linked with an understanding of basic propagation, and some aspects cannot be dissociated from it.

In Chapter 12 it is explained that radio waves are propagated as an expanding electro-magnetic wavefront whose intensity decays as it moves further out from the source. The wavefront is formed of electric and magnetic fields which exist at right angles to one another, in the plane of the wavefront, and over small areas may be considered to be made up of parallel lines of electric and magnetic force. The intensity of the wavefront is usually varying sinusoidally with time, as the waves expand like ripples on a pond, and the peak value of the wave decays as the wave moves further and further away from its source. The arrangement of one cycle of the wave along its direction of travel at a particular moment in time, is represented in **Fig. 13.1**, as it appears in general, and also as it appears to the observer whom it is approaching.

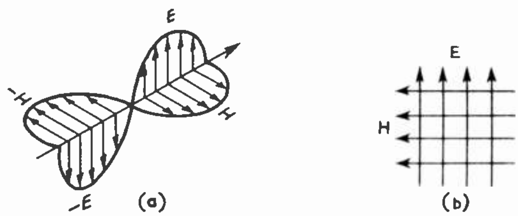


Fig. 13.1. Instantaneous representation of a travelling radio wave: (a) along the path of travel and (b) for the wave approaching the observer.

Some imagination is necessary fully to appreciate the structure of the wave as it propagates through space as a whole. The behaviour of the separate but related electric and magnetic fields is more easily visualized if one imagines them as oscillations travelling along a slack string which is being excited at one end. An observer at a point along the string will see approaching him a succession of crests and troughs which pass him along the string. A complete electro-magnetic wave is then two such strings next to one another but oscillating at right angles and with the crests on each string passing the observer at the same instant.

The distance between successive peaks in the intensity of the electro-magnetic wave as it moves along its direction of propagation is called the *wavelength* and is customarily measured in metres. The wave moves through space with the velocity of light, or more exactly light moves through space with the velocity of electromagnetic waves since light is really e.m. radiation in a narrow band of extremely high frequencies. This velocity is approximately 3×10^8 metres per second (180,000 miles per second), and is known as the *velocity of propagation*. As stated above, from the point of view of a fixed observer, the wave will pass by as a succession of peaks and troughs. The portion of a wave between successive peaks (or troughs) is called one *cycle*, and the number of such *cycles* which pass the observer in one second is known as the *frequency* of the wave, usually measured in cycles per second. At the time of preparation of this edition of the Handbook it had been generally agreed internationally to refer to frequency in Herz, thus commemorating the name of one of the earliest experimenters in the physics of electro-magnetic radiation. The meanings of the abbreviations which will usually be encountered in technical books are given in Table 13.1.

TABLE 13.1

Usual abbreviations for frequency and their relative values. The newly adopted unit of the Herz is shown in brackets.

c/s	(Hz)	cycles per second
kc/s	(kHz)	Kilo-cycles per second = c/s × 10^3
Mc/s	(MHz)	megacycles per second = c/s × 10^6
Gc/s	(GHz)	gigacycles per second = c/s × 10^9

Note: Gc/s is sometimes referred to as kMc/s (kilo-megacycles per second).

Since the wave travels with a known velocity, it is easy to establish a relationship between the frequency and the wavelength, and if f is the frequency in cycles per second and λ (lambda) the wavelength in metres, then

$$f \times \lambda = 3 \times 10^8 \qquad \ldots \text{(i)}$$

For example, a wavelength of 160 metres corresponds to a frequency of $1{\cdot}87 \times 10^6$ c/s or 1·87 Mc/s.

Referring again to the wave illustrated in **Fig. 13.1**, an observer facing the wave would " see " the field as shown in Fig. 13.1(b) which would be alternating at the frequency f. It is this alternating field that excites a receiving aerial into which it delivers some of its power. The arrows indicate the conventional directions of the electric and magnetic fields, E and H, relative to the direction of motion. If the polarity of one component along either E or H in the diagram is reversed, then the wave is receding instead of approaching the observer. The components E and H cannot be separated; together they are the wave and together they represent indestructible energy. They wax and wane together and when the energy disappears momentarily from one point, it must re-appear further along the track; in this way the wave is propagated.

Travelling and Standing Waves

The wave illustrated in Fig. 13.1 is unrestricted in its motion and is called a travelling wave. As long as they are confined to the vicinity of the earth all waves must, however, eventually encounter obstacles. In order to understand what happens in these circumstances, it is convenient to imagine that the obstacle is a very large sheet of metal, since metal is a good conductor of electricity and, as will be seen, an effective barrier to the wave (**Fig. 13.2**). In this case, the electric field is " short-circuited " by the metal and must therefore always be zero at the surface. It manifests itself in the form of a current in the sheet (shown dotted) like the current that is induced in a receiving aerial, and this current re-radiates the wave in the direction from whence it came.

In this way a wave is reflected from the metal but due to the relative directions of field and propagation, the components of forward and reflected waves appear as in **Fig. 13.2(b)**. It will be seen that at the surface of the metal the two electric

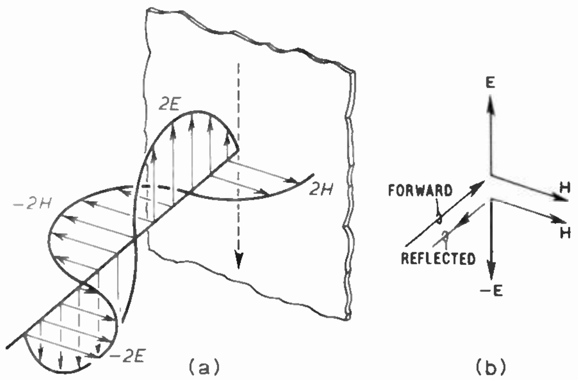


Fig. 13.2. Standing waves due to short circuit reflection at a metal surface, with analysis into forward and reflected wave components.

fields are of opposite polarity and thus cancel each other out, but the magnetic fields are additive.

If the two waves are combined the resultant wave of **Fig. 13.2(a)** is obtained in which the electric and magnetic maxima no longer coincide, but are separated by quarter wave intervals. Such a combined wave does not move, because the energy can alternate in form between electric field in one sector and magnetic field in an adjacent one. This type of wave is called a *standing wave* and exists near reflecting objects comparable in size with the length of the wave. If a small receiving aerial and detector were moved outwards from the front of the reflector the output of the receiver would vary with distance and thus indicate the stationary wave pattern.

In practice radio waves more usually encounter poor conductors, such as buildings or hillsides, and in this case some of the power from the wave is propagated into or through the obstacle, the remainder being reflected. Later in this chapter, aerials will be described which incorporate conductive elements deliberately used as reflectors. It will also be shown that oscillatory currents in aerials or feeders can be reflected to produce standing waves.

A travelling wave moving away into space represents a flow of energy from the source. The transmitter can be regarded as being at the centre of a large sphere, with the radiated power passing out through its surface. The larger the sphere the more thinly will it be spread, so that the further the receiver recedes from the transmitter, the smaller will be the signal extracted by means of a receiving aerial of given size from a given area of the wave front.

Resonant Aerials

If an oscillatory current is passed along a wire, the electric and magnetic fields associated with it can be considered as a wave attached to the wire and travelling along it as far as it continues, and if the wire finally terminates in, say an insulator, the wave cannot proceed but is reflected. This reflection is an open-circuit reflection and the wire carries a standing-wave field complementary to a short-circuit reflection as in Fig. 13.2 with corresponding standing wave voltages and currents. **Fig. 13.3** shows two typical cases where the wire is of such a length that a number of complete cycles of the standing wave can exist along it. Since the end of the wire is open circuit, the current at that point must be zero and the voltage a maximum. Therefore at a point one quarter wavelength from the end, the current must be a maximum and the voltage will be zero. The positions of maxima are usually known as current (or voltage) *anti-nodes* or *loops* and the intermediate positions as *nodes* or *zeros*. At positions of current loops, the current-to-voltage ratio is high and the wire will behave as a low impedance circuit. At voltage loops the condition is reversed and the wire will behave as a high impedance circuit. A wire carrying a standing wave as illustrated in **Fig. 13.3** is an efficient radiator of energy, and can be used as an aerial which exhibits similar properties to a resonant tuned circuit. Such a wire is then known as a *resonant* or *standing wave* aerial. If the wire is not of a resonant length, the standing wave pattern is discontinuous, and the aerial behaves like a de-tuned circuit which requires additional inductance or capacitance to bring it into resonance.

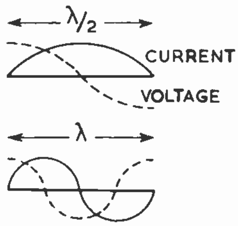


Fig. 13.3. Standing waves on resonant aerials, showing voltage and current variation along the wire. The upper aerial is a half wavelength long and is working at the fundamental frequency; the lower is a full-wave or second harmonic aerial.

The length for true resonance is not quite an exact multiple of the half-wavelength because the effect of radiation causes a slight retardation of the wave on the wire and also because the supporting insulators may introduce a little extra capacitance at the ends. An approximate formula suitable for wire aerials is:

$$Length\ (feet) = \frac{492\ (n - 0{\cdot}05)}{f}$$

where n is the number of complete half-waves in the aerial and f is the frequency in megacycles per second.

Radiation

The actual physics of the radiation of energy from the wire is an involved matter, which for interested readers is better described in detail in textbooks devoted entirely to aerial theory. It suffices to say here that the current flowing up and down the wire gives rise to a magnetic field around the wire, while the charges in motion (which constitute the current) carry with them an electric field. Due to the reversing nature of the current, the two fields are mutually supporting and expand away from the wire, carrying with them energy from the exciting current. There exists in the immediate vicinity of the wire an oscillating field known as the *induction* field (similar to that surrounding an induction coil or a magnet), but this decays in strength rapidly as the distance from the wire increases. At a distance of $\lambda/2\pi$, or approximately one sixth of a wavelength, it is equal in strength to the *radiation* field, but beyond one or two wavelengths has fallen to a negligible level.

Directivity

The radiation field which surrounds the wire is not uniformly strong in all directions. It is strongest in directions at right angles to the current flow in the wire and falls in intensity to zero along the axis of the wire. The wire is then said to exhibit some *directivity* in its radiation pattern, such that the radiated energy is concentrated in some directions at the expense of others. Later in this chapter it will be explained that by using numbers of wires in differing arrangements, even greater directivities can be obtained over that normally exhibited by the single straight wire aerial. Such aerial arrays are called *beams* because they concentrate radiation in the desired direction like a beam of light from a torch. Because a number of wires or elements are needed to create the additional directivity, most beam aerials occupy more space than simple ones, and the extent to which they can usefully be employed is governed largely by their physical dimensions at the wavelength to be used. When space is limited to an average garden, useful directivity can be obtained only from aerials operating on 14 Mc/s or higher bands, the aerial systems for 1·8 Mc/s, 3·5 Mc/s, and 7 Mc/s being generally restricted to simple arrangements involving single wires. In the v.h.f. bands, the wavelength is sufficiently short that the various elements of the aerial may be made rigid and self-supporting and the aerial system may therefore be quite elaborate.

Dipoles

One of the most commonly used words in aerial work is *dipole*. Basically a dipole is simply some device (in the present context an aerial) which has two "poles" or terminals into which radiation currents flow. The two elements may be of any length, and a certain amount of confusion sometimes arises from the failure to state the length involved. In practice it is usually safe to assume that when the word dipole is used by itself, it is intended to describe a half-wavelength aerial, i.e. a radiator of electrical length $\lambda/2$, fed by a balanced connection at the centre. Any reference to gain over a dipole is assumed to refer to this $\lambda/2$ dipole. When reference to another form of dipole is intended, it is usual to state the overall length, e.g. a *wavelength dipole*.

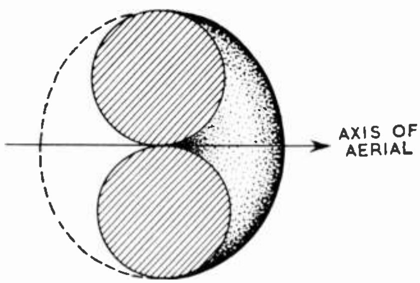
A further reference sometimes encountered is to the *monopole* or *unipole*. This is an unbalanced radiator, fed against an earth plane, and a common example is the ground plane vertical described later in this chapter.

Gain

If one aerial system can be made to concentrate more radiation in a certain direction than another aerial, for the same total power supplied, then it is said to exhibit *gain* over the second aerial, in that direction. In other words, more power would have to be supplied to the reference aerial to give the same radiated signal in the direction under consideration, and hence the better aerial has effectively gained in power over the other. Gain can be expressed either as a ratio of the power required to be supplied to each aerial to give equal signals at a distant point, or as the ratio of the signals received at that point from the two aerials when they are driven with the same power input. Gain is usually expressed in decibels: a table of conversion from voltage and power ratios will be found in the RSGB *Radio Data Reference Book*.

It is important to note that in specifying gain for an aerial, some reference to direction must be included, for no aerial can exhibit gain simultaneously in all directions relative to another aerial. The distribution of radiated energy from an aerial may be likened to a balloon filled with incompressible gas, with the aerial at the centre. The amount of gas represents the power fed to the aerial, and the volume of the balloon can only be increased by putting in more gas. The balloon may be distorted to many shapes, and elongated greatly in some directions so that the amount of gas squeezed in those directions is increased, but this can only be achieved by reducing the amount of gas in some other part of the balloon: the total volume must remain unaltered. Likewise the aerial can only direct extra energy in some required direction at the expense of less in others.

Since gain is expressed in relative terms, it is useful to refer to the gain of any aerial relative to some agreed standard aerial. The performance of any two aerials may then be directly assessed by comparing their relative gains over the standard. Aerial A may exhibit a gain of 3db relative to the reference aerial, and aerial B a gain of 9db. Then aerial B will have a gain relative to A of 9–3 = 6db.



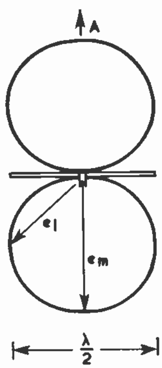


Fig. 13.4. Directive polar diagram of a half-wave dipole, in section and as a solid of revolution around the dipole axis.

It is unfortunate that two standards exist side by side and will be encountered in other references to aerials. The standard often adopted in the USA, and by others, is the theoretical " point source " or *isotropic* radiator, which radiates equal power in all directions, i.e. its solid polar diagram is a sphere. This is a strictly non-practical standard, which cannot be constructed or used, but has the advantage that it does not require a specification of direction in the comparison. The other standard is the "half-wave dipole " which has in its own right a directional pattern as illustrated in Fig. 13.4. This is a practical aerial which can be built and is therefore a more realistic basis for comparison, but it should be noted that gain expressed *relative to a $\lambda/2$ dipole* means by inference relative to the maximum radiation from the dipole, i.e. in direction A in Fig. 13.4.

It will be found in practice that gain is frequently expressed in db without reference to the standard employed. This can lead to a dispute of 2·15db in claimed results, being the relative gain of the two standards employed (the gain of a $\lambda/2$ dipole relative to the isotropic source). In such cases it is safer to assume the more conservative figure when comparing different aerial performance unless one is sure that the same standard aerial has been assumed in each case.

Because direction is inevitably associated with a statement of gain, it is usually assumed, in the absence of any qualifying statement that the gain quoted for any aerial is its gain in the direction of its own maximum radiation. Where the aerial system can be rotated, as is the case on 14 Mc/s and higher, this is not so important, but when the aerial is fixed in position, the superiority it exhibits in one direction over another aerial will not hold in other directions because of the different shapes of the two aerial patterns. Aerial A may have a quoted gain of 6db over aerial B, but only in the directions which favour the shape of its radiation pattern, relative to aerial B.

Radiation Resistance and Aerial Impedance

When power is delivered from the transmitter into the aerial, some small part will be lost as heat, since the material of which the aerial is made will have a finite resistance, albeit small, and a current flowing in it will dissipate some power. The bulk of the power will be radiated and, since power can only be lost in a resistance, it is convenient to consider the radiated power as dissipated in a fictitious resistance which is called the *radiation resistance* of the aerial. Using ordinary circuit relations, if a current I is flowing into the radiation resistance R, then a power of I^2R watts is being radiated. It was shown in Fig. 13.3 and mentioned in connection with standing waves, that the r.m.s. current distribution along a resonant aerial or indeed any standing wave aerial is not uniform but is approximately sinusoidal. It is therefore necessary to specify the point of reference for the current when formulating the value of the radiation resistance, and it is usual to assume the value of current at the anti-node or maximum point. This is known as the current loop, and hence the value of R given by this current is known as the *loop radiation resistance:* in practice the word loop is omitted but inferred.

A half-wave dipole has a radiation resistance of about 70 ohms. If it is made of highly conductive material such as copper or aluminium, the loss resistance may be less than one ohm. The conductor loss is thus relatively small and the aerial provides an efficient coupling between the transmitter and free space.

When the aerial is not a resonant length, it behaves like a resistance in series with a positive (inductive) or negative (capacitive) reactance and requires the addition of an equal but opposing reactance to bring it to resonance, so that it may be effectively supplied with power by the transmitter. The combination of resistance and reactance, which would be measured at the aerial terminals with an impedance meter, is referred to in general terms as the aerial *input impedance*. This impedance is only a pure resistance when the aerial is at one of its resonant lengths.

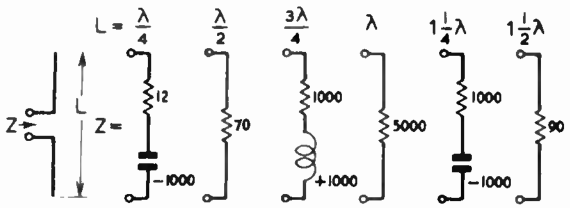


Fig. 13.5. Typical input impedance (Z_0) value of dipoles of various lengths. The values for $L = \lambda/2$ and $3\lambda/2$ are always approximately as shown but values for other lengths vary considerably according to the length/diameter ratio of the aerial. The values given are typical for wire h.f. aerials. Note how the reactance changes sign for each multiple of a quarter wavelength.

Fig. 13.5 shows, by means of equivalent circuits, how the impedance of a dipole varies according to the length in wavelengths. It will be seen that the components of impedance vary over a wide range.

The input impedance of the aerial is related specifically to the input terminals, whereas the radiation resistance is related to the current at its loop position. It is possible to feed power into an aerial at any point along its length, so that the input impedance and radiation resistance of even a resonant aerial may be very different in value although both are pure resistances. Only when the feed point of the aerial coincides with the position of the current loop on a single wire, will the two be approximately equal (**Fig. 13.6(a)**). If the feed point occurs at a position of current minimum and voltage maximum, the input impedance will be very high, but the radiation resistance remains unaltered (**Fig. 13.6(b)**).

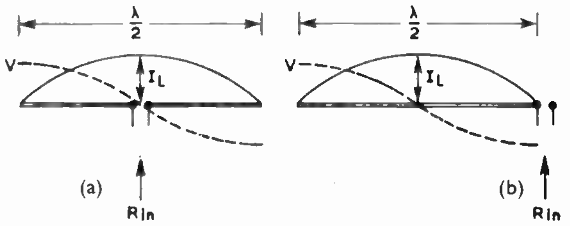


Fig. 13.6. The input resistance of a half-wave dipole is low at the centre and high at the end, although the loop radiation resistance is the same in each case.

For a given power fed into the aerial, the actual feed point current, measured on an r.f. ammeter will be very low, but because the input impedance is high, the power delivered to the aerial is the same. Such an aerial is described as *voltage fed*, because the feed point coincides with a point of maximum voltage in the distribution along the aerial. Conversely an aerial fed at the low impedance point of the current maximum is described as *current fed*.

The input impedance of a current fed half-wave dipole is approximately equal to 70 ohms (the theoretically perfect radiation resistance is 73·14 ohms), and will be much the same value irrespective of the size of wire or rod used to fabricate the dipole. The input impedance of a voltage fed half-wave dipole is very high, and its precise value depends not only upon the loop radiation resistance of the dipole, which is independent of the method of feed, but also the physical size of the wire used. The dipole wire between the current loop and the current zero at the ends may be considered to act as a quarter-wavelength transformer between the loop radiation resistance at the centre, and the input impedance at the end (**Fig. 13.7**). As a transformer, the wire must exhibit a certain characteristic impedance Z_0, and the transformer ratio will depend upon the value of Z_0, which in turn depends upon the ratio of the conductor diameter to its length. If the input impedance is R_i, and the radiation resistance R_r, then approximately $R_i = \frac{Z_0^2}{R_r}$: the value of R_r is fixed by the current distribution, and hence the value of R_i will change as Z_0, or the wire length/diameter ratio changes. Typical values of R_i for different ratios are given in **Table 13.8**; the behaviour of transformers is dealt with in more detail in the section on Transmission Lines.

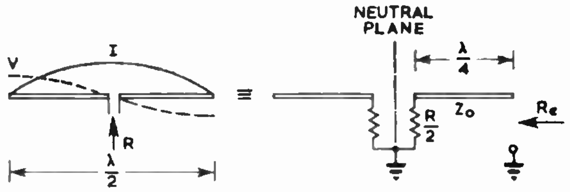


Fig. 13.7. The end impedance of a half-wave dipole can be derived by considering one half of the centre impedance transformed along a quarter-wavelength of transmission line formed by the dipole wire.

In certain cases, even though a half-wave dipole is centre fed at the current loop position, the input impedance will not equal the radiation resistance. If the halfwave dipole is folded (see page 13.22), a transformation again occurs, to produce a different input impedance. If the dipole is the driven element of a multi-element array such as a Yagi, then the input impedance will also be modified (usually reduced) because of the presence of the other elements. When an aerial of fixed physical length is used at a number of different frequencies, it behaves exactly the same as the dipole illustrated in Fig. 13.6. If a wire is made up to be 33 ft. long and centre-fed, it would be resonant ($\lambda/2$) at 14 Mc/s, three quarter wavelength ($3\lambda/4$) at 21 Mc/s and full wave (λ) at 28 Mc/s. At 14 Mc/s it is easy to supply power into its 70 ohm input resistance, but at full wave resonance at 28 Mc/s the high value of 5000 ohms presents practical difficulties. Again at 21 Mc/s where the impedance is complex and evidently high, special arrangements are necessary. These problems are discussed later in connection with tuned feeders and aerial couplers and in relation to specific aerials.

This same aerial on 7 Mc/s would only be $\lambda/4$ long, with a low resistance of 12 ohms and a high capacitive reactance. It would require a *loading coil* of 23 μH inductance (1000 ohms reactance) for resonance at 7 Mc/s. Unless this loading coil were physically large it might well have appreciable loss resistance (6 ohms for example). Thus only two thirds of the power from the transmitter would reach the aerial, the remaining third being expended as heat in the coil, and the efficiency would be only 67 per cent. With 18 ohms total

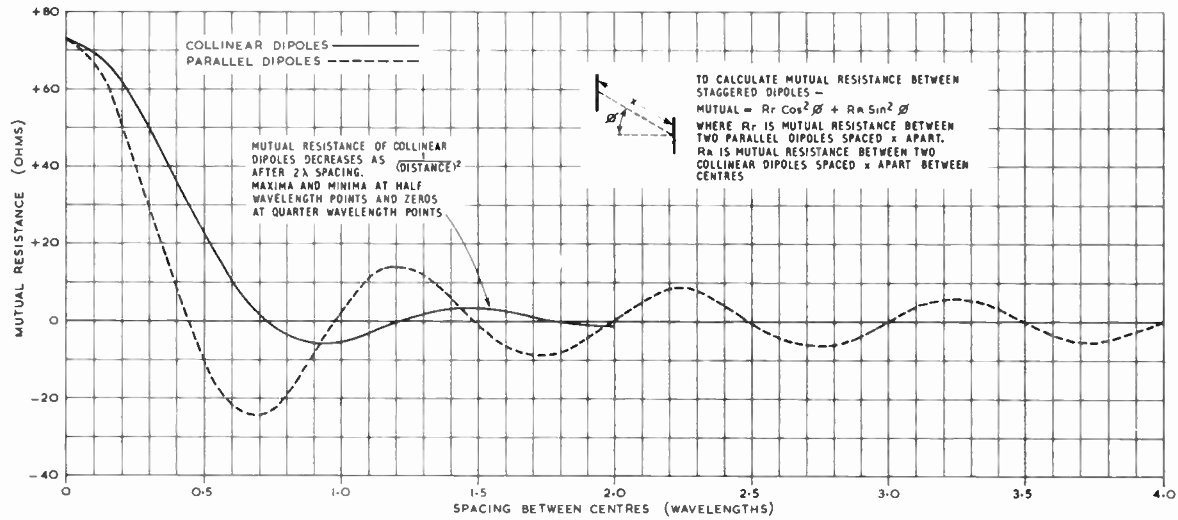


Fig. 13.8. Variation of mutual resistance between parallel and collinear half-wave dipoles.

input resistance and a reactance of 1000 ohms, the Q of the circuit (ratio of reactance to resistance) would be approximately 50, which means that tuning adjustments for resonance would be critical. On 3·5 Mc/s the same aerial would have a resistance of only 3 ohms and a reactance of 2000 ohms. It would therefore be very difficult to use it efficiently as it would need re-tuning for different parts of the 3·5 Mc/s band.

Mutual Impedance

If two aerials are in close proximity, they couple together, and the input impedance of each is modified by the presence of the other. The amount by which the input impedance of the one is varied is called its *mutual impedance* due to the other; the mutual impedance depends upon the degree of coupling and therefore the physical spacing of the aerials. For simple aerials the variation follows a regular pattern, and **Fig. 13.8** illustrates the mutual resistance between parallel and collinear half-wave dipoles for variable spacing. Beyond two wavelengths for collinear dipoles (four wavelengths for parallel dipoles) the mutual resistance decays to negligible proportions; up to those distances, it takes the form of a decaying sinusoidal wave, with zeroes occurring approximately at $0{\cdot}5\lambda$ intervals. The values shown on the vertical axis should be added to the perfect free-space input resistance of 73 ohms for each dipole to obtain the actual input resistance of each. For dipoles spaced $1{\cdot}25\lambda$ apart, the input resistance of each will be $73 + 13{\cdot}9 = 86{\cdot}9$ ohms. The reactive term of the mutual impedance varies in a similar cyclic manner.

Stacking Gain

It is possible to use a number of discrete radiators in parallel to achieve an effective gain over a single one. Examples of this are the collinear and broadside arrays made up of a number of discrete $\lambda/2$ dipoles. At first glance, it may seem unreasonable to expect gain, since the splitting of the power available between a number of similar aerials would appear to be no different to the feeding of the whole power into one of their number. That gain is achievable depends upon the fact that the distant field, which excites the receiving aerial, is directly related to the radiation current, whereas the power fed to the aerial is proportional to the square of the radiation current. **Fig. 13.9** shows two half-wave dipoles stacked one wavelength apart, and each with a radiation resistance R. They are centre fed so that the radiation resistance is nominally the same as the input impedance, and each carries a loop radiation current I. For one dipole, the distant field is E due to a radiation current $2I$, and the power required to be fed to the dipole to achieve this is $(2I)^2 \times R = 4I^2R$. If the second dipole is now introduced and the current is split between them so that each has a current I and gives a distant field $E/2$, the fields will be in phase broadside to the dipoles and will add to give the same field strength E as before, but the power delivered to each

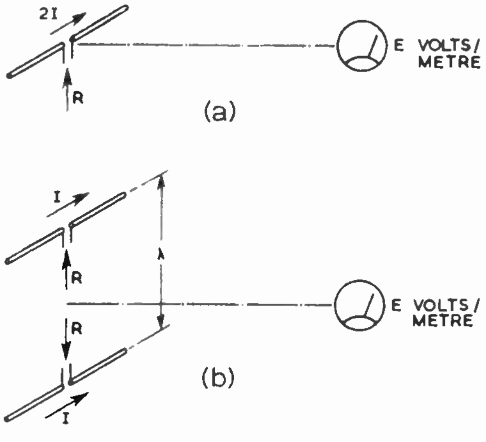


Fig. 13.9. Stacking two dipoles gives an effective gain of 3db (ignoring mutual effects) because the input power is proportional to the square of the current and the distant field only to the current itself.

dipole is now I^2R, so the total power is $2I^2R$. For the same distant field, the power required is only half that of the single dipole, so a power gain of 3db has resulted from stacking the two dipoles. What has happened is that the fields radiated by the individual dipoles have added in phase in the desired directions but have cancelled in others by being out of phase, which results in a sharpening of the radiation pattern in the desired direction, and hence gain.

This simple explanation of stacking gain assumes that the input impedance of each dipole remains unaltered when the two are used together. This is not exactly the case in practice since mutual impedance plays a part. It was shown earlier that the cyclically varying mutual resistance must be added to the basic input resistance of each dipole to determine the actual value obtained due to the coupling. Dependent upon the spacing involved, the input resistance of each dipole will be sometimes greater, sometimes smaller than the single dipole value, and hence the value of R used to determine the total power delivered to the two dipoles will be modified. The smaller its value, the less will be the power required to achieve the standard radiation current, and hence the greater the gain. The greatest gain will occur when $R + R_M$ has its smallest value for each dipole, i.e. at that spacing which gives the largest *negative* mutual resistance, or 0·7λ for two dipoles.

In the simple case, the gain was given by:

$$G = \frac{4RI^2}{2RI^2} = 2$$

Taking into account the effect of mutual impedance, the actual gain at 0·7λ spacing is given by:

$$G = \frac{4I^2R}{2 \times I^2R(1 \cdot 0.35)} = \frac{2}{0 \cdot 65} = 3 \cdot 06, \text{ or } 4 \cdot 88\text{db.}$$

This is the greatest gain which can ever be obtained from two λ/2 dipoles, and the distance between them to achieve this gain is known as the *optimum stacking distance*. In the example of the two dipoles spaced one wavelength apart, the mutual resistance is very nearly zero, and the simple figure of 3db is correct.

The principle of stacking gain may be extended to more than two radiators, and to radiators each of which is highly directive in its own right, but wherever many elements are involved, consideration must be given to the mutual coupling between each element and all other elements, in turn. The derivation of the optimum stacking distance for such systems becomes very complicated, but it is true to say that the higher the directivity and intrinsic gain of each unit in the array, or the larger the number of elements, the greater is the optimum stacking distance between them. Two dipoles require a spacing of 0·65λ, four are better spaced at 0·8λ intervals, eight at 0·9λ intervals and so on. Where the basic unit is of high gain, such as a Yagi array (see page 13.64), the optimum stacking distance lies between 1·0λ and 2·0λ, and depends upon the performance of the individual Yagis. This aspect is covered more fully in Chapter 14—*V.H.F./U.H.F. Aerials*.

TRANSMISSION LINES

Three separate parts are involved in an aerial system: the radiator, the feed line between transmitter and radiator, and the coupling arrangements to the transmitter. Wherever possible, the aerial itself should be placed in the most advantageous position, and a feed line used to connect it to the transmitter or receiver with a minimum of loss due to resistance or radiation. In some circumstances (when space is limited or when multi-band operation is required) the feed line is omitted and the end of the aerial is brought into the station and connected directly to the apparatus.

By the use of transmission lines or *feeders*, the power of the transmitter can be carried appreciable distances without much loss due to conductor resistance, insulator losses or radiation. It is thus possible to place the aerial in an advantageous position without having to suffer the effects of radiation from the connecting wires. For example, a 14 Mc/s dipole only 35 ft. long could be raised 60 ft. high and fed with power without incurring appreciable loss. If, on the other hand, the aerial wire itself were brought down from this height to the transmitter, most of the radiation would be propagated from the down-lead in a high angular direction. An arrangement of this nature would be most unsuitable for long distance communication.

Types of Line

There are three main types of transmission line:

(i) The single wire feed arranged so that there is a true travelling wave on it.

(ii) The concentric line in which the outer conductor (or sheath) encloses the wave (*coaxial feeder*).

(iii) The parallel wire line with two conductors carrying equal but oppositely directed currents and voltages, i.e. balanced with respect to earth (*twin line*).

A single wire line seldom works correctly because it is very difficult to prevent it acting as a radiator instead of a pure transmission line. In the other two types, the field is confined to the immediate vicinity of the conductors and does not radiate if proper precautions are taken. Commercially available types of feeder are dealt with in Chapter 14 (*V.H.F./U.H.F. Aerials*) with some emphasis on their v.h.f. applications. In this chapter the discussion is confined to the mode of working and application to frequencies below 30 Mc/s but the fundamental principles are the same.

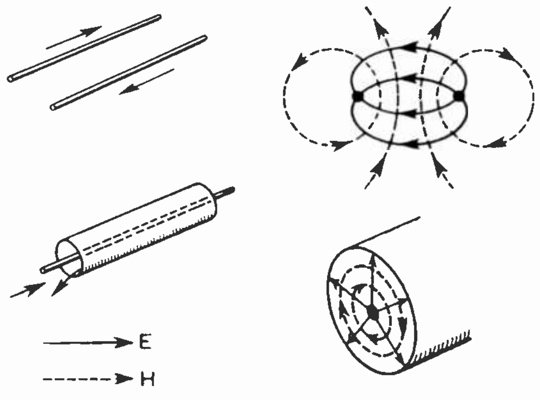


Fig. 13.10. Concentric and two-wire transmission lines, with cross-sections of their wave fields. The field directions correspond to waves entering the page.

The wave which travels on a transmission line is fundamentally the same as the free-space wave of Fig. 13.1 but in this case it is confined to the conductors and the field is curved about the conductors instead of being linear as in that diagram. Concentric and two-wire lines and the fields carried on them are illustrated in **Fig. 13.10.** In the concentric line the current passes along the centre conductor and returns along the inside of the sheath. At high frequencies the currents do not penetrate more than a few thousandths of an inch into the metal, due to skin effect; hence with any practical thickness of the sheath there is no current on the outside. The fields are thus held inside the cable and cannot radiate.

In the twin line the two wires carry " forward and return " currents and the field is concentrated in their vicinity. When the spacing between wires is a very small fraction of the wavelength, the radiation is negligible provided the line is balanced i.e. the currents are equal and opposite in the two parallel wires. In the h.f. range, spacings of several inches may be employed, but in the v.h.f. range small spacing is important.

Characteristic Impedance

In the earlier section on the behaviour of currents on radiating wires, it was stated that a travelling wave is one which moves along a wire in a certain direction without suffering any reflection at a discontinuity. The same applies to transmission lines, although in their case the presence of reflections and therefore standing waves does not cause radiation if the line remains balanced or shielded.

It is possible to terminate any transmission line with a pure resistance of such a value that no reflection of the voltage and current travelling waves occurs and all the transmitted power is dissipated in the resistance. The value of this resistance is specific to any particular type of line, depending upon the physical size and spacing of the conductors used, and it is therefore possible to attribute to a line a property known as its *characteristic impedance*, usually denoted by Z_0, which is equal in value to this resistance. When a line is terminated in a resistance equal to its characteristic impedance, there is no reflection of power, and the line is said to be *matched.*

Because no reflections occur at the end of a matched line, the situation is the same as for the case where the line is infinitely long but unterminated, for the voltage and current waves continue to travel forward without suffering reflection at any point. Arising from this analogy, another definition of characteristic impedance sometimes employed states it to be the input impedance of an infinitely long section of line of the particular dimensions involved.

Again because no reflections occur at the end of a correctly matched line, the ratio of the travelling waves of voltage and current, $V/I = Z_0$. This is the important property which enables the load presented by the aerial to be in turn presented to the transmitter, without any loss in the process. This is true irrespective of the length of line employed, and consequently the value of the characteristic impedance Z_0 is also independent of the length of the line. In order to achieve maximum efficiency from a transmission line, it should always be operated as close to a matched condition as possible, i.e. the load presented by the aerial should be arranged, either directly or by means of some impedance transformer, to present a good match to the line. The degree

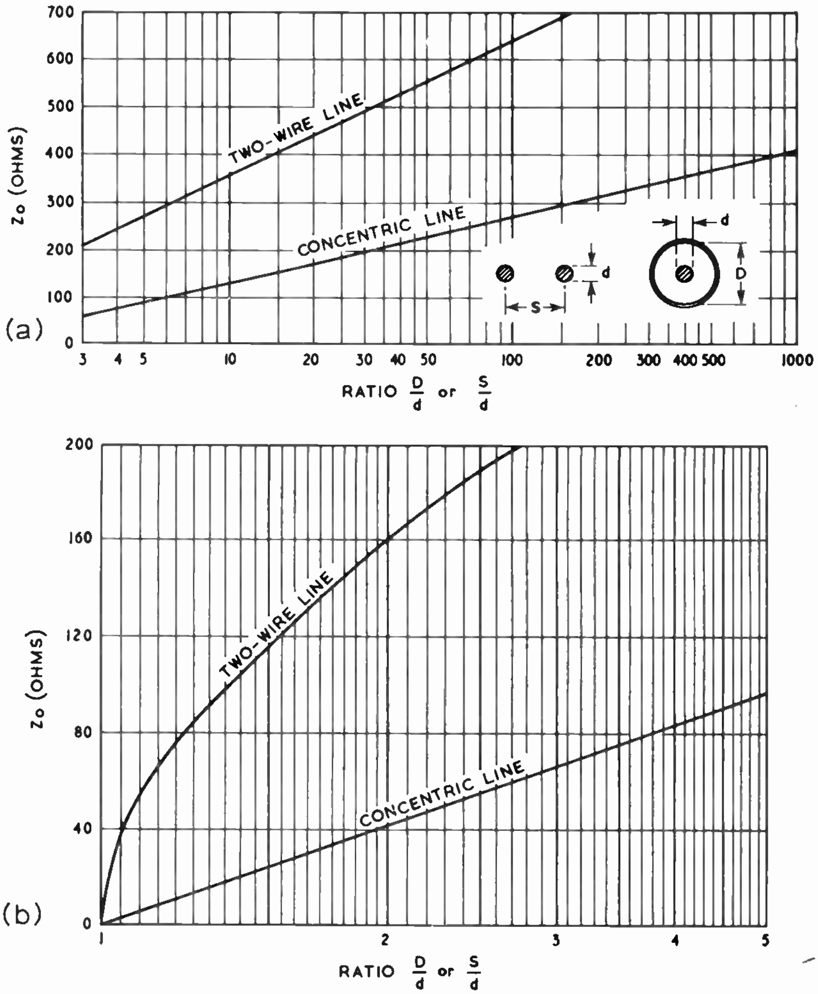


Fig. 13.11. Chart giving characteristic impedances of concentric and two-wire lines in terms of their dimensional ratios, assuming air insulation. When the space around the wires is filled with insulation, the impedance given by the chart must be divided by the square-root of its dielectric constant (permittivity). The ratio of this reduction is called the velocity factor, because the wave velocity is reduced in the same proportion.

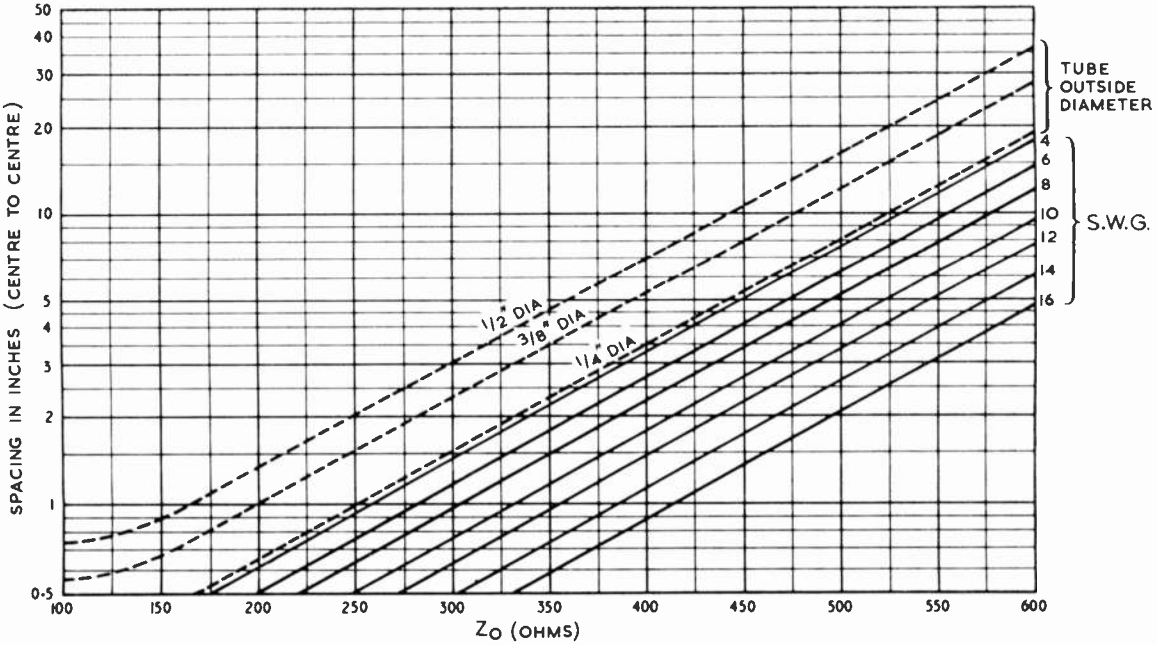


Fig. 13.12. Characteristic impedance of balanced lines for different wire and tube sizes and spacings, for the range 200-600 ohms. The curves for tubes are extended down to 100 ohms to cover the design of Q-bar transformers.

to which the load impedance can be permitted to depart from the characteristic impedance is discussed in the section on attenuation.

The characteristic impedance is determined purely by the dimensional ratios of the cross-section of the line, and not by its absolute size. A concentric line or cable with diameter ratio 2·3 : 1 and air dielectric, is always a 50 ohm line, whatever its actual diameter may be. If it is connected to an aerial of 50 ohms radiation resistance, then all the power in the line will pass into the aerial and the impedance at the sending end of the line will also appear to be 50 ohms.

Coaxial cables can conveniently be constructed with characteristic impedance values between about 50 and 120 ohms. Twin lines have higher impedances; in practice, between 80 and 600 ohms. **Fig. 13.11** and **Fig. 13.12** show the characteristic impedance of coaxial and two-wire lines in terms of the dimensional ratios, assuming air between the conductors. The formula for concentric lines of inner and outer diameters d and D respectively is:

$$Z_0 \text{ (ohms)} = 138 \, Log \, (D/d)$$

Thus if the diameter ratio D/d is 2·3 : 1 the logarithm of 2·3 is 0·362 and this, multiplied by the constant 138 gives Z_0 = 50 ohms. If the space between conductors is filled with insulating material with a *dielectric constant* ϵ (permittivity) greater than unity, the above value of Z_0 must be divided by the square root of the dielectric constant.

The usual material for insulation is polythene, which has a permittivity of 2·25. The square root of 2·25 is 1·5 so that a " solid " polythene cable has a characteristic impedance two-thirds of the value given by the formula. Many cables have a mixed air/polythene dielectric, and for these it is necessary to estimate the effect of the dielectric; this can be carried out by measuring the velocity factor as described later in this chapter.

The characteristic impedance of two-wire lines of wire diameter d and centre spacing S is given by the approximate formula

$$Z_0 \text{ (ohms)} = 276 \, Log_{10} \, (2S/d).$$

This formula applies to the straight parts of the lines in Fig. 13.12, but does not bring the impedance to zero when the conductors touch. A more precise formula is therefore used for small spacings.

$$Z_0\text{(ohms)} = 276 \cosh^{-1} \left(\frac{s}{d}\right)$$

As with concentric lines, an allowance must be made for the effect of the insulation. A thin coat of enamel or even a p.v.c. covering will not produce any material change in lines with an impedance of 300 ohms or less, but in the moulded feeder commonly known as 80 ohm flat twin the electric field between the wires is substantially enclosed in the polythene insulator, which is thus effective in reducing the impedance to two-thirds of the value given by the formula.

When an open-wire line is constructed with wooden or polythene spacers it is again difficult to estimate the effect of the insulation, but since the proportion of insulator to air is relatively small, the effect is also small. A 600 ohm line with spreaders spaced every few feet along it would normally be designed as if it were for about 625 ohms, e.g. for 16 s.w.g. wires (0·064 in. diameter) the spacing would be made 6 in. instead of 5 in. as given by the chart for 600 ohms.

A transmission line can be considered as a long ladder network of series inductances and shunt capacitances, corresponding to the inductance of the wires and the capacitance between them. It differs from conventional L/C circuits in that these properties are uniformly distributed along the line, though applications are given later where short sections of line are used instead of coil or capacitor elements. If the inductance and capacitance per unit length, say, per metre, are known, the characteristic impedance is given by:

$$Z_0 = \sqrt{L/C} \text{ (ohms)}$$

and the velocity of waves (v) on the line by:

$$v = 1 \div \sqrt{LC} \text{ metres per second}$$

The inductance and capacitance can both be determined from purely geometrical calculations on the shape of the cross-section.

For the two wire parallel line, they are obtained from the expressions:

$$L = 0{\cdot}921 \log_{10}\frac{2S}{d} \text{ microhenries/metre}$$

$$C = \frac{12{\cdot}05\epsilon}{\log_{10}\frac{2S}{d}} \text{ picofarads/metre}$$

where ϵ is the dielectric constant of the material in the space between the conductors (= 1·0 for air).

Similarly for concentric lines:

$$L = 0{\cdot}46 \log_{10}\frac{D}{d} \text{ microhenries/metre}$$

$$C = \frac{24{\cdot}1\epsilon}{\log_{10}\frac{D}{d}} \text{ picofarads/metre.}$$

Velocity Factor

When the medium between the conductors of a transmission line is air, the travelling waves will propagate along it at the same speed as waves in free space. If a dielectric material is introduced between the conductors, for insulation or support purposes, the waves will be slowed down and will no longer travel at the free space velocity. The velocity of the waves along any line is equal to $1 \div \sqrt{LC}$; from the expressions for L and C given above, the value of C depends upon the dielectric constant of the insulating material. The introduction of such material increases the capacity without increasing the inductance, and consequently the velocity is reduced, by the same factor $1 \div \sqrt{\epsilon}$ by which the characteristic impedance is also reduced. The ratio of the velocity of waves on the line to the velocity in free space is known as the *velocity factor*. It is as low as 0·5 for mineral or p.v.c. insulated lines and is equal to 0·66 for solid polythene cables ($\epsilon = 2{\cdot}25$). Semi-airspaced lines have a factor which varies between 0·8 and 0·95, while open wire lines with interval spacers may reach 0·98.

It is important to make proper allowance for this factor in some feeder applications, particularly where the feeders are used as tuning elements or interconnecting lines on aerial arrays, or as TVI chokes. For example, if $v = \frac{2}{3}$, then a quarter-wave line would be physically $\frac{1}{6}$ wavelength long ($\frac{2}{3} \times \frac{1}{4} = \frac{1}{6}$).

In practice, the velocity factor v can be found by short-circuiting a length of cable with about 1 in. of wire formed into a loop and then coupling the loop to a grid dip oscillator. The *lowest* frequency at which the cable shows resonance corresponds to an *electrical* length of one quarter-wave: then

$$v = \frac{f(\text{Mc/s}) \times \textit{Length}\text{ (feet)}}{246}$$

and should have a value between 0·5 and unity.

A particularly useful application of the velocity factor of solid polythene cables is to be found in the use of such cables to couple together dipoles in a stacked aerial array. It was shown earlier that the optimum stacking distances for parallel dipoles is approximately 0·7λ for maximum broadside gain. To achieve this, all dipoles must be driven in phase, which necessitates cables which are electrically multiples of one half-wavelength long. If the harness is constructed from solid polythene cables one wavelength long electrically, their physical length is 0·66λ, which is near enough to the optimum distance between the dipoles to permit the cables to be extended straight along the supporting boom, without any slack length to be secured.

Standing Waves

It has already been stated that when a transmission line is terminated by a resistance equal in value to its characteristic impedance, there is no reflection and the line carries a pure travelling wave. When the line is not correctly terminated, the voltage-to-current ratio is not the same for the load as for the line and the power fed along the line cannot all be absorbed—some of it is reflected in the form of a second travelling wave, which must return along the line. These two waves, forward and reflected, interact all along the line to set up a *standing wave*.

The flow of power along the line can be interpreted as the progress along the line of a voltage wave and a current wave which are in phase, the product of which is the value of the power flowing.

If the voltage V at a point on the line is given by the expression:

$$V = V_0 \cos \omega t$$

and the current at the same point by:

$$I = I_0 \cos \omega t$$

then the amount of power flowing is the product of the r.m.s. voltage and current:

$$P = 0{\cdot}707\ V_0 \times 0{\cdot}707\ I_0 = 0{\cdot}5\ V_0 I_0$$

and is independent of the time or position on the line, varying only as the peak amplitude of the voltage or current wave is altered. Since the voltage and current waves can be expressed in identical terms, it is convenient to consider the current

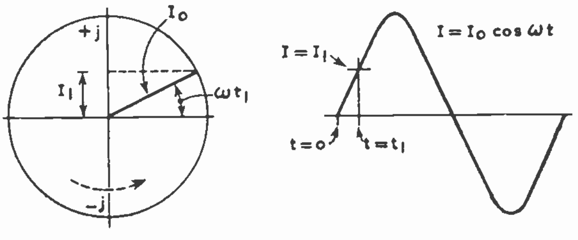


Fig. 13.31. Graphical interpretation of a sinusoidal wave represented by a vector.

wave, bearing in mind that the discussion is equally applicable to the voltage wave.

The current wave is one of a sinusoidal form, varying in amplitude at an angular rate $\omega = 2\pi f$ where f is the frequency at which the transmitter is generating the r.f. power. Such a wave may be represented graphically by a vector of constant magnitude (or length) rotating at an angular speed ω. The actual instantaneous value of the wave at any moment is obtained by projecting the length of the vector on to a line passing through the origin (Fig. 13.13). The nature of vectors is fully covered in any textbook dealing with a.c. waves.

Thus the distribution of current along the line *at a particular moment of time* due to the passage of the current wave, is also sinusoidal, with the phase of the current lagging constantly along the length of the line. It can therefore be represented by a whole series of vectors, each appropriate to a particular physical point on the line. This illustrates the physical meaning of the statement that a piece of line is

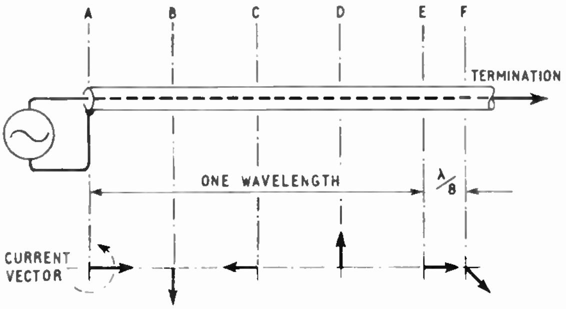


Fig. 13.14. Initial variation of incident current vector, according to position along the line at any instant of time.

" one wavelength " long, since this is the distance between adjacent points along the line at which the current is equal in amplitude and phase, i.e. the vectors are identical (**Fig. 13.14**).

The vectors are drawn for a perfect loss-less line, in which the current wave suffers no attenuation during its passage along the line. It is convenient to explain the principles of generation of standing waves by reference to the behaviour of such a line and subsequently to examine what happens when the line exhibits some amount of attenuation, as is the case in practice to a greater or lesser degree.

The current wave moving along the line will eventually reach the far end, and will then be influenced by the termination it meets. In order to establish the magnitude of this effect we must now look at the various loads that can exist. These may be divided broadly into three groups depending upon whether all, some, or none of the incident power is reflected:

(i) A resistive termination equal in value to the characteristic impedance of the transmission line. By definition there will be no reflection from such a termination, and all the current will flow into this load. Hence all the power delivered by the generator will be dissipated in this load (**Fig. 13.15(a)**).

(ii) A resistive termination of value other than in (i) above. This may or may not have a reactance associated with it. At such a termination, a reflection of the current wave will occur, the actual amount of the reflected current being dependent upon the relative values of the resistive part of the load and the characteristic impedance of the line. In general, the greater the difference between these two, the larger the proportion of current reflected. Also, if the load resistance is greater than the characteristic impedance, a phase reversal occurs, i.e. the reflected current wave is 180 out of phase with the incident wave. In the most general case where a reactance is also involved, a further phase shift will occur, the amount depending upon the ratio of load resistance to reactance. Thus some of the generator power is dis-

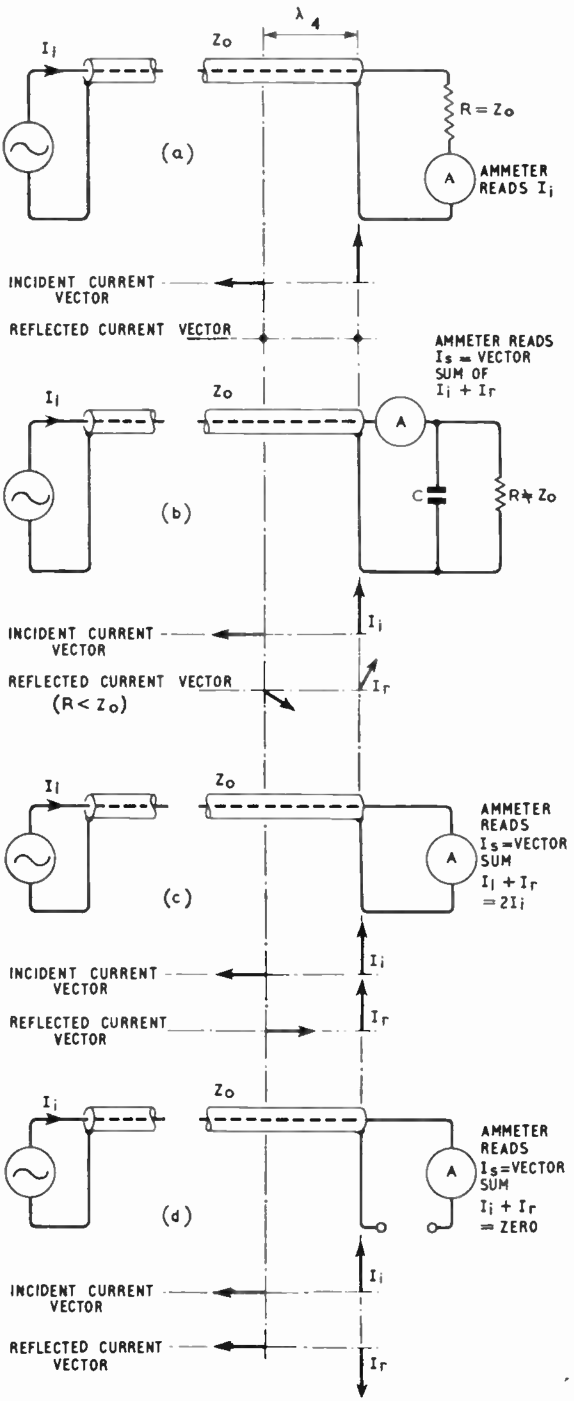


Fig. 13.15 (a) Arrangement of incident and reflected current vectors in the vicinity of a correctly terminated load. (b) Arrangement of incident and reflected current vectors in the vicinity of a mismatch load of complex impedance. (c) Arrangement of incident and reflected current vectors in the vicinity of a short-circuit. (d) Arrangement of incident and reflected current vectors in the vicinity of an open circuit.

sipated in the resistive part of the load and some is reflected back along the line **(Fig. 13.15(b))**.

(iii) An entirely reactive termination. This includes both the limiting cases of a short circuit and an open circuit. Since there is no resistive component in the load, no power can be dissipated, and must, therefore, be all reflected. This means that the whole of the incident current wave is reflected back down the line. There will be a phase change relative to the incident wave, the actual value of which will depend upon the reactance of the load.

In the limiting case of a short circuit, the current in the short circuit will have a maximum value and, therefore, the phase change is zero; in the case of an open circuit, no current can flow in the load, and hence the phase change is 180°, this being a special case of the more general condition covered in (ii).

In every case, apart from that of a perfect termination, some proportion of the incident current is reflected at the far end of the transmission line, and commences to flow back along the line at the same rate as the incident current flowing towards the end. This reflected current wave will, depending upon the circumstances of the termination, commence with an amplitude and phase both differing from the incident current wave. However, since the reflected wave is travelling back along the line, its value at any one moment in time may also be expressed in terms of a reflected current vector, which is rotating in a clockwise direction, opposite to the incident current vector (since the waves are travelling in opposite directions) but nevertheless rotating at the same angular rate, since the frequency of the wave remains unaltered. This is illustrated in Fig. 13.15(a)-(b) where the incident and reflected current wave vectors are shown at a precise moment of time for both the end of the line and a point one quarter wavelength away from the end.

The reflected wave of current will now travel back along the line, towards the generator supplying the power. At any physical point on the line, the net current at any instant is merely the vector sum of the two waves, and the resultant vector gives the amplitude and phase.

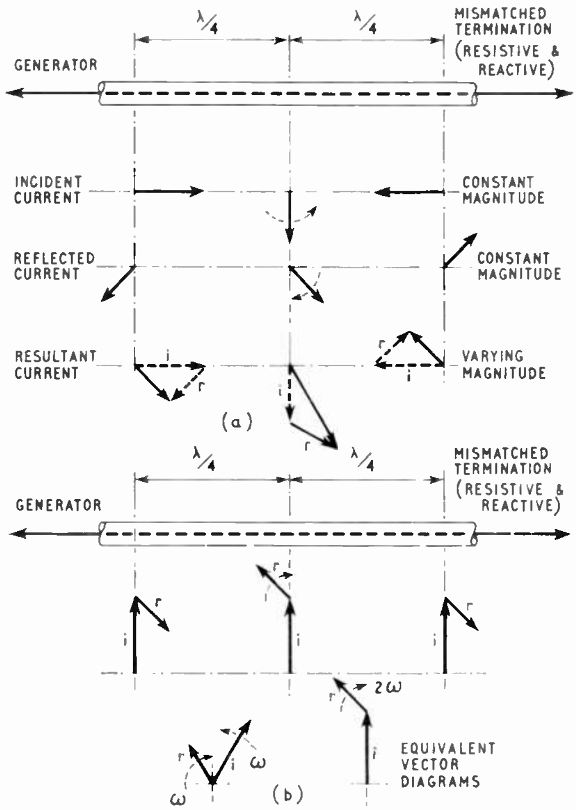


Fig. 13.16. (a) Combination of incident and reflected currents along a typical section of the line. (b) Effect on (a) of maintaining the incident current vector stationary. The reflected current vector effectively rotates at twice the speed, to produce maxima and minima of current which have a crest-to-crest distance of half a wavelength.

Although the explanation has so far been based on the conditions existing on the line at one instant of time, all the currents shown in **Fig. 13.16** are varying sinusoidally with time at the same rate, and the vector relationships of magnitude and phase correctly represent the behaviour of the r.m.s. value of the currents at each point, irrespective of time.

In order to derive the curve showing the variation of amplitude of the standing wave along the line, which is the aspect usually of interest, it is sufficient to consider only the resultant of the reflected current vector rotating at a constant rate equal to twice the generator frequency, about a fixed incident current vector. The doubling of the rotation rate arises from the contra-rotation of the two vectors, each at carrier frequency, with varying position along the line.

This can now be translated back to the equivalent wave, which exists *in space* along the transmission line **Fig. 13.17** and gives rise to the familiar standing wave pattern, having a maximum value equal to the sum of the incident and reflected wave vectors, and a minimum value equal to the difference. The variations with time have been eliminated by maintaining a constant incident current vector.

The shape of the standing wave curve is not a pure sine-wave, being generated by the resultant of two vectors. The shape is approximately sinusoidal for low values of s.w.r., but departs increasingly from this form as the s.w.r. increases, to an extent where in the limit of an open or short circuit termination it is close to half sine-waves **(Fig.13.18(a)** and **(b))**. Because of the fact that in space the reflected current wave vector is effectively rotating at twice the generator frequency, the positions of successive maxima and minima occur at quarter-wavelength intervals along the line.

The distance from crest to crest of the standing wave is **one half-wavelength** at carrier frequency, an important distinction from the travelling wave, which by definition has a crest to crest distance of one wavelength. In general, the conditions of voltage, current and therefore impedance repeat themselves every half-wavelength along the line, and use is made of this property when impedance matching is being considered

The *standing wave ratio* (s.w.r.) k is the ratio of the maximum and minimum value of the standing or space wave existing along the line. Also, by definition, the *reflection coefficient* r is the ratio of the reflected current vector to the incident current vector. Thus the maximum value of the standing wave will be $(1 + r)$ and the minimum value of the standing wave will be $(1 - r)$, and the s.w.r. and reflection coefficient are related by the expression:

$$\text{s.w.r. } k = \frac{1 + r}{1 - r}$$

The value of r used in the above expression lies between zero (matched line) and unity (complete mistermination) but it is usually expressed elsewhere as a percentage. Typical values are given in Table 13.2.

As defined above, the standing wave ratio will have a range of values from unity (matched line) to infinity (complete mistermination). Occasional references will be found to values of s.w.r. in the range zero to unity. These refer to a scale of values obtained by inverting the expression given, and are exactly reciprocal to the more generally used figures, i.e. an s.w.r. of 0·5 : 1 is exactly the same as an s.w.r. of 2·0 : 1. This system is more frequently encountered in microwave work.

TABLE 13.2

V.S.W.R. in Terms of Reflection Coefficient

Z/Z_0 or Z_0/Z = v.s.w.r.	Reflection coefficient percentage
1·0	0
1·5	20
2	33
3	50
5	67
9	80
inf.	100

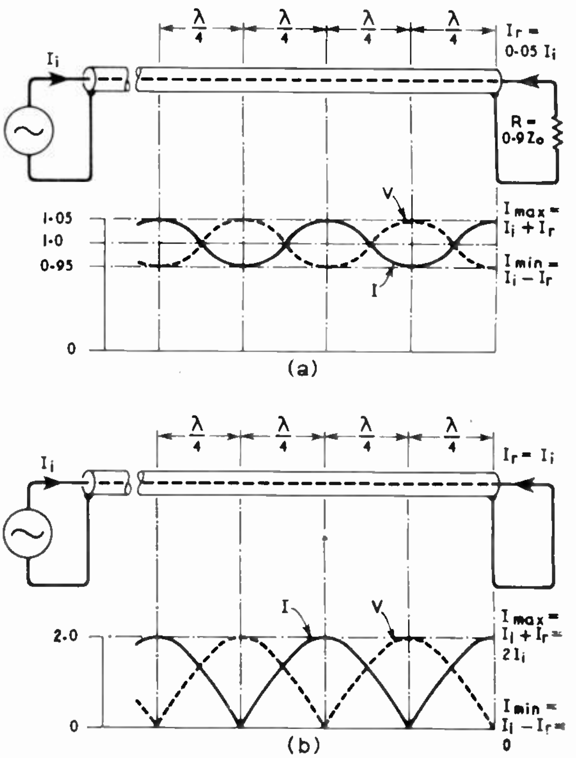


Fig. 13.18. (a) Standing wave pattern for a near-matched line (R = 0·9 Z_0). (b) Standing wave pattern for a completely mismatched line (R = s/cct.). The curve of Fig. 13.17 is typical of the transition at medium s.w.r., from the near-sinusoid of Fig. 13.18(a) to the rectified sinusoid of Fig. 13.18(b).

Fig. 13.18(a) shows the current standing wave pattern on a line which is terminated in a purely resistive load smaller than Z_0. The standing wave is a maximum at the load and therefore a minimum at a point λ/4 back from the load.

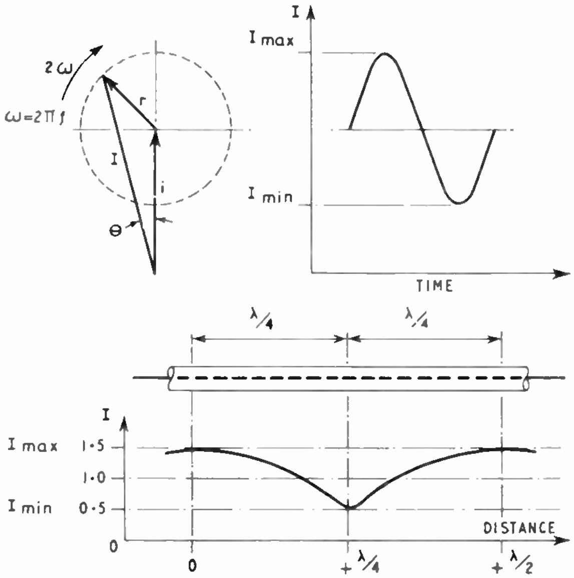


Fig. 13.17. Development of the standing wave pattern due to the resultant of the incident and reflected current vectors. The variation of the resultant line current with distance is derived directly from the variations of incident and reflected current with time; the time variation is effectively removed by maintaining a constant incident current vector.

If a line is terminated in a mismatch R_L, which is entirely resistive, then the value of the s.w.r. is given by the simple relationship:

$$\text{s.w.r.} = \frac{R_L}{Z_0}\left(\text{or } \frac{Z_0}{R}\right)$$

This follows from a consideration of the power distribution at the resistive load. Considering the balance of power flow at the point of connection of the load to the line, the incident power, i.e. that associated with the forward travelling wave, is given by

$$P_{in} = I_i^2 \times Z_0$$

and the reflected power, i.e. that associated with the backward travelling wave, is given by

$$P_{ref} = I_r^2 \times Z_0$$

both from the earlier definition of $Z_0 = \frac{V}{I}$ along the line.

The power in the load resistance R_L will be

$$P_L = (I_i + I_r)^2 \times R_L$$

and must equal the difference between the incident and reflected power, so that

$$Z_0(I_i^2 - I_r^2) = R_L(I_i + I_r)^2$$

and

$$\frac{R_L}{Z_0} = \frac{I_i + I_r}{I_i - I_r} = \frac{1 + r}{1 - r} = k \text{ (s.w.r.)}$$

Therefore any line terminated in a pure resistance will have on it a standing wave, the value of which is given by

$$s.w.r.\ (k) = \frac{R_L}{Z_0}$$

This relationship between R_L, Z_0 and the s.w.r. is a basic one which can be exploited to effect when considering the use of transmission lines as impedance transformers. This is covered later.

The reflected current wave travelling back along the line from the mistermination at the load end, will ultimately reach the generator supplying power. This generator will itself possess an internal impedance which may or may not match the Z_0 of the line. The reflected wave will therefore be re-reflected at the generator to some degree, and will travel forward again towards the load, as a new incident wave which will in turn be reflected at the load end, to exactly the same extent as was the original incident wave. This process of reflection at the load, and re-reflection at the generator will continue indefinitely provided that no power is being lost in the line during the passage of a travelling wave of current, i.e. the current is not attenuated by the line due to its ohmic resistance. Thus eventually all the power originally supplied by the generator will arrive at and be dissipated in the load. There will be finite intervals of time as each part of the output power arrives, those contributions which have made several journeys up and down the line being delayed behind the main contribution delivered to the load without reflection, by an amount depending upon the electrical length of the line. When a transmission system conveying information by means of an audio modulation process is involved the time delays are so short that the ear cannot hear the "echo." When visible images or other pulse forms of transmission are being employed, however, the echo or ghost pulses delayed due to reflections along the line may well become visible to the naked eye as a second image or pulse delayed in time behind the main signal, and diminished in amplitude by an amount dependent upon the inherent loss in the line itself, and the magnitude of the reflection occurring at the load.

The net effect of the passage up and down the line of travelling waves of current is to modify in turn the value of the incident and reflected currents. However, each contribution to the incident current due to re-reflection at the generator is accompanied by a corresponding contribution to the reflected current, due to further reflection at the load. The overall effect of this is that the ratio of incident and reflected currents remains unaltered, and is dependent *only* upon the relative values of the characteristic impedance of the line and the impedance of the load. **The standing wave ratio along the line is dependent only upon the nature of the load at the far end, and no amount of alteration at the generator end can alter the magnitude of this standing wave.**

This is one of the most important aspects of the behaviour of transmission lines.

Since no power is lost in the lossless line, the current waves are not attenuated during their passage along the line, and hence the s.w.r. remains at a constant value for the whole length of the line.

All that has been said so far about the travelling and standing waves of current, is equally applicable to the voltage waves. There is however one important difference in the generation of the standing wave of voltage along the line. In Fig. 13.18(a) the current standing wave has a maximum value at the load, because the load resistance is lower than the Z_0 of the line. Since no power is lost in the line, it must all be dissipated in R, and consequently the voltage at that point must be a minimum. In the extreme case of a short circuit (Fig. 13.18(b)), the voltage must obviously be zero. The voltage standing wave along the line will therefore be 90° out of phase with the current standing wave, so that maxima of current coincide with a minima of voltage and conversely. The voltage standing wave for both conditions of Fig. 13.18 is shown in dotted lines.

Input Inpedance

When a transmission line is operated in a mismatched condition, it is usually necessary to know what value of impedance is seen across the input to the line. This is the load which is presented to the transmittter, and is termed the *input impedance* of the line.

The conditions of voltage and current along a misterminated line are shown in Fig. 13.18, and are dependent upon the nature of the load impedance which fixes the value of voltage and current at its point of connection: the variation of the standing waves is then tied to these values as the "starting" conditions at the far end of the line. In the same way that the voltage and current at the far end are related to the load impedance, the ratio between them at any other point along the line can also be used to determine an impedance which can be said to exist across the line at that point. If the line were cut at a point A and this value of impedance connected, the standing wave pattern back towards the transmitter would be unaffected, and the input impedance would remain the same as before. Equally, the impedance across the line at the point of cutting is the input impedance seen looking back into the section of line which has been cut off. This is shown in **Fig. 13.19.** The actual value of the impedance at any point along the line is determined by the ratio of the actual line voltage and current at the point taking into account any phase difference between them. Fig. 13.18 shows that a voltage maximum is determined by a

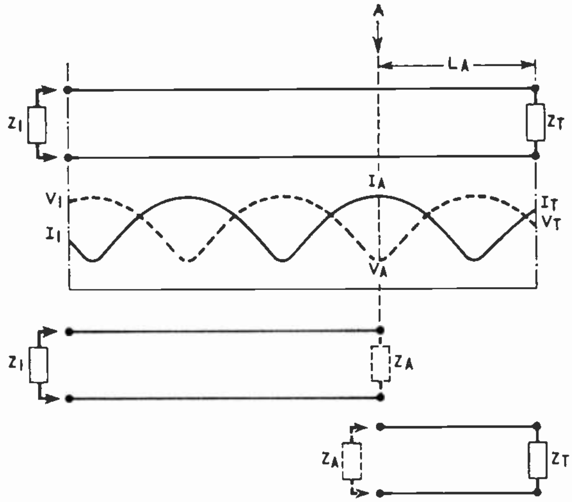


Fig. 13.19. The impedance across the line at point A is such that if used to terminate the shortened line, the input impedance Z_I would be unaltered. The value of Z_A depends upon the ratio V_A/I_A and the phase difference between them: in this example it is less than Z_T or Z_I, and because it occurs at a point of current maximum is the lowest impedance at any position along the line.

purely resistive termination greater than Z_0, and consequently at this, and all other voltage maxima, the voltage and current will be in phase and the effective impedance at those points will be resistive. Equally, at current maxima the impedance will also be resistive, although of a different value. In the regions between these points, the voltage is alternately lagging and leading the current, so that the impedance will be alternately inductive and capacitive in nature along successive half-wavelengths of line, changing sign at the positions where the reactive term becomes zero in **Fig. 13.20.**

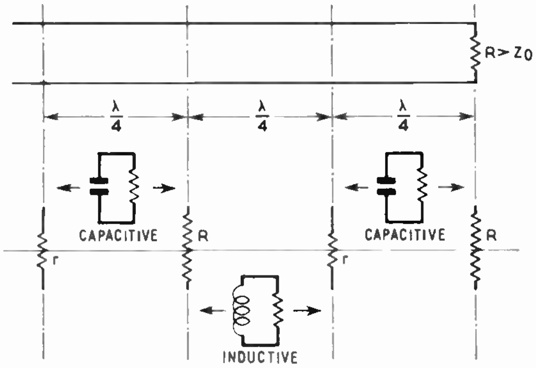


Fig. 13.20. Variation of input impedance along a misterminated line. The impedance is alternately inductive and capacitive for succeeding quarter wave sections. At the precise quarter wave points it is purely resistive.

The input impedance is dependent not only upon the load impedance, but also the length of line involved. A common fallacy is the belief that by altering the length of a line, the match can be improved. In fact the adjustment of line length merely results in the presentation to the transmitter of a more acceptable load impedance into which it can deliver more power. The s.w.r. on the line remains unaltered in every way.

In Fig. 13.18 the case of a purely resistive termination is illustrated. In practice the far end termination may be inductive or capacitive. The effect of this is to cause the positions of voltage maxima to move along the line towards the generator by an amount determined by the nature of the reactive termination. If this is capacitive, as in **Fig. 13.21**, there will then exist along the line a short region *AB* of capacitive impedance before the first voltage minimum. The reactive termination could be replaced by the input impedance of a further short length of line BC terminated in a suitable pure resistance: the length *AB* is then such as to make up the quarter wavelength *AC*. The lower the capacitive reactance of the termination, the closer is the point of voltage minimum.

An important aspect of the input impedance of a mismatched line is its repetitive nature along the line. The conditions of voltage and current are repeated at intervals of one half wavelength along the line, and the impedance across the line repeats similarly. *The input impedance is always equal to the load impedance for a length of line any exact number of half-wavelengths long (neglecting line losses) irrespective of its characteristic impedance.* Such a length of line can be used to connect any two impedances together without

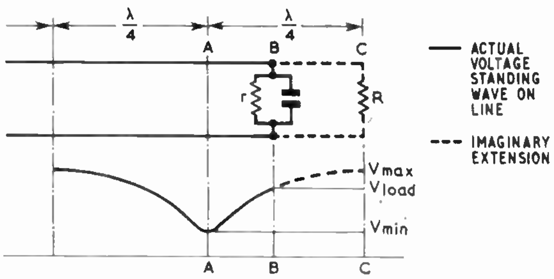


Fig. 13.21. Derivation of the standing wave pattern on a line terminated in a capacitive load impedance. The identical pattern is provided by a hypothetical resistive load at a particular distance beyond the actual line load.

introducing any unwanted variations, and without regard to the Z_0 of the line being used. This property is the basis of the operation of " tuned " lines as aerial feeders, and its application is discussed later. Also, any existing feeder arrangement can be extended in length by any number of half-wavelengths without altering the s.w.r. or the load as seen at the input end. Because the repetitive nature of the line impedance is a function of frequency, being tied to intervals of one half-wavelength, the precise duplication of the load impedance at the input can only occur at a number of discrete frequencies for any specific length of line, for which the line is 1, 2, 3 . . . n half-wavelengths long. The extent to which the input impedance departs from its ideal value equal to the load impedance, as the frequency is varied slightly, is a measure of the impedance bandwidth of the line. For a given percentage change in frequency, the input impedance will vary more rapidly the longer the physical length of line involved, as the extra electrical length introduced by the change of frequency is greater, and causes a proportionately larger change in the standing voltage and current at the input to the line. This limitation can be countered to some extent by using a line of low Z_0, for which the rate of change of impedance is inherently lower. Particular attention should be paid to the impedance bandwidth aspects of lines when they are operating with high s.w.r. such as matching or TVI stubs, or as " tuned " feeders. In the case of TVI stubs which operate with an infinite s.w.r. the effective Q is quite high and their useful bandwidths as filters is strictly limited, often to the extent of $\pm$ 2 per cent of nominal design frequency.

Line Transformers

When a transmission line is used in the resonant condition described above, such that the input impedance is always equal to the load, it can be considered as a 1 : 1 impedance transformer. It is also possible to use a length of line as a transformer having a ratio other than 1 : 1, to connect together without mismatch, two resistive loads of different values. Fig. 13.20 shows that a resistive load at the end of the line gives rise to a resistive impedance of a different value at a point one quarterwavelength away. If the load resistance is R_1 and the resistance at the quarterwave position R_2, then the standing wave pattern due to R_1 is the same as that to be obtained by connecting R_2 at its appropriate position and the line s.w.r. remains unaltered. The s.w.r. has earlier been

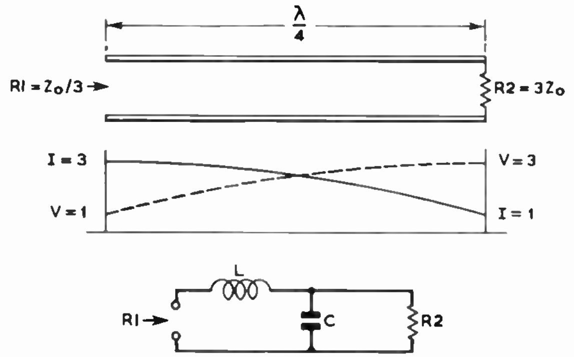


Fig. 13.22. Illustrating the principle of the quarter-wave transformer in the case of a 9:1 impedance ratio. The V/I ratio is inverted by the standing wave on the line with v.s.w.r. = 3, giving R1 R2 = Z_0^2. This corresponds to the series parallel action of a resonant circuit in which L/C = Z_0^2 R2 is the parallel resistance and R1 the equivalent series resistance.

shown to be equal to R/Z_0, so in **Fig. 13.22** it can be defined either as R_1/Z_0 or Z_0/R_2, (keeping s.w.r. greater than unity).

$$\text{s.w.r. } k = \frac{R_1}{Z_0} = \frac{Z_0}{R_2} \text{ and } R_1 \times R_2 = Z_0^2$$

Such a length of line is called a *quarterwave transformer*, and possesses the property that the load impedance is inverted about the line Z_0 to become the input impedance. This holds good no matter what the load impedance (resistive or reactive or both), although it is usually employed to transform a purely resistive load to one of a more convenient value to match into another line.

As with most transmission line problems, the behaviour of the quarterwave transformer can be considered in terms of an equivalent lumped constant circuit of inductance and capacitance, and the relationship between *Rin* and *Rout* is that of the series/parallel conversion provided by the resonant tuned circuit of Fig. 13.22.

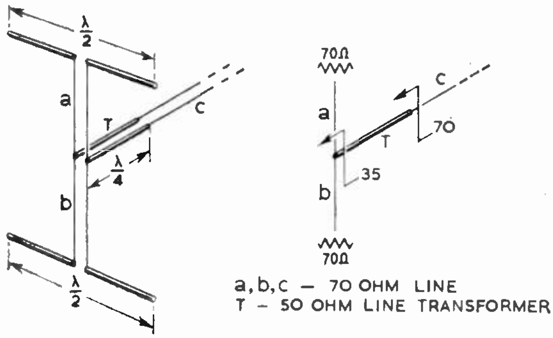


Fig. 13.23. Use of a quarter-wave 50 ohm line transformer to match two co-phased dipoles into a single 70 ohm main feeder. The impedance at the paralleling point is $\frac{70}{2}$ = 35 ohms, which is transformed up to 70 ohms again.

The use of such a transformer to match an aerial array of two half-wave dipoles into a 70 ohm line is shown in **Fig. 13.23.** The bandwidth of a single section quarterwave transformer is not large, and a substantial improvement can be obtained by achieving the required transformation in two quarterwave sections in series, transforming to an intermediate value at their junction, which should ideally be the geometric mean of the end impedances. In such an arrangement the effective ratio of each transformer is reduced, which itself improves their bandwidth, and in addition the two tend to be mutually compensating with change of frequency to produce a further improvement, **Fig. 13.24.** The choice of cable impedances limits the application of this technique, but it can be usefully applied at v.h.f. and u.h.f. using fabricated coaxial lines of appropriate Z_0, and at h.f. with balanced lines of appropriate spacings.

The inverting properties of a quarterwavelength of line can also be used to relate the voltage at the input in terms of the current in the load (and vice versa). In Fig. 13.22, the s.w.r. on the line can be defined in a number of ways.

$$\text{s.w.r.} = \frac{V_2}{V_1} = \frac{Z_0}{R_1} = \frac{Z_0 I_1}{V_1}$$

so that $V_2 = I_1 \times Z_0$

and similarly $V_1 = I_2 \times Z_0$

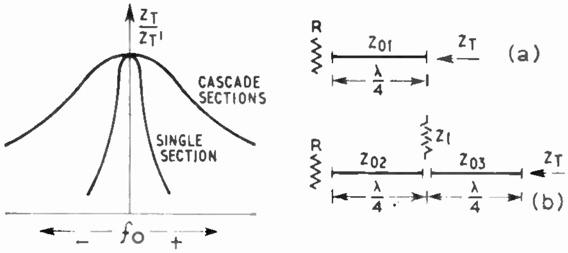


Fig. 13.24. Bandwidth of single and double section quarterwave transformers. Z_T is the actual input impedance at each frequency, while Z_T' is the theoretical design frequency value. In the two-section case, Z_i is the geometric mean of R and Z_T.

The voltage at the input end is determined *only* by the characteristic line impedance and the current at the other, and does not depend upon the value of the load in which the current flows. This property finds applications when it is necessary to drive a certain current through a load impedance (usually a radiating aerial) without reference to the actual value of the load impedance itself.

Stubs

The limiting case of a mismatched line occurs when the far end is completely misterminated by either a short or open circuit. There is then no load resistance to dissipate any power, and a 100 per cent reflection of voltage and current takes place. The standing wave pattern is then that of Fig. 13.18(b). Sections of line with open or short circuit terminations approximate to pure capacitance or inductance, the only loss being that associated with the line attenuation, which can usually be ignored when the length involved is short electrically. Such pieces of line, usually of lengths not exceeding λ/4 are called *stubs* and can be used to compensate for, or match out unwanted reactive impedance terms at or close to a mismatched termination, and allow the major part of the feed line to operate in a matched condition.

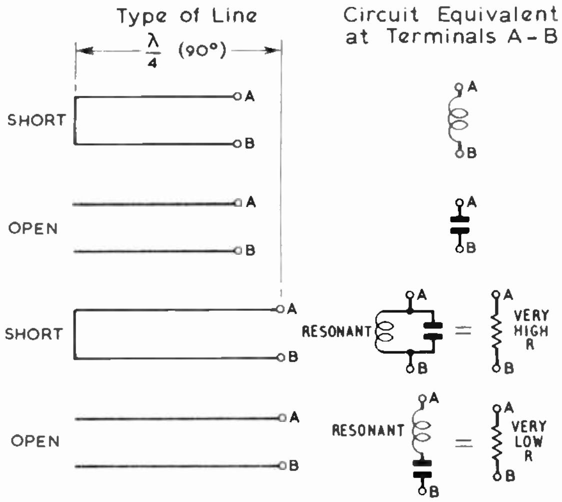


Fig. 13.25. Circuit equivalents of open and short-circuited lines.

Figs. 13.25 and 13.26 illustrate the manner in which the input reactance varies with line length. For sections shorter than $\lambda/4$, the short circuited line is inductive, and the open circuited line capacitive. A short circuited line can also be made to behave like an open circuited line, of the same electrical length (and vice versa) by adding a further $\lambda/4$ of line: this is another application of the inverting properties of a $\lambda/4$ section of line described previously. The behaviour of the two types of stub is completely complementary, and if two are joined together at their input ends they will make up between them precisely one quarterwavelength when their respective reactances are equal and opposite in sign. This is exactly analogous to a parallel tuned circuit, and a short-circuited $\lambda/4$ line is said to be a " resonant " section, possessing all the properties of the parallel tuned circuit. The reactance-frequency slope of the resonant line is similar in shape to that of the lumped constant circuit, and a change of Z_0 is equivalent to a change in the L/C ratio. Such sections of line may be used as shunt chokes across other lines without effecting the latter, and may also at v.h.f. be used as tuned circuits for receivers and transmitters, when the physical lengths involved are conveniently short.

In complementary fashion, a $\lambda/4$ open circuit line section behaves as a series resonant circuit, with a very low input resistance at the centre frequency. In the ideal case this is a perfect short circuit: in practice the line losses involved degrade this to a finite but low value of resistance, which for all practical purposes can be considered short circuit. The reactance-frequency slope of both open and short circuit resonant sections is shown in Fig. 13.27 which illustrates the effect of variations in Z_0.

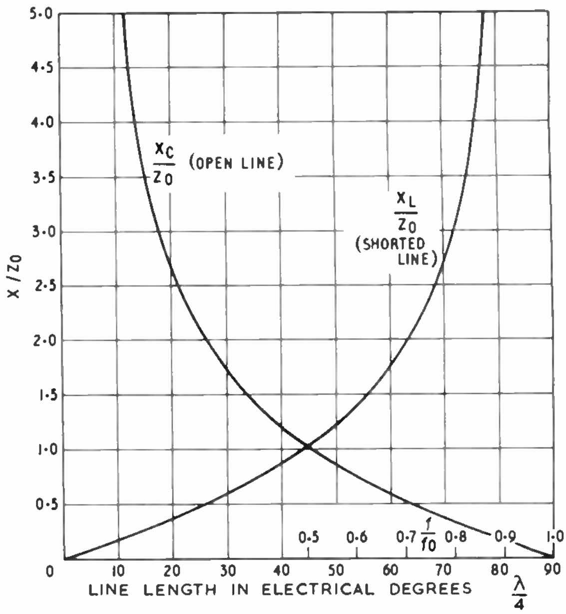


Fig. 13.27. Reactance curves for open and short circuited lines. The quantity X/Z_0 multiplied by the characteristic impedance of the line equals the reactance at the input terminals. Longer lines alternate as capacitance or inductance according to Fig. 13.26. The curves also indicate the manner in which input reactance varies with frequency for nominal quarter-wave sections.

The input reactance of a loss free short circuited line is given by the expression:

$$Xin = Z_0 \tan \theta \text{ (inductive)}$$

and for an open circuited line by:

$$Xin = -Z_0 \cot \theta \text{ (capacitive)}$$

where θ is the electrical line length and Z_0 its characteristic impedance (see page 13.8).

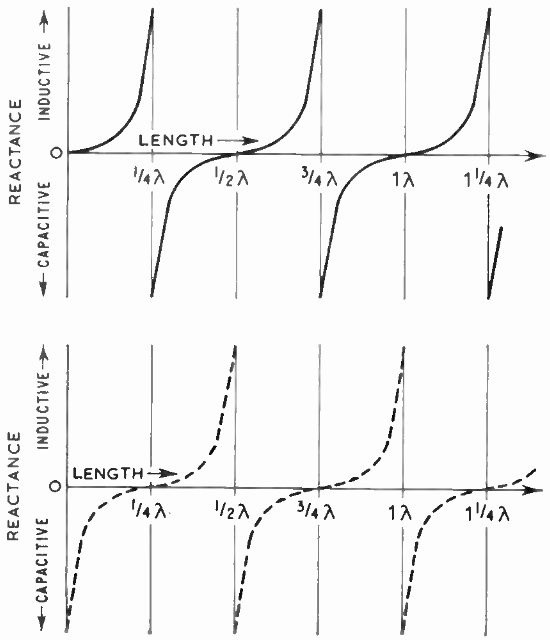


Fig. 13.26. Variation of input reactance with length for short circuited (solid) and open circuited (broken) lines. The two sets of curves are identical and displaced by a quarter-wavelength.

The practical applications of matching stubs are considered on page 13.25 in the section dealing with Impedance Matching. Their use as single frequency filters is covered in detail in Chapter 18—*Interference*.

Attenuation and Loss

In practice all transmission lines have associated with them some loss which is experienced when power is transmitted along them. This loss may be due to radiation (from balanced lines), resistive losses in the conductors, and leakage losses in the insulator; but however it arises, it is a function of the actual construction of the line and the materials employed. With the moulded twin or coaxial lines for amateur and television use, this attenuation is less than 1db per 100 ft. (when correctly matched) at frequencies below 30 Mc/s. One db loss per 100 ft. means that at the far end of the feeder the amplitude of voltage or current is just under 90 per cent of that at the input end, or the power approximately 80 per cent. A second 100 ft. extension would deliver 80 per cent of the power left at the end of the first 100 ft. or 64 per cent of the original input. For each extra 100 ft. the factor is multiplied again by 0·8 but in decibel ratios 1·0db loss is added. Open wire lines can be remarkably efficient: a 600 ohm two-wire line made from 16 s.w.g. copper wire is spaced about 6 in., carefully insulated and supported, has a loss of only a few decibels *per mile* in the h.f. range.

The *matched loss* of the line is quoted in the manufacturers' published information usually as *n* db per 100 ft. It increases more or less in proportion to the square root of the frequency at which it is being used, i.e., a line having a quoted matched loss of 3db/100 ft. at 10 Mc/s will have a matched loss of approximately 9db/100 ft. at 100 Mc/s (3db $\times \sqrt{10}$). The effect of this loss is to attenuate the magnitude of the current and voltage waves as they flow along the line, and since the loss has no sense of direction, it will equally well attenuate the reflected wave of current as well as the incident wave.

If the line is operated with standing waves, the loss is greater than in the matched case, because increased voltages and currents are circulating on the line and the average heating of conductor and insulator is greater for the same power output. The extra loss due to standing waves increases with the s.w.r.

When power is fed into the sending end of a practical line which possesses some degree of attenuation and is terminated in a short circuit, it is again propagated along the line as incident waves of current (and voltage). The magnitude of the incident current decreases as the wave propagates along the line due to the presence of the attenuation, although its frequency and hence relative phase remain unaltered.

When it arrives at the termination it will have the same phase, but a smaller magnitude than in the example of the loss-less line, i.e. as far as the load is concerned it is equivalent to a reduced power flowing into the generator end of a loss-less line. At the point of connection of the load, a reflected wave will be set up as before, its value and phase being dependent upon the nature of the load, and also relative to the value of incident current *at the load*. This reflected wave then flows as before back along the line, experiencing attenuation during its passage, until it arrives back at the generator, in the same phase as in the loss less case, but with a further reduction in magnitude (Fig. 13.28). The standing

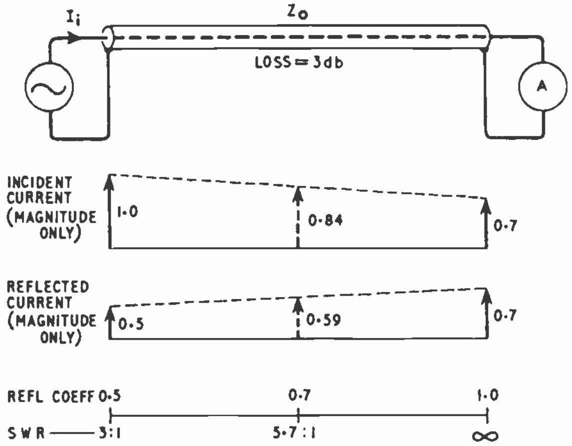


Fig. 13.28. Effect of line attenuation on incident and reflected currents.

wave ratio is related to the reflection coefficient at any point on the line, and that the reflection coefficient is defined by the ratio of the incident and reflected currents at the point in question. Applying this to the case of the line with a finite loss it becomes evident that the reflection coefficient and hence the s.w.r. are not the same at each end of the line, and indeed change constantly along the length of the line. The effect of the line attenuation is to cause an improvement in the s.w.r. towards the generator. In other words, when feeding power to a mis-matched load through a transmission line having a finite loss, the mismatch appears less severe at the transmitter.

This property can be used when it is required to present a dummy load of reasonable accuracy to a transmitter (or receiver). If a long length of cable of fairly high attenuation at the operating frequency is available, it is sufficient merely to connect one end of this to the equipment to be tested, and ignore the other end. The input s.w.r. will be reasonably low, the actual value depending upon the attenuation of the cable. For example, if sufficient cable is available to have a matched loss of 10db, the input s.w.r. will be only 1·22 : 1, and the load impedance presented to the equipment will be $Z_0 \pm 20$ per cent. This can be improved still further by using a low dissipation resistor approximately equal to Z_0, on the far end of the cable "load." This technique is particularly useful at v.h.f. and u.h.f. where the smaller diameter cables tend to be quite lossy.

In the case of practical lines, in which the whole object is to achieve as small an attenuation as possible, the improvement in the match presented to the transmitter is not great, and indeed for typical h.f. band installations, is negligible, since the matched feeder loss is only likely to be of the order of 0·5db, and hence the reflected current at the transmitter will only be reduced by 1db (go-and-return) or 10 per cent. Since the reflected current is typically only 10 per cent of the incident current anyway (corresponding to an s.w.r. of 1·22), the net effect of a line attenuation of 0·5db would be to reduce the reflected current to 9 per cent of the incident current (corresponding to an s.w.r. of 1·20). For a similar piece of line at v.h.f. (say ten times the frequency of the h.f. case), then the line loss will have increased by $\sqrt{10}$, or approximately a factor of 3, i.e. 1·5db (or 3db go-and-return).

Then in this case, for an s.w.r. at the load of 1·22, the reflected current at the transmitter will be reduced to 7 per cent instead of 10 per cent, giving an effective s.w.r. at the transmitter of 1·15. Taking this a stage further to u.h.f., again by a factor of ten in frequency the return line loss will have increased to 10db, giving an effective s.w.r. at the transmitter of 1·06. These results are tabulated in Table 13.3.

TABLE 13·3

Frequency	Line Loss	S.W.R. at load	S.W.R. at transmitter
14 Mc/s	0·5db	1·22	1·20
140 Mc/s	1·6db	1·22	1·15
1400 Mc/s	5·0db	1·22	1·06

Effect of line attenuation on the s.w.r. presented at the transmitter, for a constant s.w.r. at the load.
Load S.W.R. = 1·22 = 10% reflection coefficient.

We have so far considered the nett effect of line attenuation upon the s.w.r. presented to the transmitter. Of greater interest is the resultant additional loss of power in a practical line, with a finite matched loss, when it has upon it a standing wave. In the case of the loss-less line all the power delivered to the line is eventually dissipated in the load, although a certain proportion is delayed in time due to multiple reflections. When the mistermination of the line is not gross, the amount of power returned from the first reflection at the load is not large.

Assuming that the generator completely misterminates the line at the sending end then the whole of the power from this first load reflection will reappear at the load after a journey back down the line and up again during which time it is attenuated by twice the matched line loss (go and return). At this stage the major proportion of this attenuated reflected power is then delivered to the load, and a small proportion is re-reflected, so that the nett power delivered to the load after only one reflection represents an appreciable proportion of that delivered by the generator, and is only slightly less than that which would have been dissipated in the load, had there been no attenuation in the line.

In practice, any reactive term present in the load offered to the transmitter by the line is removed by re-tuning of the output stage, and any variation in the resistive term merely alters the effective loading on the transmitter and limits the amount of power delivered to the line (none is "lost" because the d.c. input to the transmitter is also limited by the variation in loading).

In other words the transmitter presents virtually a complete mistermination to the reflected power in the line and the additional power lost due to the presence of reflections from the load is relatively small, arising only from the attenuation of the reflected wave. This point is frequently misunderstood, and the reader is led to believe that once power has been reflected from a misterminating load this reflected power is lost. This case *only* occurs when the transmitter itself presents a perfect termination to the line and can absorb this reflected power, which does not arise in practice.

This power loss by mismatch is, however, important in the case of a receiving system. In such an arrangement, the input impedance of the first stage of the receiver is transformed to present as good a match as possible to the transmission line transferring power from the aerial. Thus the receiver becomes the load, and the aerial becomes the generator. In such a case, the load is adjusted to match the line by design, and by its very nature, the aerial is also a reasonable match to the line. In such circumstances, any signal power reflected from the receiver input is re-radiated by, or effectively dissipated in the radiation resistance of the aerial. This reflected power does not return to the receiver

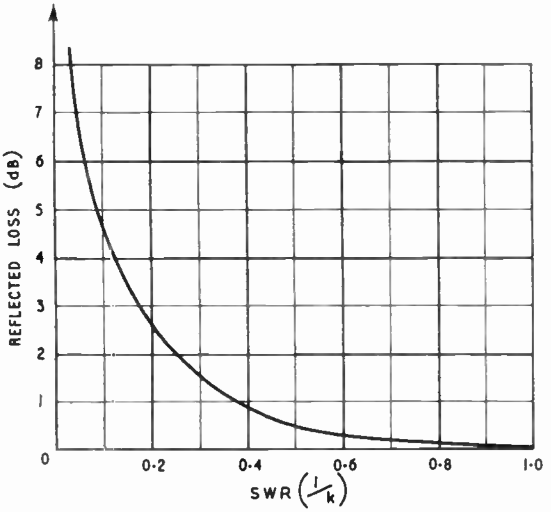


Fig. 13.29. Curve of mismatch loss, i.e. the amount of power reflected from a mismatch for a given s.w.r. This is also the curve of absolute power loss (when the generator terminates the line, and therefore absorbs the reflected power) for lines of zero matched loss.

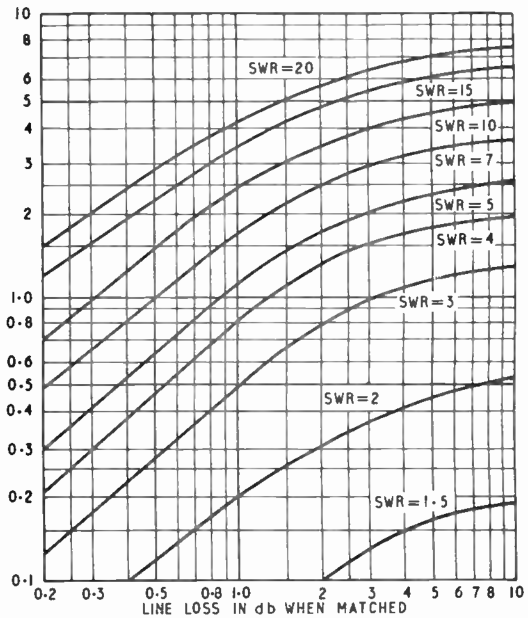


Fig. 13.30. Curves of absolute power loss for any given s.w.r. and matched line loss (assuming that the generator completely misterminates the line).

input to provide a contribution to the total signal received, but is lost. The extent of the additional power lost due to such a mismatch on a receiving aerial feeder is illustrated by **Fig. 13.29.** The difference between this case and that of the transmitter can be seen by comparing this graph with that of **Fig. 13.30,** which shows the additional loss due to a mismatch on a line in the transmitting case. It is, of course, true to say that in practice neither case is perfectly satisfied, i.e. the transmitter does absorb some of the reflected power in the one case, and the aerial does re-reflect some of the reflected power in the other case. The respective curves deal with the limiting conditions, and represent the maximum additional loss that can be experienced in each case.

It will be seen from Fig. 13.30, that when the matched line loss is very high, the additional loss due to the presence of a standing wave is independent of the matched line loss.

Power Handling

There is always a definite limit to the amount of r.f. power which can be transmitted along any form of transmission line with adequate safety. The limitation is set either by excessive heating of the conductors, which can lead to deformation or destruction of any artificial insulant between them, or by voltage breakdown between the conductors, either in air or again in the insulating material.

Open wire balanced lines will generally carry a greater power than any other form, there being adequate ventilation around the conductors and usually a relatively large spacing between them. Concentric lines are usually manufactured to lower impedances, with smaller spacings between the conductors, and because the inner conductor must be maintained accurately in alignment, there is often a considerable amount of dielectric material involved. The inner conductor is therefore well insulated thermally from the surrounding air and will rise to a higher temperature for a given current flow, than its open wire counterpart.

In amateur practice, most lines other than the poorest quality coaxial cables are quite suitable for use up to the maximum authorized power, although for the u.h.f. bands the smaller diameter cables should be avoided when carrier powers in excess of 50 watts are employed. This is usually done anyway, because the cables of lower power rating are also those with the highest attenuation per unit length.

The amateur should pay attention to the power rating of cables employed, when a high s.w.r. exists on the line, either by accident or design. In such circumstances, the value of the current at maxima of the standing wave may rise to such an extent that local overheating will occur, with consequent damage. The voltage maxima may also approach the limiting value for flash-over. These conditions are unlikely to obtain in any well designed installation, but the possibility of their arising, particularly under fault conditions, should not be overlooked.

Practical Feeders

A variety of concentric and two-wire lines is commercially available, the most common being 50 ohm and 70 ohm coaxial cables with a solid or stranded inner conductor and a braided wire sheath. Two-wire lines are made with impedances of 70, 80, 150 and 300 ohms, are generally constructed in moulded polythene insulation and thus quite flexible and weatherproof. Such lines are sufficiently low loss to be suitable for amateur transmission. Their general properties are described in more detail in Chapter 14 (*V.H.F. Aerials*).

The choice of type of feeder may be governed by circumstances, but in the matter of characteristic impedance it is always best to suit the line to the aerial impedance; the use of impedance transformers at the aerial connection is always a disadvantage and may lead to practical difficulties.

For distances greater than a few wavelengths, the loss in small-size manufactured lines may be enough to warrant the construction of an open-wire twin line. This will necessarily be of a fairly high impedance, say, 300-600 ohms. Soft-drawn enamelled instrument wire of 14-18 s.w.g. can be used, although when the line is to be strained between insulators, it is better to use hard-drawn or cadmium-copper wire. Pre-stretching to about 10 per cent extra length by means of a steady pull will appreciably harden soft-drawn copper wire and render it less liable to stretching in service.

Spacers made from $\frac{3}{8}$ in. diameter wood dowelling or polythene rod should be fitted every few feet along the open line. Wooden spacers should be thoroughly impregnated by boiling them in wax. A small hole should be drilled in each end to take a 20 s.w.g. binding wire. A groove in the end is also a helpful aid to secure binding. It is not often practicable to thread the spacers on to the line wires themselves **(Fig. 13.31)** and the arrangements shown inset in the illustration may usefully be employed.

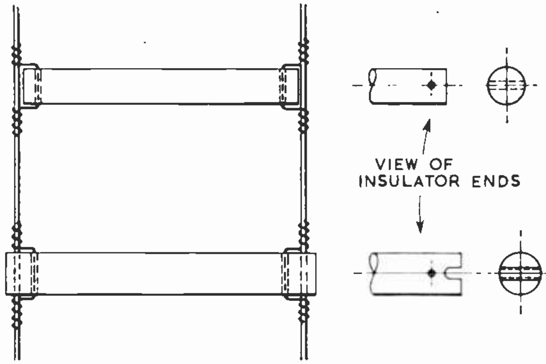


Fig. 13.31. Spacing insulators for open wire line. The lower pattern with an end locating groove is to be preferred. In each case, the insulator is held in place by a 20 s.w.g. binding wire passing through a hole in the insulator body.

Moulded twin lines sometimes suffer noticeably from the effect of moisture in damp weather. This is most noticeable in the case of tuned lines, because water has a very high dielectric constant and affects the velocity factor of the line as well as its loss factor. Soot deposited on the surface will tend to retain moisture and accentuate these troubles. The remedy is to clean the line periodically and give it a dressing of silicone wax polish to repel any moisture. Tubular 300 ohm line should be sealed so that water cannot enter the interior. Although the p.v.c. outer jacket of a concentric cable is good enough to allow the cable to be buried, the open end of the cable is very vulnerable for it can " breathe in " moisture which does irreparable damage to the line. There is a variety of satisfactory sealing materials available,

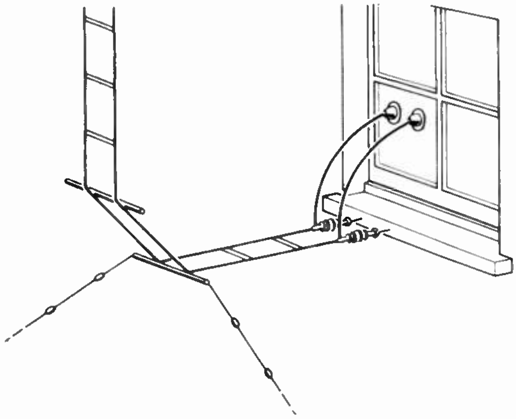


Fig. 13.32. Preferred layout of balanced line installation for optimum performance. A lower end fixing and a slow bend are included in this example.

such as Bostick cement, Bostick sealing strip (putty) and Sylglas tape which is loaded with a silicone putty.

Particular attention should be paid to the adequate sealing of plugs and sockets used out of doors on coaxial cables. Even those intended for professional use (e.g., BICC Teleconnectors, US type C and BNC) are not entirely waterproof, and the usual Belling-Lee coaxial plug is virtually unsealed in its own right. Such fittings should be securely taped, for at least 1 in. along the cable with a water repellent tape. The most effective and reliable sealing is obtained by a layer of self-amalgamating P.I.B. tape, covered by a wrapping of a putty loaded fabric tape. (The former is manufactured by Rotunda Ltd., Denton, Manchester, and the latter, known as Densotape, by Winn and Coales Ltd., Denso House, Chapel Road, London, S.E.27).

In selecting an appropriate feeder for use with an aerial system, due allowance must be made for the physical layout involved. Any installation employing a balanced line must be arranged so that the line is kept well clear of masts, buildings and other obstructions. Failure to do this can result in severely unbalanced currents in the two conductors, which will then become a source of radiation from the feeder itself, with undesirable effects on the required aerial performance. Unbalance can also be caused by abrupt changes of direction of the feeder, and to a lesser degree by slackness in the line which can allow it to swing about in the wind and to twist.

As far as possible, all balanced lines should be lightly tensioned in straight clear runs, with bends broken up into a slow change of direction. Mechanical restraints to the feeder can be secured to the extended arms of selected spacer insulators, the straining wires being either broken up with insulators or preferably made from artificial fibre cords (e.g., nylon, terylene). A practical layout is illustrated in Fig. 13.32 in which it will be noted that the lower end of the feeder is tensioned to a secure anchorage, and flexible tails are used to connect to the feed-through insulators to relieve them of any strain. If at any point along the feeder it becomes impossible to avoid obstructions, the balanced line is best terminated in a tuning or coupling unit (see page 13.36) and the rest of the line to the transmitter run in a coaxial cable.

Coaxial cables lend themselves to more unobtrusive installation. If they are correctly terminated in an unbalanced load, no currents will flow on the outside of the outer conductor and they may therefore be laid close to any other surface or object (brickwork, tubular masts, etc.) without affecting their behaviour in any way. It is only necessary to ensure that any bend is made on a sufficiently large radius to avoid distortion of the cable itself. Typically, small ¼ in. diameter television type cables may be formed round a ½ in. radius, whereas larger cables such as UR67 should be formed on no tighter than a 2 in. radius. The recommended minimum bending radius for cables can often be found in manufacturers' literature, but a useful rule of thumb is to use a radius equal to at least twice and preferably four times the cable diameter. It is easy to damage coaxial cables by excessively tight fastenings, such as staples driven hard into the cable support, and care should be exercised when making such fixings to avoid bruising of the cable.

When plugs and sockets or other forms of termination are used on out-of-door coaxial cables, the cable should always be arranged in such a way that any water running down the outside is shed from the outer before it can run into the termination. The right and wrong ways for fixing such terminations are illustrated in Fig. 13.33.

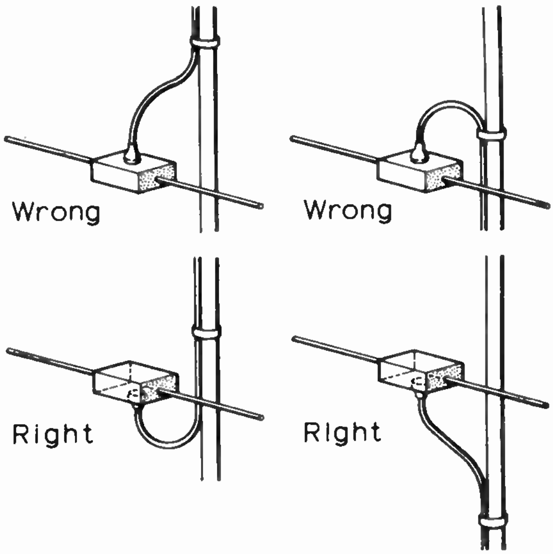


Fig. 13.33. Method for terminating outdoor co-axial cables. The reverse loop in the cable prevents water from running back into the end termination of the cable.

The correct operation of coaxial cables relies entirely on their being terminated in an unbalanced load. Since most aerial systems present a balanced impedance at their feed point, some form of balance to unbalance transformer must be employed. Suitable devices are described later in this chapter.

IMPEDANCE MATCHING

In nearly every arrangement of aerial and feeder system there is a requirement to provide some means of impedance transformation, or *matching*, at a particular point between the aerial and the transmitter. It may be in the form of an aerial coupler described above, which transforms the varying input impedance for a tuned line system down to a suitable load value for the transmitter. It may be a network at or near the aerial terminals which transforms the aerial input impedance to the correct value to match a length of flat line back to the transmitter. In each case the function of the transformer is to adapt some arbitrary impedance in order correctly to terminate the line which connects it back to the transmitter. The matching network has no effect on the impedances encountered on the aerial side and can therefore not alter the s.w.r. on the line which connects the network, be it constructed of lumped components or transmission line stubs, back to the aerial terminals. It is important to understand that the process of matching refers to the elimination of standing waves only on the line between the point of application of the matching and the transmitter delivering the power (Fig. 13.34).

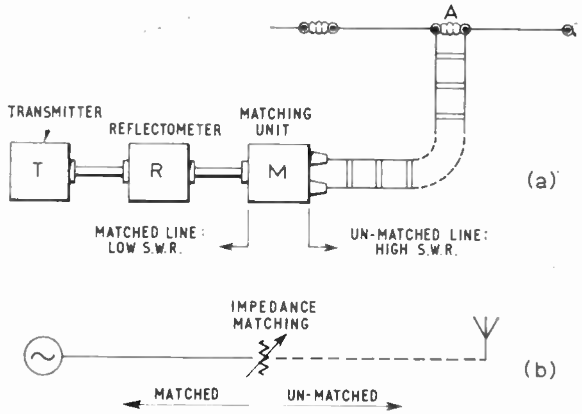


Fig. 13.34. Application of impedance matching. Introduction of the matching unit can improve matters only on the section M—T. The feeder on the aerial side of the matching unit, M—A, remains "unmatched" with a high s.w.r. It should therefore be as short as possible and of low attenuation (see text). The matching unit, often called an aerial coupler in this application, can be adjusted using a reflectometer to indicate minimum reflected power in the "matched" section of the line to the transmitter.

In discussing the operation of tuned line feeder systems it was explained that they were acceptable, carrying as they do a high standing wave, only because the inherent line loss and hence the additional loss due to the standing wave is very low. It was also explained on page 13.15 that the bandwidth of the aerial system reduces rapidly when an increasing length of tuned line is employed, because the latter is in effect a series of cascaded resonant circuits, so that even in the narrow span of an amateur band re-tuning of the coupling circuits may be necessary as the frequency is changed. Any necessary matching should therefore be carried out as near as possible to the aerial. In a similar way, a reduction of bandwidth occurs if a large ratio of impedance is to be transformed, because high ratio transformers have smaller bandwidths and greater loss. It is therefore best to use aerials and feeders with impedances as nearly alike as possible. This cannot always be done, especially if a long line is necessary, because it would then be preferable to employ a high impedance line in order to minimize losses. Another factor which must always be considered is that the feeder should not carry unintended waves, a matter discussed later under *Marconi Effect*.

The technique employed to achieve the necessary matching will depend upon the nature of the unmatched impedance and the physical considerations of the aerial. In some cases it is possible to modify the aerial itself to achieve the required match to the feed line, without significantly altering its characteristics as a radiating element. Examples of this are the use of folded wires, and the adjustment of reflector spacing and length in Yagi type beam aerials. In the case of simple wire aerials it is not often practicable to match on the aerial wire, and a suitable network, either of lumped constant components or more usually comprising an arrangement of transmission line stubs, is interposed between the aerial terminals and the main feeder. The use of networks of coils and capacitors is restricted by physical convenience to the provision of parallel resonant pre-tuned traps in multi-band aerials, and to matching transformers for use at the base of a vertical radiator.

Folded Dipoles

The half-wave dipole is a balanced aerial and requires a balanced feeder. Normally this is a 70 ohm line, but it is sometimes necessary to step up the aerial input impedance to a higher value; for example, where the feeder is very long and

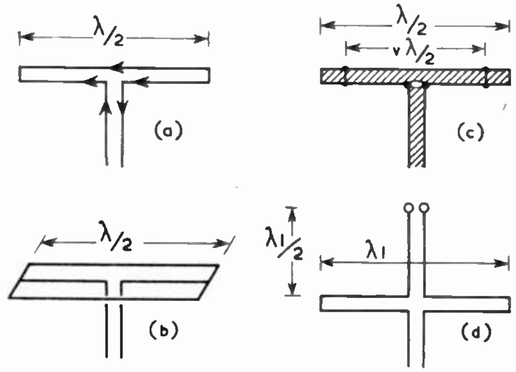


Fig. 13.35. Folded dipoles, including (a) simple fold; (b) three-fold; (c) folded dipole made from 300 ohm ribbon feeder with two (d) arrangements for working on half- and full-wave resonance.

300 or 600 ohm line is employed to reduce feeder loss. Another is in a beam aerial in which the dipole input impedance is too low to match 70 ohm line. A separate transformer to step the aerial impedance up to these values would reduce the bandwidth, but it has been found that by folding the aerial the step-up can be accomplished without any reduction of bandwidth. **Fig. 13.35 (a)** shows a full-wave wire folded into a half-wave dipole. It is not a loop aerial because the two halves are only separated by a few inches.

The relative direction of currents, at current maximum, are shown by arrows in the diagram. In a straight full-wave wire they would be of opposite polarity, but folding the full-wave has " turned over " one current direction; thus the arrows in the diagram all point the same way. The folded dipole is therefore equivalent to two single-wire dipoles in parallel. The radiation resistance of the folded dipole made of the same diameter conductors is 300 ohms, which is four times that of the single wire.

One of the advantages of folding a dipole is that the bandwidth is increased because it becomes equivalent to an aerial made of a much larger diameter conductor. Such aerials have a lower Q than those made of small gauge wires. Low Q is desirable in an aerial provided it is produced by radiation load and not by conductor or insulator loss.

A folded dipole connected to a 300 ohm feeder will operate over the whole of the 14 Mc/s band with no appreciable change of impedance. For comparison a single wire dipole connected through some kind of transformer and 100 ft. of 300 ohm line might well show a 2 : 1 change of impedance over the same band.

If the aerial is made three-fold (**Fig. 13.35(b)**, it is equivalent to three radiators in parallel, its radiation resistance is multiplied nine times and it would be used with a 600 ohm line.

In some multi-element arrays the radiation resistance is very low, and folding the driven element is a useful aid to correct matching since with N wires the resistance is always raised N^2 times.

The folded dipole will not normally operate on even harmonics of the frequency for which it is cut because the " folded " currents cancel each other. However, the alignment of currents is correct for the odd harmonics, so that a folded dipole for 7 Mc/s can also be used on 21 Mc/s.

A folded dipole one wavelength long can be made to radiate but only if the centre of the second conductor is broken; in such a case the input impedance is *divided* by four. The centre impedance of a full-wave dipole is 4000 to 6000 ohms; that of the folded version 1000 to 1500 ohms. **Fig. 13.35(d)** shows how such an aerial can be made to work at both fundamental and second harmonic frequencies by means of a length of open circuit twin line (called a *stub*) used as a " frequency switch." At the frequency at which the aerial is $\lambda/2$ long, the stub is one electrical quarter-wave long and, since its other end is open, the input end behaves as a short circuit and effectively closes the gap in the aerial. At the full-wave frequency the stub is one half-wave long and behaves as an open circuit. The stub may be made of 300 ohm twin feeder and in application to rotating beam aerials can be fixed to the mast or dropped down inside it if it is tubular.

The folded dipole can be conveniently made from 300 ohm flat twin feeder as in **Fig. 13.35 (c)**, because the spacing is not important so long as it is a small fraction of a wavelength. Since the velocity factor of ribbon feeder is about 0·8, it is necessary to place the short circuits between the wires at this fraction of a quarter wavelength from the centre rather than at the end—e.g. 27 ft. apart in a 33 ft. long 14 Mc/s aerial.

Other ratios of transformation than four or nine can be obtained by using different conductor diameters for the elements of the radiator but these are more practicable on the higher frequencies where tubular elements can be employed. When this is done, the spacing between the conductors is important and can be varied to alter the transformation ratio. The relative size and spacings can be determined with the aid of the nomogram in **Fig. 13.36.**

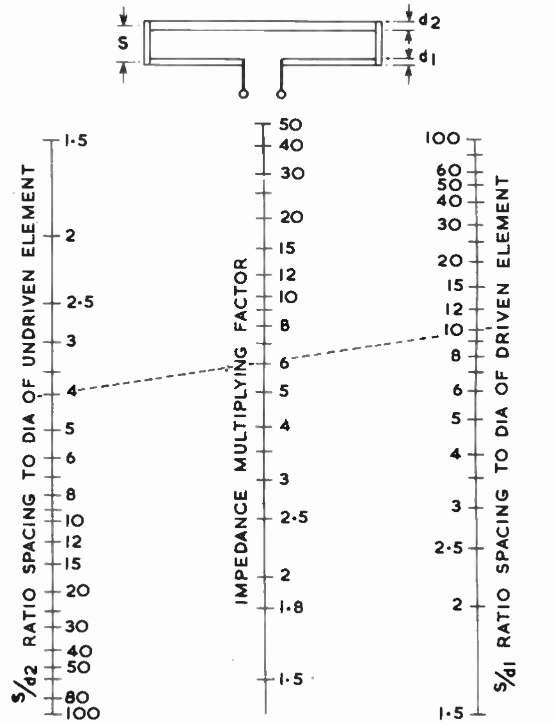


Fig. 13.36. Nomogram for folded dipole calculations. The impedance multiplying factor depends on the two ratios of conductor diameter to spacing between centres, and is always 4 : 1 when the diameters are equal. A ruler laid across the scales will give pairs of spacing/diameter ratio for any required multiplier. In the example shown the driven element diameter is one-tenth of the spacing and the other element diameter one quarter of the spacing, resulting in a step up of 6 : 1. There is an unlimited number of solutions for any given ratio. The chart may also be used to find the step-up ratio of an aerial of given dimensions.

Aerial Transformers

In aerial systems comprising full-wave (or end fed half-wave) radiators, it is possible to obtain some control over the high input impedance by utilizing the radiator itself as a quarter-wave transformer of single wire line. The principle was explained on page 13.4. The loop radiation resistance of the radiator is principally a function of its length, and is for resonant lengths independent of the conductor size. By varying the latter to alter the effective L/d ratio of the radiator, the Z_0 of the single wire line can be altered and hence the input impedance at the voltage feed point. This is not practicable as an empirical adjustment after erection, but does permit some elementary matching to be built-in to the aerial at the design stage to restrict the extent of the initial mismatch to the open wire main feeder.

The effective Z_0 of the radiator as a transmission line is a function of the length/diameter ratio, and since the length is determined by the operating frequency, the diameter must be adjusted to control Z_0. This may be achieved by using a thicker wire, or more effectively by using several wires connected in parallel and spaced out to form a cage. Another alternative is to use several wires which are spread out from their common end at the feed point, to form a fan or semi-cone. These methods are illustrated in **Fig. 13.37.** The curve in **Fig. 13.38** shows the variation of input impedance of a full wave biconical dipole in terms of the cone angle: it is correct for solid cones and indicative of the general tendency for elemental cones comprising several wires. The values of input impedance for a fan dipole will be 50-100 per cent higher for the same angle.

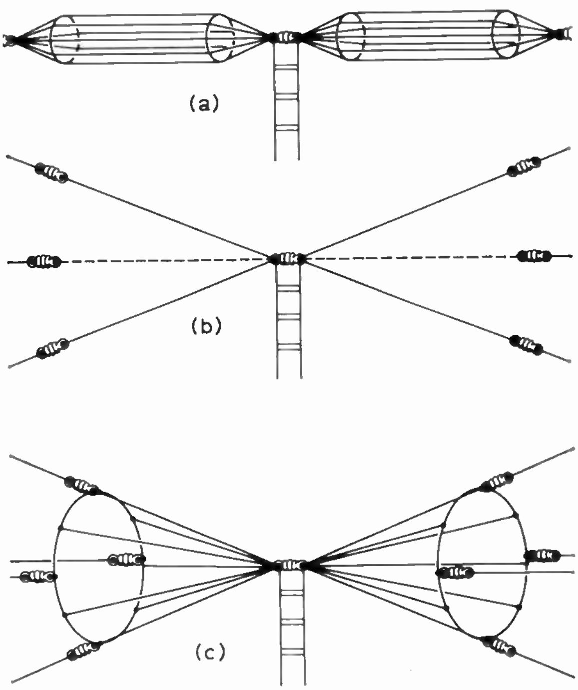


Fig. 13.37. Broadband dipoles. (a) Cage. (b) Fan (two or three wire). (c) Conical. The semi-conical form is often used for broadband vertical monopoles over an earth plane.

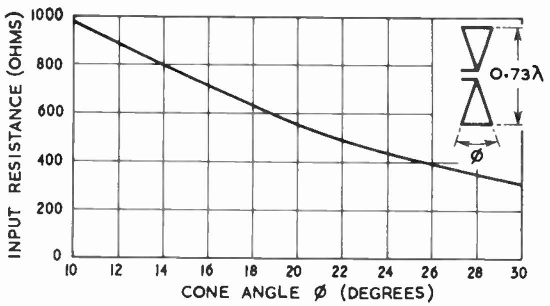


Fig. 13.38. Input resistance on a full-wave bi-conical dipole as a function of the cone angle. The overall length should be 0·73 wavelength at the mid-frequency independent of the cone angle.

The use of radiators of low Z_0 is to be recommended wherever possible in arrays of full-wave radiators. In much the same way as in the case of folded dipoles as $\lambda/2$ radiators, the bandwidth is markedly improved by using fan elements for λ radiators. In order to compensate for the larger diameter, the overall length for resonance must be reduced to something approaching 0·8 λ, being shorter for the lower L/d ratios.

Tapped Aerial Matching: The Delta Match

The input impedance of any simple wire aerial varies from a low value, approximately equal to the radiation resistance, at points of maximum current, up to a very high value at points of maximum voltage. It is therefore possible to find a point along the wire which can present the correct load impedance to a feed line, provided that the line impedance is greater than the radiation resistance of the aerial. This is called the *delta match*, and is a convenient way of connecting an open wire balanced line to a simple halfwave dipole. The arrangement is shown in **Fig. 13.39** the precise dimensions A and B of the delta section are obtained experimentally, and for the case of a 600 ohm feed line are given by:

$$A = \frac{118}{f(\text{Mc/s})} \qquad B = \frac{148}{f(\text{Mc/s})}$$

where A, B are measured in feet.

The delta match is difficult to adjust in practice to finalize optimum values for the dimensions, and there is also a tendency for the delta section to radiate since the wires carrying equal and opposite currents are not sufficiently close throughout their length for the radiation fields from each to cancel.

The Tee and Gamma Match

The principle of the folded dipole, in which the feed current divides between the driven and undriven conductors of the radiator, is adapted to provide a matching device known as the *tee match*, for connection to a balanced line, and its further development, the *gamma match* which combines the impedance transformation with an unbalanced connection suitable for the direct feed of balanced aerials from coaxial lines.

Since the tee section forms with the aerial conductor a short circuit stub less than $\lambda/4$ long, it will present to the feed line an inductive reactance which must be tuned out to leave only the transformed aerial impedance. This can be achieved by the use of a series capacitor as shown in **Fig. 13.40** This capacitor will require to be adjusted empirically for minimum s.w.r. on the main feed line: it will also need to be sealed adequately against weather and dirt accumulation, which would otherwise make it liable to voltage breakdown, when

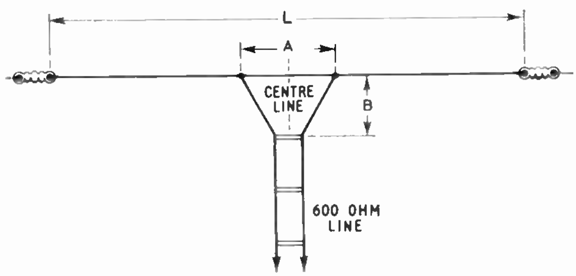


Fig. 13.39. Use the delta match with a half-wave aerial fed with 600 ohm open line.

the aerial is powered. Its physical size and rating should be adequate to withstand the full line voltage with an adequate factor of safety.

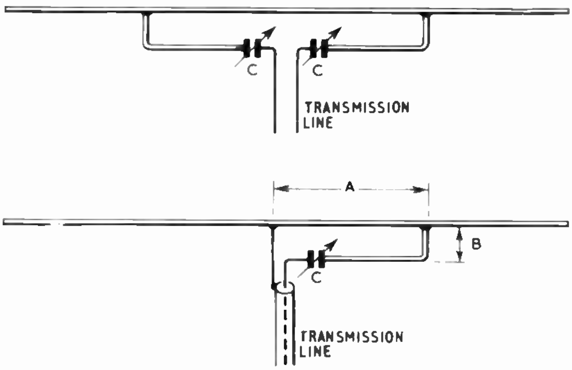


Fig. 13.40. Tuning out residual reactance in the Tee and gamma-match systems with series capacitors. Capacitors with a maximum capacitance of 150pF are suitable at 14 Mc/s. For higher frequencies proportionately smaller capacitances may be used.

As in the delta match, the exact dimensions of the matching section are a matter of experimental adjustment for minimum s.w.r. on the main feed line. For a halfwave dipole fed with 600 ohm line, the approximate dimensions are given by:

$$A = \frac{180 \cdot 5}{f(\text{Mc/s})} \qquad B = \frac{114}{f(\text{Mc/s})}$$

where A is in feet and B in inches.

These formulae apply only when the extra conductor is of the same diameter as the aerial conductor. If a different size of conductor is used for the matching section, and impedance transformation is obtained analogous to that achieved by varying the diameter ratio in a folded dipole. This is useful in parasitic type beam aerials, when the very low input impedance of the driven element can be stepped up to match straight into a 75 ohm coaxial line by employing a gamma matching rod much thinner than the actual driven element tube.

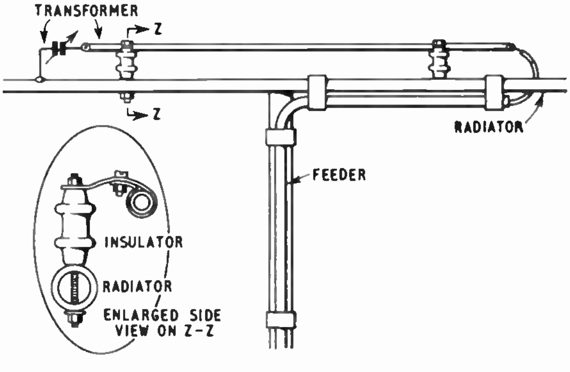


Fig. 13.41. The Clemens transformer, a flexible arrangement which will match a wide range of dipole impedances into a concentric 50 or 70 ohm line, and at the same time convert from balanced-to-unbalanced circuits.

The Clemens Match

A principal disadvantage of the gamma match is the requirement to adjust the length of the matching section to obtain optimum performance. An improved version was developed which utilizes a fixed dimension for the tapping point. It is known as the *Clemens match* and is illustrated in **Fig. 13.41.**

This arrangement is the best, and will cater for a very wide range of impedance transformations. The coaxial feeder cable is bonded to the centre " neutral " point of the aerial, and then carried along one side to a distance of $0{\cdot}06\lambda$ (⅛ of the length of the aerial) where the outer conductor is bonded to the aerial and the inner conductor is joined to a tube of about one-third of the aerial diameter. The tube is spaced a few inches from the aerial and its far end is connected to the aerial through a capacitance. The spacing of the tube, varying from $0{\cdot}01$ to $0{\cdot}02\lambda$, adjusts the ratio, and the capacitor tunes out the transformer reactance and also helps with the tuning of the radiator. The capacitance will not exceed 50 pF at 14 Mc/s or higher frequencies. For the purpose of making adjustments, a variable capacitor may be used, being finally replaced by a high-voltage foil/mica capacitor of the required value, or by a short length of open-circuit concentric r.f. cable, reckoning 70 ohm cable as 22 pF per foot, and 50 ohm cable as 30 pF per foot. The cable can be tucked inside the transformer tubing, and should be sealed against damp, Bostik sealing strip or Sylglas tape both being very reliable for the purpose.

Linear Transformers

The most convenient way of performing impedance matching between the aerial and the main feeder is by employing *linear transformers.* These are made up from short sections of transmission line and are hence physically convenient to assemble and connect at the aerial terminals. They are also lighter than the equivalent coil and capacitor networks and more dependable when exposed to the weather.

The behaviour of short sections of transmission line with various terminations was explained on page 13.16 where it was shown that quarter wave sections can be considered equivalent to parallel and series tuned circuits. Shorter line sections behave like inductance or capacitance depending upon their length and whether they are open or short circuit at the far end.

Stub Matching

If the main feeder is connected directly to the aerial terminals the aerial will present other than a matched load, and a system of standing waves will be set up along the feed line. Moving away from the aerial, the impedance along the line will alter in sympathy with the pattern of voltage and current waves, as explained and illustrated in Fig. 13.19 There will occur a position along the line *not more than* $\lambda/2$ *from the aerial* at which the resistive component of the impedance will equal the line characteristic impedance, but will have associated with it some degree of reactance. By connecting in parallel with the main line at that point a transmission line stub, which is either open or short circuited at the far end to give an equal but opposite reactance at its input end, the unwanted reactance on the main line can be cancelled or " tuned out," and the only term left is the resistive one which then presents a correct match to the remainder of the line. This is the principle of *stub matching*.

It is possible to calculate the length and position of the stub from a knowledge of the s.w.r. on the un-matched line and the position of a voltage maximum or minimum. The charts shown in **Fig. 13.42** and **13.43** give the length of the stub and its position from the point of maximum or minimum voltage *in the direction of the transmitter.* They are drawn specifically in terms of a stub of the same impedance Z_{01} as the main feed line. However a line of different impedance Z_{02} may be used to provide the stub, by determining the ratio $\frac{X}{Z_{01}}$ from Fig. 13.27, and thence X for the stub length given by the stub-matching charts. The new ratio $\frac{X}{Z_{02}}$ may then be used to find a new stub length again from Fig. 13.27. It may be advantageous to use a stub of lower Z_0 than the main feed line when the latter is relatively high, since the bandwidth of the match will then be improved.

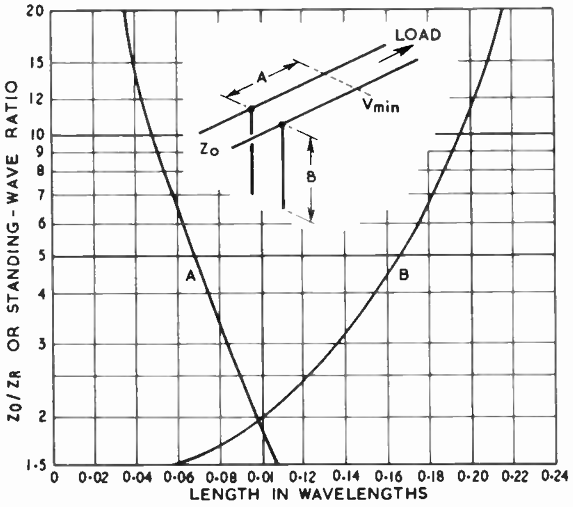


Fig. 13.42. Position and length of an open stub as a function of Z_0/Z_R or s.w.r.

The curves of Fig. 13.42 apply when the point of reference is a voltage minimum, and those of Fig. 13.43 when the reference is a voltage maximum. Since these voltage conditions along the line alternate at intervals of $\lambda/4$, either short circuit or open circuit stubs may be employed at the appropriate points on the line to achieve exactly the same match on the remainder of the line. Also because of the inverting properties of a quarterwave section of line, the open circuit stub may be replaced by a short circuit stub which is $\lambda/4$ longer, and vice versa. This system of stub matching is extremely flexible, and within certain limits permits the matching to be carried out at a point which is physically convenient, and with a stub which can be either open or short circuit to suit the particular mechanical layout. Generally speaking the use of short circuit stubs is preferable. The electric field at the remote end of an open circuit stub is very high, and the "end effects" of the supporting insulator are difficult to assess accurately. Their action is to modify the electrical length of the stub and hence to "de-tune" it. In a short circuit stub,

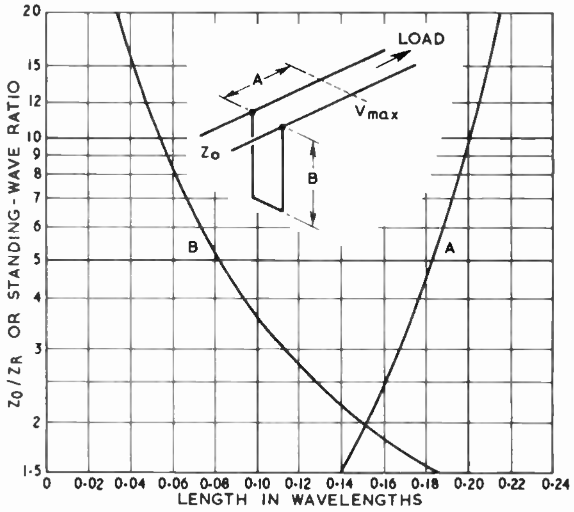


Fig. 13.43. Position and length of a closed stub as a function of Z_R/Z_0 or s.w.r.

the remote end is accurately defined by the shorting bar and the length is not subject to the same end effects.

The voltage at the end is practically zero, and they can be tensioned or otherwise supported without the need for a strain insulator. There is also no objection to a d.c. connection from the centre of the shorting bar to earth, to provide lightning protection for the aerial and feeder system.

Another useful feature of stub matching is that it is not necessary for the aerial to be resonant, i.e., cut to an exact number of quarterwavelengths at the operating frequency. If the aerial is of an arbitrary length, the input impedance will be complex, and the series reactance necessary to tune the aerial is effectively provided by the short piece of line between the aerial and the first voltage maximum or minimum.

Quarterwave Resonant Sections

When the standing wave ratio is high, the length of the stub together with its distance from the reference point add up to one quarterwavelength, and because of this, such stubs are sometimes referred to as quarterwave resonant line transformers. A typical stub matching arrangement is shown in **Fig. 13.44(a)** and re-arranged in **Fig. 13.44(b)** to illustrate the quarterwave feature of this system of matching. The action is electrically identical, but it may sometimes be physically more convenient to mount a fixed $\lambda/4$ stub across the aerial input, and to adjust the tapping point of the main feeder along this stub for minimum s.w.r. on the main line.

Quarterwave resonant line transformers can be used when the impedance ratio is fairly high (e.g., greater than five-to-one) and two varieties using short- and open-circuited stubs are shown. Both make use of the fact that there is a standing wave on the stubs, the main feeder being moved along the stub to a point where the voltage/current ratio becomes equal to the main line impedance. They are effectively tuned resonant circuits, with the feeder tapped into them. The open stub, is employed when the feed point impedance of the aerial is too low for the line, i.e., when the overall aerial length is an odd number of half-waves, and the closed stub is used for a high impedance feed, i.e., when the aerial is a number of full waves long.

In practice the stub can be made to have the same impedance as the main line, though this is not essential. The line should be supported so that it remains at a fixed angle (preferably a right angle) to the stub. The tapping position is varied until the standing wave or the reflection coefficient on the main line is minimized. If a good match is not produced the length of the stub can be altered; should this not be successful, it is probable that the wrong type of stub is being used. It may be noted that the aerial itself need not be a resonant length, though the aerial-plus-stub must be; also, the stubs may be increased in length by one or more quarter-waves in order to bring the tap position nearer to ground level, changing from " open " to " short-circuit " or vice-versa with each quarterwavelength addition.

The technique of stub matching is applicable both to balanced and unbalanced lines, and may be used with advantage on coaxial lines at v.h.f. where the physical length of the stub is not great. Again because of the " end effect " of the open circuit stub, a short circuit version is to be preferred, which then completes the outer screen of the coaxial system and eliminates the possibility of unwanted signals being picked up on the inner conductor of an open circuit stub.

Series Quarterwave Transformers

The impedance inverting properties of a $\lambda/4$ section of line were explained earlier in the section dealing with transmission

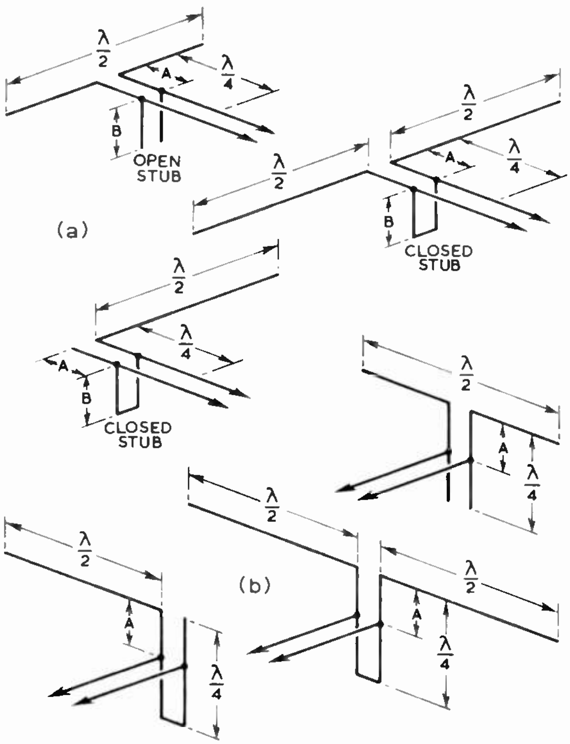


Fig. 13.44. Use of matching stubs with various aerials.

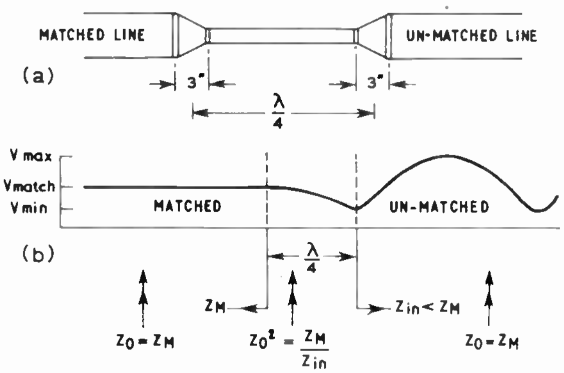


Fig. 13.45. Series quarter-wave transformer (Q-bar) using "pinch" matching. The voltage standing wave is a minimum at the point of commencing the quarter-wave matching section, which is pinched together until the standing wave disappears on the following length of line. The "pinch" can be determined experimentally using string to hold the wires together. The string can then be replaced by permanent insulators.

lines. It is therefore possible to match two resistive loads by the insertion of a series $\lambda/4$ section of line of an appropriate Z_0. In the description of stub matching it was explained how the arbitrary impedance of the aerial sets up a standing wave pattern on the main feed line. Moving along the line from the aerial, a position of voltage minimum is located. At this point the impedance will be purely resistive and lower in value than the line characteristic impedance. By pinching the line conductors together to reduce the spacing over the following quarterwavelength of main line, the Z_0 of that section can be adjusted to the correct value for the section to act as a $\lambda/4$ transformer and step up the low impedance at the voltage minimum point to the correct value to terminate the remainder of the line at its normal spacing, **Fig. 13.45.** The $\lambda/4$ section of modified Z is often called a *Q-bar transformer*. In an open wire system it is physically convenient to draw the conductors together over the required $\lambda/4$ section of line, experimentally adjusting the spacing for minimum s.w.r. on the main line. This is facilitated by the fact that the position of the $\lambda/4$ section is fixed by the location of the point of voltage minimum on the line, which is itself determined only by the actual aerial impedance. When the optimum spacing has been determined, permanent rigid spacers of the required length may be fitted throughout the transformer section to maintain the correct Z_0.

The bandwidth of a series quarterwave transformer match is better than that to be obtained from an equivalent stub match, and can be made even better by the use of cascaded quarterwave sections as explained on page 13.16.

The series transformer technique can also be applied to coaxial line systems, but it is then in general less flexible than the stub match. The range of commercially available characteristic impedances available for the transformer sections is limited to one or two discrete values, and the length of such sections at h.f. precludes the fabrication of a special section of line. The technique is however more amenable at v.h.f. when the physical lengths are short, and self-supporting. $\lambda/4$ sections of rigid line may be readily constructed to a wide range of impedance at will.

Tapered Lines

When the required transformer ratio is not high, for example, when an 800 ohm rhombic aerial is to be matched to a 600 or 300 ohm line, it is practicable to use a simple tapered line. At the aerial end, the spacing is correct for 800 ohms and gradually decreases to that for the lower impedance. This arrangement gradually converts the impedance from one value to the other. It is a wide-band device and, provided the taper is not less than one wavelength long, it will give a good, though not perfect, match.

Network Transformers

At ground level, and particularly where a concentric (coaxial) feed line is used, it is often more convenient to employ a coil and capacitor network. This could be similar to any of the couplers used near the transmitter (though a simpler network).

A frequent requirement in radio frequency circuits is the ability to transform some particular value of resistance to some other value, without introducing a significant power loss in the process. An example of this is the matching of the effective r.f. anode load of a valve into the grid of the succeeding stage. One of the easiest ways of achieving this

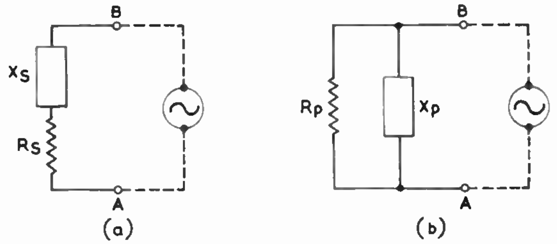


Fig. 13.46. Equivalent impedance circuits.

transformation is by the use of inductance and capacitance in a single network, using entirely reactive elements in the transformer. Clearly if such elements are perfect, then the power losses will be zero and the transformation will occur at unity efficiency. Since in practice, radio frequency inductors and capacitors can be made to a relatively high quality or high-Q, the losses in a practical arrangement are quite small, resulting only from the dissipation in the d.c. resistance of the coil due to the passage of current, and the dielectric loss in the capacitor due to the leakage current, which in turn is proportional to the voltage across it.

Consider two elementary networks in series and shunt elements respectively as shown in **Fig. 13.46.** Now for these two networks to be identical, that is to present the same load to a generator connected across the terminals AB, the admittance seen at these terminals must be the same in each case.

$$\frac{1}{Rp} + \frac{1}{jXp} = \frac{1}{Rs + jXs} \qquad \text{....(i)}$$

From this it is simple to derive the following two identities which are the key to all network problems:

$$Rp = Rs\left(1 + \frac{Xs^2}{Rs^2}\right) \qquad \text{....(ii)}$$

$$Xp = Xs\left(1 + \frac{Rs^2}{Xs^2}\right) \qquad \text{....(iii)}$$

Using these basic formulae it is possible to convert a network of series elements to its equivalent network of parallel elements (and conversely). Consider next the circuit shown in **Fig. 13.47.** From an inspection of Fig. 13.46. and equation (ii) above, it can be seen that it is possible to select a value of Xs to put in series with the resistance Rs, such that the shunt equivalent resistance is equal to Rp, but has in parallel with it a residual reactance Xp. If now a reactance of value $-Xp$ is placed across the terminals AB, the net effect will be to produce at those terminals a pure resistance equal to Rp, which is the requirement of the transformer. Then the basic form of the transformer is an L network comprising only reactive elements as shown in **Fig. 13.48.** This also meets the terms of the original specification. If the ratio $R_2/R_1 = p$, then it can be shown that:

$$X_1 = \pm R_1\sqrt{(p-1)} \qquad \text{....(iv)}$$

$$X_2 = \mp pR_1/\sqrt{(p-1)} \qquad \text{....(v)}$$

Then X_1 and X_2 must have opposite signs in all cases, or in more general terms, the network must consist of one inductive and one capacitive element. It is more usual in practice to make the series element inductive and the shunt element capacitive for reasons which vary with individual circumstances (e.g. a series d.c. path to the anode of a valve), but there is no basic objection to the reversal of this practice in cases where it may be more suitable.

Using the fundamental equations (iv) and (v) it is now

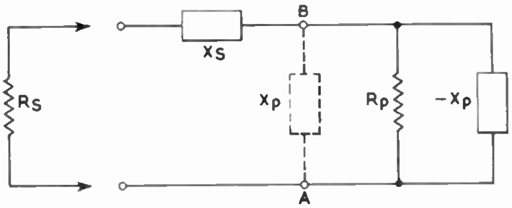


Fig. 13.47. Transformation : stage 1.

possible to design any required network by a process of arithmetic as the following example will show:

Example

A single wire short aerial for 3·6 Mc/s has a base impedance of 15-j200 ohms. It is required to match this to a feeder with a characteristic impedance of 75 ohms (**Fig. 13.49. (a)**). This aerial is equivalent at its terminals to a resistance of 15 ohm in series with a capacitance of 220 pF. The steps of the calculation are as follows:

(a) Connect an equal and opposite reactance +j200 in series to tune out the aerial capacity and leave only the resistive term of 15 ohms: the transformation required is then from 15 ohms resistive to 75 ohms resistive (**Fig. 13.49. (b)**).

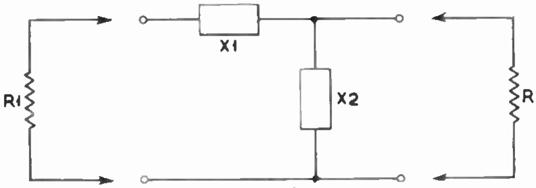


Fig. 13.48. Transformation : stage 2.

(b) From equation (iv), X_1 (series term) $= \pm 15\sqrt{5-1}$
$= \pm 30$ ohm

(c) From equation (iv), X_2 (shunt term) $= \mp \dfrac{5 \times 15}{\sqrt{5-1}}$
$= \mp 37{\cdot}5$ ohm

Thus the transformer section required is as shown in **Fig. 13.49. (c)** and the complete arrangement in **Fig. 13.49 (d).** To convert the reactance into physical components, since $X_L = 2\pi fL$ and $X_C = 1/2\pi fC$:

$+j230$ is the reactance of 10 microhenries at 3·6 Mc/s.
$-j37{\cdot}5$ is the reactance of 1250 pF at 3·6 Mc/s.

The complete network is then shown in **Fig. 13.49 (e).** By making each element variable over a convenient range, it is possible to correct for any inaccuracies in the original figure taken for the aerial base impedance, and also for variations in the working frequency.

A suitable protective cover for the network is a large air-tight can, used upside down, with components mounted on the lid. A metallized lead-through seal from an old capacitor or transformer can be soldered into the lid for the aerial lead. A sealed r.f. connector is preferable for the cable connection. The can should have a coat of paint, and Bostik putty can be used to seal all the joints; a small bag of silica-gel may be placed in the can to guard against condensation. The components of the network should be suitable for the transmitter power; if fixed capacitors are used they should be of the foil-and-mica type.

Baluns

The great majority of aerials are inherently balanced devices, such that equal voltages exist to earth from each input terminal. The exceptions are the electrically short aerials often used on the lower frequency bands, verticals which are driven against either a ground plane or true earth, and aerials fed with gamma matching.

It is not possible to connect to a balanced matching aerial an unbalanced feeder such as coaxial line and maintain zero potential on the outside of the line. In such cases currents are forced to flow down the *outside* of the outer conductor by the voltages appearing at the aerial terminal to which it is connected. These currents give rise to unwanted radiation from the line itself since the field due to them cannot be cancelled out by the field due to the current flowing in the inner conductor, which is contained entirely within the outer conductor of the line and cannot penetrate beyond it.

The existence of the unwanted radiation field around the feeder will modify the pattern of the aerial, and possibly also its input impedance due to the coupling now existing between feeder and aerial, and will represent a loss of energy in undesired directions. As with an inadequately balanced twin wire line, it is also a potential source of TVI, BCI and r.f. feedback problems within the station itself.

If it is required to drive a balanced aerial directly from an unbalanced line, then some form of choke must be employed to prevent the unwanted currents flowing back down the outside of the line. The necessary choking action is obtained by incorporating a balance-to-unbalance transformer, often called a *balun*. The action of transmission line baluns depends principally upon the use of a quarterwave section of line as a parallel resonant circuit to present a high impedance path to the unwanted currents. A number of balun

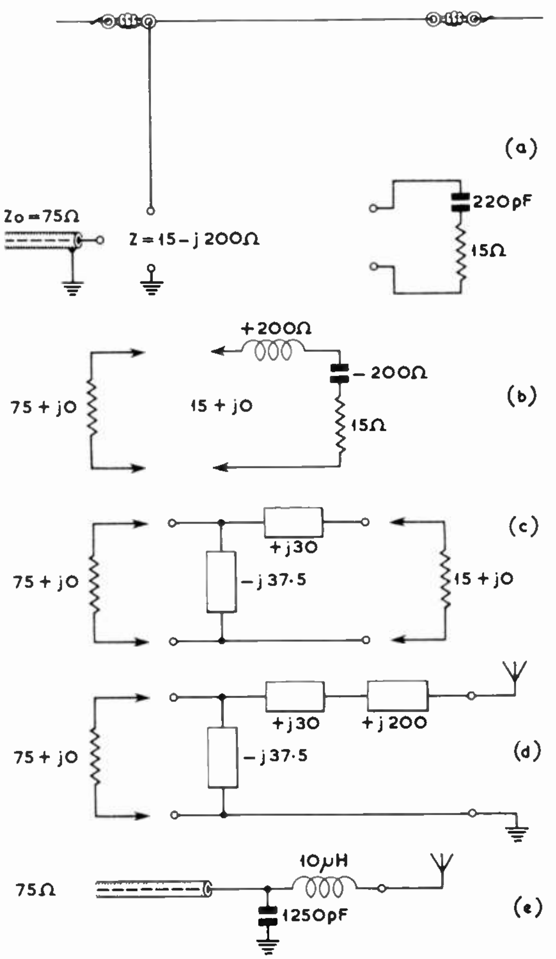


Fig. 13.49. Matching of two arbitrary impedances.

designs based on this principle are described in Chapter 14, and although they will all work in theory at h.f. as well, their physical length precludes their convenient use below the v.h.f. bands. At the longer wavelengths the most convenient form for a balun is provided by a pair of bifilar wound choke coils connected as shown in **Fig. 13.50.** These coils may be sealed up in a suitable weatherproof housing at the aerial feed point to provide a convenient centre insulator for the aerial, with a coaxial input for unbalanced feed.

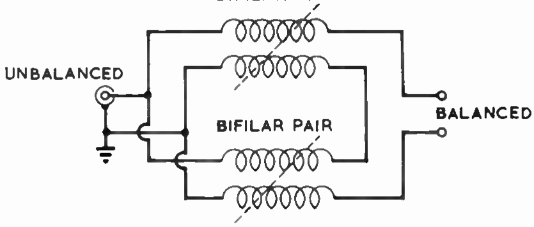


Fig. 13.50. Pairing of bifilar choke coils.

A suitable design for powers up to 1kW on all the h.f. bands has been developed by G3HZP. It is a one-to-one ratio bifilar wound transformer with a toroidal ferrite core, exhibiting an impedance of 65 ohms ± 15 ohms, an insertion loss of less than 0·1db, and a balance better than 10 per cent over the frequency range 1·5 Mc/s to 30 Mc/s. The impedance varies slightly over the frequency range, but this is of no consequence, and the balun is equally effective with 50 or 80 ohms coaxial cable. If desired, the turns ratio can be varied, thereby introducing an impedance transformation property.

Construction of the Toroidal Balun

Two Mullard FX1588 toroidal cores are stacked, and bound with acetate, polyester or silk tape before the primary and secondary coils are wound. The primary, L1, consists of 10 turns, while the secondary comprises two windings of five turns connected as shown in **Fig. 13.51.** The wire used is 0·110 in. × 0·060 in. enamelled copper tape, 20 in. being required for L1, and 13½ in. for each section of L2. When the winding is completed, the balun can be completely enveloped for protection with one of the following materials: bitumen, Chatterton's compound, several coats of tropical varnish or polyurethane varnish of the type supplied for small boats or yachts, fibreglass (Holts car kit), Araldite, or solidifying silicone grease (EP5555, ICI Ltd.). Alternatively, it can be encapsulated in a small hermetically sealed metal box.

Three methods of mounting have been found suitable. Perhaps the simplest method is to attach the toroid to a triangular sheet of Perspex, and completely envelop the assembly in Araldite. The coaxial line can be readily secured with a sheet metal clamp bolted to the Perspex; the toroid is held in position by the combined action of the mounting wires and Araldite. The dimensions and drilling points are shown in **Fig. 13.52**; the inner holes X retain the balun lead-

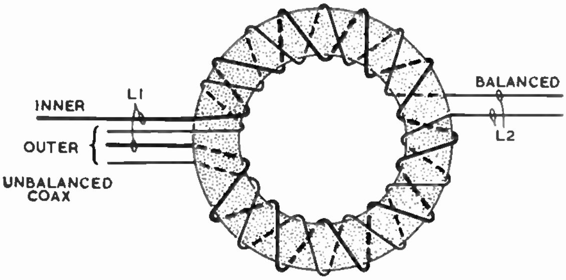


Fig. 13.51. The construction of the windings on the toroidal core. L1 consists of 10 turns, and L2 is formed with 5 + 5 turns.

out wires, and the holes Y secure the aerial wires. The jumper between holes X and Y should preferably be a length of copper braid to prevent damage owing to flexing of the assembly in high winds.

A second fairly simple method of mounting is to enclose the balun in a short length of Marley plastic drainpipe, the ends being covered with discs of paxolin. The discs can be cut with a fly-cutter, the pilot holes being used to allow a brass screw to clamp the ends together, or alternatively used as fixing points to tie the unit to a mast. The coaxial cable

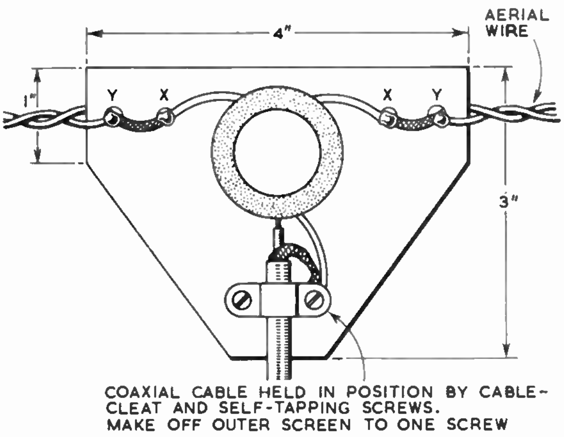
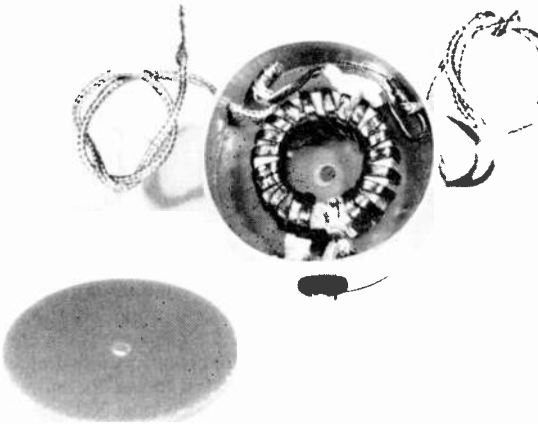


Fig. 13.52. A triangular block of perspex makes a strong mounting for a balun positioned at the centre of a half-wave dipole. The balun is anchored with its connecting wires and a coating of Araldite, the coax is held with a cable cleat, and the aerial wires are passed through holes Y. Braid is used to connect the output of the balun to the aerial wires to avoid straining the connections owing to flexing in the wind.

is held tight by a rubber grommet in the centre of the tube wall, while the balanced winding can either be connected to the dipole by braid fly leads, or alternatively a small strain insulator cut from Perspex sheet can be mounted within the unit; this possesses the added advantage that all joints are totally enclosed. The unit should be sealed with Evo-stik or Twinpack Araldite.

The third, more complex, system of mounting the balun is to seal it hermetically within a metal box. Suitable measurements are 2 in. × 1½ in. × 2½ in., with the type 83 coaxial socket mounted on the 2 in. × 1½ in. dimension. Glass feed-through insulators are used for the aerial connections. Unless a suitable box is available, one can be fashioned without too much difficulty from tinplate and the joints soldered.

The completed balun. The two right-hand leads are the balanced aerial connections (L2); the other end of each of these windings is connected to the earthy end of L1.



An inexpensive mount using a short length of plastic drainpipe.

METHODS OF FEEDING AN AERIAL

End Feeding

The simplest way to energize an aerial is to bring one end of it into the station and connect it to earth through a tuned circuit, link coupled to the transmitter.

This method is strongly to be deprecated for a number of reasons. The impedance seen at the end of the aerial wire will in general be of some unknown value, and unless some form of adjustment of the coupling between the link and the p.a. coil is provided, together with adjustment of reactance in the p.a. tank circuit, it will be impossible to achieve anything like an efficient match between aerial and transmitter, with a consequent inability to draw power into the aerial. Additionally there will be no discrimination against the coupling into the aerial of the unwanted harmonic currents flowing in the p.a. tank circuit: the aerial will undoubtedly respond to these in a harmonic mode and radiate them as a source of interference. Finally, the need to bring the aerial wire right into the station to make the connection to the transmitter means that there will exist in the room and its immediate environs high radiated fields. These will give rise to feedback problems in audio and r.f. circuits, and will contribute nothing to the useful radiation, being generally absorbed in the fabric of the building in which the station is housed.

Modern transmitter design tends to use a pi-circuit for the p.a. tank, designed to work into an output impedance within a limited range of low impedances, typically 40–80 ohms. Since the end impedance of a current fed aerial, i.e. one which is an odd number of quarter wavelengths long, is also low, and of the same order as the figures quoted, it would be quite possible to couple the aerial straight into the output socket of the transmitter and achieve a reasonable match by means of adjustments to the variables in the p.a. tank circuit. For precisely the same reasons as above, this technique is also strongly to be deprecated, tempting as it may be to the amateur beginning on 1·8 Mc/s with a simple transmitter and a short wire aerial. It is perhaps permissible in field day conditions, when the transmitter is in the open, and can literally be connected between the aerial and earth without degrading the radiation characteristics of the aerial. In *all* other circumstances, a separate tuned circuit should be employed as an aerial coupling network, and located at a place suitable to the aerial in use. This network may then be link coupled to the transmitter tank circuit by a low impedance coaxial line, to achieve adequate separation between the radiating part of the aerial and the building housing the station.

If the total aerial length is such as to provide a high impedance load the aerial section nearest to the transmitter will be carrying a high r.f. voltage, and there may be considerable loss of power should it have to pass through or near to surrounding structures. On the other hand, an aerial tuned to an odd number of quarter wavelengths and consequently of low input impedance, can also be inefficient as a result of the comparatively high resistance of the earth connection. It is better therefore to use an intermediate length, for example, an odd number of eighth wavelengths where the impedance will have some intermediate value and therefore a moderate current at the earth connection.

A suitable coupling circuit arrangement is illustrated in **Fig. 13.53**. The inductor and capacitor should be similar to those in the transmitter output tank circuit. The link coil may vary from about four turns at 3·5 Mc/s to one turn at 28 Mc/s and should be tightly coupled into the earthed end of the transmitter tank coil, the aerial tap then being adjusted for correct loading of the transmitter, with moderately broad resonance of the aerial coupler. The chart of **Fig. 13.54** indicates whether the aerial tap should be low on the coil (low impedance) or nearer the top or "hot" end (high impedance).

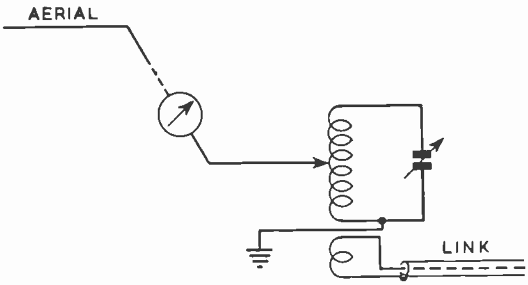


Fig. 13.53. Aerial coupler for end-fed aerials.

With a non-resonant length of aerial it may occasionally be found that, as the tap position is moved up the coil (from the "earthy" end) in order to draw more power, the coupler circuit will be thrown out of tune. This problem is considered later under Aerial Couplers.

Earth Connections

The earth connection should preferably be taken to a copper spike or tube several feet long, or to the nearest large earthed conductor such as a water main pipe. The joint should be heavily coated with bituminous paint to avoid corrosion. *The earth connection of the electricity supply is dangerous* and may introduce noise into the receiver or be responsible for spreading interference to nearby television receivers. The transmitter earth connection should be separate from the safety earth for the equipment in the station, as it is part of the aerial system, carrying r.f. power. If the station is at the top of a building, the aerial coupler may fail to function properly on some frequencies because the earth

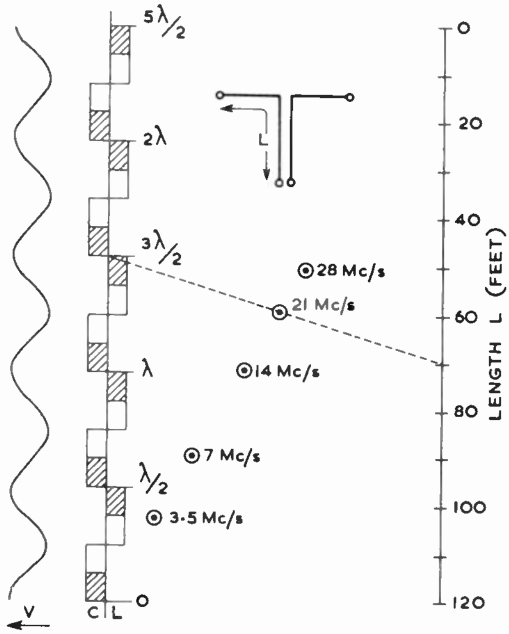


Fig. 13.54. Standing wave chart for tuned feeders. A line through the length *L* of feeder plus half-aerial, and through the frequency point will show, on the wavelength scale, the nature of the input impedance. Rectangles to the left of the line are regions of capacitive impedance, those to the right inductive. The shaded areas correspond to high impedance input (high voltage) and call for parallel-tuned couplers (Fig. 13.55 (b)) and the blank areas to low impedances (high current) which require close taps on a parallel-tuned coupler, or a series circuit (Fig. 13.55 (a)). The chart may also be used for feeders alone and, as coded, applies for the use of high resistance loads. When the load is lower than Z_0 the code must be reversed; the shaded areas indicate low input impedance, inductive to the left. The velocity factor of the feeder is not allowed for in the chart, and the physical length of the feeder must be divided by this factor when computing length *L*.

lead is long enough to develop resonance effects. In such cases a "shortening" capacitor of 50–100 pF may be inserted into the r.f. earth lead to de-tune it. This capacitor should be an air spaced type.

Further information on earth connections and their relative importance will be found later in the section dealing with Low Frequency Aerials.

Non-radiating Feeders

In the preceding section, it was explained that advantages could be obtained by using transmission lines to transfer the power from the transmitter to the aerial without loss by radiation. In order to achieve best results, it is also desirable to present to the transmitter output circuit a resistive load of optimum value for maximum power transfer.

Broadly speaking, there are two methods by which the aerial may be coupled to the transmitter. One involves the use of a *flat* or matched line (called flat because there are no standing waves upon it).

The other employs a *tuned* line, which is not matched, and carries upon it an often substantial standing wave. In such a case, use is made of the property of transmission lines by which the load impedance repeats itself at intervals of one half wavelength along the line.

Each of these systems has merit in certain applications, depending largely upon the type of aerial employed, and the input impedance of the aerial. In general, flat lines are used with low impedance aerials, and tuned lines with aerials presenting a relatively high input impedance e.g. full wave dipoles, or with wire aerials intended for multi-band use.

Tuned Lines

In dealing with line losses on page 13.18 it was explained that, for lines with a very low matched loss, the increased loss due to the presence of even severe standing wave ratios is not excessive. From the curves of Fig 13.30, it will be seen that a line with a matched loss of 0·2db will only introduce a further 0·7db when carrying a 10:1 s.w.r., and only 0·3db if the s.w.r. can be held down to 5:1. The typical matched loss of high impedance open wire lines is only a few decibels per mile if well constructed; in practical terms, therefore, a matched loss of 0·2db represents a run of a least 200 ft. or so of such a line. It is quite reasonable to operate such a line under conditions of relatively high s.w.r. or mismatch without greatly impairing its transmission efficiency, and consequently a line of this type may be used to feed a fairly wide range of aerial impedances in an acceptably efficient manner.

The input impedance at the lower end of the line will depend upon the actual load impedance presented at the far end. and upon the length of feeder involved. In general, it will be quite arbitrary and the feeder is employed solely as a means of connecting the aerial to the transmitter without attempting to transform it to any particular value. It is therefore necessary to terminate the tuned line in a matching network which will transform the impedance at the lower end of the line into the optimum load resistance for the transmitter. Such a network is termed an *aerial coupler*, and usually comprises a tuned circuit with a low impedance link winding on the coil. The tuned circuit will require to be of a series or parallel form depending as the impedance at the lower end of the line is lower or higher than the Z_0 of the open wire line. The design of aerial couplers is dealt with later in the chapter.

Without a knowledge of the exact aerial impedance, it is not possible to determine the precise value of the input impedance to the feeder or to determine the values of capacitance and inductance in the coupler. It is however possible readily to assess whether a series or parallel coupler is required, because simple wire resonant aerials are fed at points of either voltage or current maxima and therefore present either a very high (2,000-5,000 ohms) or very low (70-100 ohms) input impedance at their feed point. This was explained on page 13.5. A simple chart may then be prepared, from which a knowledge of the feeder length involved, and the feed of the aerial (current or voltage) can be used to determine at least the nature, if not the precise value of the impedance presented to the coupling circuit. Such a chart is shown in Fig. 13.54; the same chart may also be used if the composite length of feeder and half aerial is known in the case of centre fed aerials. In such cases the problem of voltage or current feed to the aerial is automatically accounted for by the length L.

The impedance presented to the coupler will contain both resistive and reactive terms. The reactive term is accounted for by adjustment of the tuning control on the coupler (usually a variable capacitor); provision must also be made for obtaining the correct transformation of the resistive term.

This is best done by providing a range of tappings on the coil of the coupler unit, in order that the downcoming feeder may be tapped in at a position along the coil best suited to the value of the input resistance. A wide range of adjustment is desirable to accommodate very low impedances on taps near the coil centre, and very high impedances towards the outer ends of the coil. Only when the impedance is very low, corresponding to a current maximum on the line, will it be necessary to use a series coupler.

When the feeder input resistance is of the same range as the actual feeder Z_0, either due to a particular length of feeder in use, or because of the actual aerial impedance at the top, it may not always be possible to tune the feeder with the available adjustments on a series coupler. In such cases, a parallel coupler is to be preferred, and the feeder should be tapped down the coil to obtain the best match to the transmitter.

In general, very high impedances at the bottom of the feeder should be avoided by judicious selection of the length of feeder employed, using the chart of Fig. 13.54. High impedances are often difficult to match into the parallel tuned circuit because of the tendency to " run out of coil " when selecting the correct tappings. The high impedance also associates with it a high voltage which will appear across the components of the tuned circuit and requires disproportionately large components to avoid the danger of flash-over. This is particularly evident between the vanes of the air spaced tuning capacitor during peaks of amplitude modulation, when the vane spacing is inadequate for the high voltages developed. Regions of current maxima on the tuned feeders are not quite so unattractive, although they do result in increased losses in the coupling unit due to the high circulating currents in the coil. The ideal situation is one which, by judicious selection of aerial length and feeder length, provides a medium impedance to the coupler on all bands. In such cases a parallel coupler with a reasonable number of tap adjustments on the coil will be satisfactory.

There are two disadvantages associated with the use of tuned lines for aerial feeders. When considering the variation of input impedance with line length, on page 13.15, it was shown that the rate of change of input impedance with frequency became greater as the length increased. The effect of this is to reduce the bandwidth of the aerial and feeder together, as the line length employed becomes greater, to an extent where it may well become necessary to retune the aerial coupler for changes of frequency within the narrow span of one amateur band. A compromise must therefore be sought between the low loss of the open wire line (even with a high standing wave) and the reduced flexibility of operation necessitated by frequent adjustments to the coupler.

The degree of compromise can only be assessed for each particular installation, but as a general rule it would be desirable to locate the coupler no more than 1·0 to 2·0 wavelengths from the aerial at the highest operating frequency. Because of the restrictive nature of open wire feeder installations, it is in any case necessary to locate a coupler at the point of entry of open-wire line into the building, be it house or garden shed, and to complete the feeder run to the transmitter in low impedance unbalanced line. It is very bad practice to attempt to extend open wire balanced lines inside any form of building, and particularly into the station itself, because of their tendancy to radiate energy if the slightest current unbalance occurs.

The tendency to unbalance is the second disadvantage in employing open wire lines with a high s.w.r. upon them. There will exist along them regions of fairly high current maxima considerably in excess of the normal matched line current, and these will greatly accentuate radiation from the line if the slightest unbalance occurs along the line at those regions. Such radiation, in addition to being a possible source of TVI, BCI, and even r.f. feedback in the shack, is effectively power lost to the aerial, and may be considered an additional source of line loss due to the high s.w.r. The ideal figures of Fig. 13.30 may be increased considerably by this additional power loss due to radiation, to a point where it would be more efficient to use some form of impedance transformer close to the aerial, and a suitable unbalanced matched feeder down to the transmitter. It is therefore particularly important when using open wire balanced feeders as tuned lines with large standing waves, to ensure perfect electrical and mechanical symmetry in the construction and installation of the feeder.

The use of tuned feeders is essential to any aerial system comprising a simple wire radiator intended to work on a number of different bands. Their use in such a context is best illustrated by example.

Centre-fed Horizontal Wire

The wavelength to which a wire aerial tunes is approximately twice its length in metres. Hence an aerial 21 metres long (approximately 66 ft.) tunes to a wavelength of 42 metres corresponding to a frequency of 7 Mc/s. At this frequency it is a half-wave aerial, and, if fed in the centre with a balanced two-wire feeder, is called a half-wave dipole. Such an aerial would be full-wave on 14 Mc/s and three half-waves on 21 Mc/s and so on. On 3·5 Mc/s it would be a short (quarter wave) dipole and its radiation resistance would be about 12 ohms; the wire resistance and insulator loss might contribute several ohms extra resistance, which would consume a large proportion of the transmitting power, thus preventing its useful radiation.

The 7 Mc/s half-wave dipole shown in **Fig. 13.55** can be fed with 70 ohm or 80 ohm twin feeder since its radiation resistance lies between 60 and 90 ohms depending on its height, the match between aerial and feeder being sufficiently accurate for all practical purposes.

On 14 Mc/s the same 66 ft. aerial would be extremely inefficient because the feed portion occurs between two voltage maxima. At this frequency its impedance would be 5000 ohms and the v.s.w.r. on the 80 ohm twin line 5000÷80—about 62. The line would therefore have appreciable loss. On 21 Mc/s the centre would again have a current maximum and an impedance of about 90 ohms, which would be satisfactory for the feeder. At 28 Mc/s a high mismatch would again be presented.

The directional properties of this aerial on the various bands will be considered later; for the present, a way to feed with reasonable efficiency on all the bands must be found. Obviously some compromise is needed, and tuned feeders provide a solution.

The first compromise is to sacrifice the feeder match on 7 and 21 Mc/s in order to reduce the v.s.w.r. on 14 and 28 Mc/s. This can be done by using 600 ohm open wire line, which will work quite efficiently with a fairly high v.s.w.r. The impedance of 600 ohms is a mean value between say 70 ohms

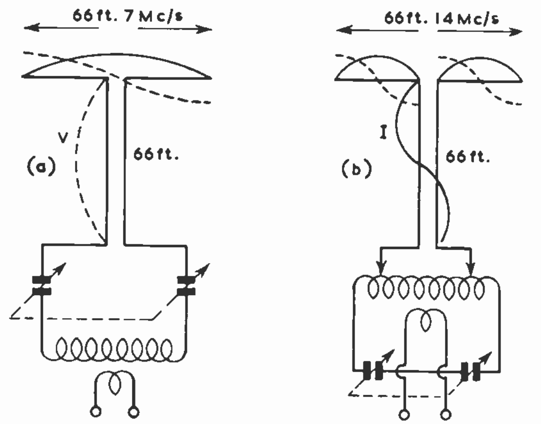


Fig. 13.55. Centre-fed aerials using tuned feeders. The aerial to the left is a half-wave long and has a low input impedance; the feeder is a half-wavelength long and thus repeats the low impedance, so that series tuning could be used. On the right is a similar aerial operating at twice the frequency of the other (full wave). The input impedance is high, and the feeder one wavelength long; parallel tuning would be best. The effect of feeder length is discussed in the text, and illustrated in Fig. 13.26. The tuning capacitor sections should each be 50 pF maximum for 14 Mc/s and higher frequencies; proportionately larger values are needed for lower frequencies. The voltage rating of the capacitors should be the same as for those in the transmitter output tank circuit.

(on 7 Mc/s) and 5,000 ohms (on 14 Mc/s) and the v.s.w.r. will be as high as 9 on these and the other bands.

Fig. 13.55 shows the aerial and its feeder (also 66 ft. long) and indicates the form of the standing waves on both aerial and feeder when used on 7 and 14 Mc/s. Two different coupling circuits are shown for the transmitter.

Since the feeder is 66 ft. long, it will be a definite number of half wavelengths long on all the bands from 7 to 28 Mc/s and the input impedance will therefore be the same as the aerial impedance for each band. On 7 Mc/s for example, the impedance is low and the *series* circuit of Fig. 13.55 (a) would be used, the coil and tuning capacitors having about the same electrical and physical size as the transmitter tank circuit. The link coupling coil for the transmitter feeder connection should consist of three turns of insulated wire between the centre turns of the tuning coil. Alternatively the parallel tuned coupling circuit of Fig. 13.55 (b) could be used, and the feeder tapping points set close to the centre of the coil. Similar arrangements could also be used for 21 Mc/s with a size of coil suitable for that band. On 14 and 28 Mc/s the impedance is high, and in these cases the parallel tuning network should be used because at high impedance the taps will be at the outer ends of the coil.

If the feeder is made only 33 ft. long it will be one quarter wavelength on 7 Mc/s, one half wavelength on 14 Mc/s, three-quarters of a wavelength on 21 Mc/s and a full wavelength (four quarter wavelengths) on 28 Mc/s. The impedance at the input to the feeder will be the same as with the 66 ft. feeder on 14 and 28 Mc/s but on 7 and 21 Mc/s where the aerial impedance is low, the odd quarter wavelength feeder will transform it to a high value on these bands also; thus, the parallel tuned coupler would be used on all bands.

With the aerials described the s.w.r. is high and consequently the coupling circuit will carry large voltages and currents. It is essential therefore to use good quality components. The best arrangement for multi-band working is to employ a plug-in coil base with six contacts, two for the coil, two for the feeder and two for the transmitter link coil with a separate coil unit for each band. The tuning capacitor must have two sections in order to preserve feeder balance, as discussed later under aerial couplers. For 100 watts of r.f. power the feeder current may not exceed about 0·25 amp in each line when the impedance is high, but it may well exceed 1 amp for the low impedance cases, e.g. 66 ft. feeder on 7 and 21 Mc/s. This is a disadvantage in that the same r.f. ammeters cannot be used in all cases, but when the current is low and the voltage high, a small neon lamp may be used instead as a voltage indicator.

Judicious choice of feeder lengths will avoid a high voltage feed point on most bands. A suitable length may be found with the aid of Fig. 13.54. On this standing wave chart the total lengths may (if desired) include one half of the aerial so that the voltage wave illustrated starts from one aerial insulator. It will be seen that total lengths of 45 or 90 ft. bring the feeder input into the high current sections of the chart in nearly every case when using tuned feeders. It is not essential to make the aerial itself a resonant length; in fact the top can advantageously be made a little more than 66 ft. long to avoid a high voltage at the centre of the aerial on even harmonic bands.

Flat Lines

The most important feature of flat lines lies in the fact that the input impedance of the line is equal to the load impedance, and by definition to the characteristic impedance Z_0, *irrespective of the line length.* It follows from this that the route and length of feeder employed may be entirely decided by the physical layout of the aerial system, and its position relative to the transmitter. Because the matched loss of coaxial cables is considerably greater than for open-wire lines, such cables are always employed as nominally flat lines, with a minimum standing wave and hence minimum additional loss. In some cases, particularly at v.h.f. where line loss is important, open-wire lines may be intentionally worked in a matched condition to ensure absolute minimum attenuation of signal.

It was shown in the section on Transmission Lines that a matched line can only be achieved by variation of the load impedance, adjustments being made until it equals the Z_0 of the line. Thus, unlike the tuned line method of aerial feeding, the matching arrangements must be incorporated *at the aerial end* of the flat line to obtain correct operation.

If the line is to be operated in a matched condition right up to the aerial itself, then some means of matching must be incorporated into the actual aerial design, and, if the line employed is of the coaxial or unbalanced type, some form of unbalanced-to-balanced transformer (or *balun*) will also be required at the aerial terminals. Techniques of matching at the aerial itself are discussed in the earlier section on Impedance Matching, which also considers the design of balun transformers.

Because the use of flat lines dictates the need for impedance matching facilities at the aerial end of the line, it is possible to standardize on a characteristic impedance for flat lines for all applications, and arrange to match any particular aerial impedance which is encountered to that standard line impedance. The commonest coaxial line impedances are 75 ohms and 50 ohms. The former is used in Europe for domestic TV purposes: the latter is more common in the American continent, but is being increasingly used for amateur installations in Europe possibly because of the tendency to use commercial equipment of American origin.

It is advisable to develop a station where possible around one or other of these standard values, rather than to attempt the simultaneous use of both. All auxiliary equipment such as TVI filters, s.w.r. indicators etc., can then be constructed to the selected standard impedance, and a greater degree of operational flexibility results without the need for additional impedance transformers. Transmitters and receivers can also be designed to work into this standard impedance, with provision for some small range of adjustment to cover impedance variations due to the residual small standing wave which exists unless the matching on the line is perfect. By arranging for the transmitter to work into a load impedance equal to the Z_0 of the line selected, that line becomes effectively an extension of the transmitter output terminals right out to the point of connection of the aerial. The use of an additional aerial coupler at the transmitter end of the feeder becomes unnecessary, the lower end aerial coupler of the tuned line being exchanged for the upper end aerial matching network of the flat line. This difference in the two methods of aerial feeding is illustrated in **Fig. 13.56** The decision on which method to employ is largely dictated by individual circumstances and the type of aerial to be fed.

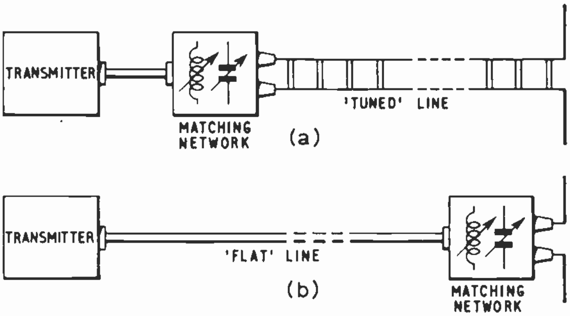


Fig. 13.56. Transmission lines. (a) Tuned line. (b) "Flat" line.

Modern practice tends towards the use of matched unbalanced coaxial line wherever possible, the use of tuned unmatched lines being restricted to cases where a length of such line is an inherent requirement of the aerial design, and is used to convert a nominally high impedance aerial down to a low impedance over a number of bands without the need for re-adjustment of any matching network at the aerial itself. An example of this is the G5RV multi-band dipole described on page 13.72.

AERIAL COUPLERS

It has been assumed in this chapter that the aerial or its feeder will not be connected direct to the transmitter output circuit because this is an obsolete practice not suited to modern conditions. The normal output from a transmitter is through a coupling network and harmonic filter into a co-axial r.f. connector. In some cases, where the aerial is already fed by co-axial lines, it can be connected directly to the transmitter output socket, but more generally it is necessary to use an intermediate coupler or transformer. Even if the line is of the same impedance as the transmitter output (say 70 ohms), it may be a balanced twin feeder and cannot therefore be connected to the unbalanced concentric transmitter output. It is therefore common practice to use an *aerial coupler* or *aerial tuning unit* (often abbreviated to a.t.u.).

The aerial coupler is sometimes built as part of the transmitter but this is not essential and it may be more convenient to extend the output cable from the transmitter to another part of the station, particularly if a change-over arrangement is required to connect the same aerial to the receiver.

The coupler may have to tune the aerial or the feeder to resonance, in addition to transforming from the one impedance to another. Moreover, it may have to take a balanced input and link it to an unbalanced concentric line. The general design of the network has already been illustrated in connection with the simpler aerials and it is now necessary to consider how to choose the right one for each purpose, what value of components to use and the special requirements of balanced feeders. Series tuning was once popular for use with low impedance tuned feeders or aerials, but is a handicap if the same aerial is to be used on several bands, and can usually be avoided by compensation before transforming as described on page 13.32.

Unbalanced Aerials

Fig. 13.57 shows some variations of an aerial coupler suitable for unbalanced aerials or feeders. The general principle is that the inductance acts as a two-winding (or auto) transformer, the capacitor serving to resonate the coil to ensure a high degree of coupling, to help the network to discriminate against unwanted frequencies, and to assist in harmonic suppression. The link may have either a direct tap connection or a separate coil of a few turns interwound with the main inductance at the earthed end.

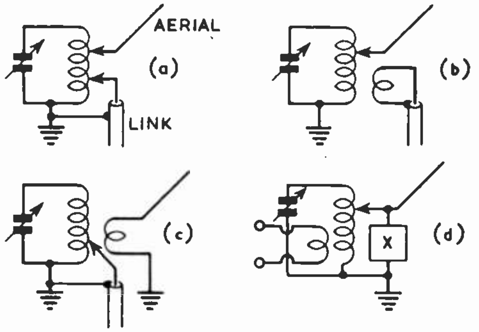


Fig. 13.57. Various forms of aerial coupler for end-fed aerials. The use of reactance X for compensation of very reactive aerials or feeder loads is discussed in the text.

The tuning components should each have a reactance of about 500 ohms at the operating frequency. This may most conveniently be translated into practical figures by reckoning the capacitance as 1 pF per metre of wavelength, and the inductance as 0·25μH per metre. For example, at 7 Mc/s, approximately 40 pF and 12μH could be used. There is nothing critical about these values, though for a very high impedance aerial load (as may occur when using tuned feeders) it is advisable to increase the inductance and use less capacitance. Coils of about 3 in. diameter wound with 14–16 s.w.g. wire as shown in **Fig. 13.58** will handle the power output of a 150 watt amateur transmitter without overheating. The tuning capacitor should be of good quality, have a plate spacing of $\frac{1}{16}$ in. or more and have adequate capacitance to tune to resonance at about half its maximum value. The link coil may be self-supporting inside the end of the tuning coil or alternatively a few turns of well insulated wire supported between the turns of the tuning inductance. The link may vary from four or five turns at 3·5 Mc/s to one turn at 28 Mc/s.

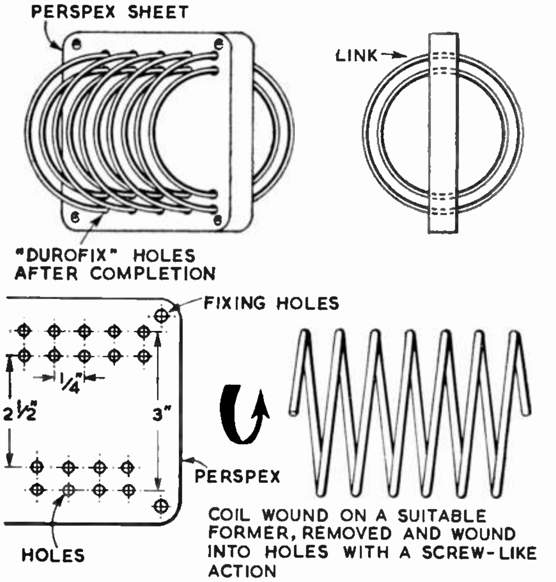


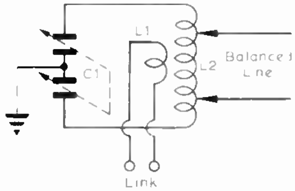
Fig. 13.58. A suitable method of constructing aerial coupler coils. The coils should be 2½ in. and 3 in. in diameter, and the turns spaced ¼ in. using 14 s.w.g. wire. If the link coil is the outer one, turns can be varied easily.

The method of setting up an aerial coupler is first to reduce the link coupling or move the tapping point nearer the earthed end of the coil, and to find the resonant point of the circuit by noting its loading effect on the transmitter anode current. The aerial may next be connected or coupled in and the circuit retuned. Less power will now be drawn, so the link coupling must be increased. The final adjustment is to bring the link to full coupling and the aerial tap or coupling as tight as consistent (high up the coil) with the transmitter still drawing maximum input power. The tuning should be very broad, so that considerable movement of the capacitor is needed to produce significant changes; this is the condition of high efficiency when the network consumes a negligible amount of power. If the aerial is coupled too tightly, it will damp the tuned circuit too much to suit the coupling of the link coil. The transmitter may then draw more anode current, but the aerial current will be less than optimum.

Balanced Circuits

When the aerial or the feeder is balanced it is necessary to use an aerial coupler which is also balanced, such as that shown in **Fig. 13.59** using a split-stator tuning capacitor. The capacitance of each half of the capacitor may be the same as

Fig. 13.59. Aerial coupler suitable for balanced aerials or feeders.



recommended for single-ended couplers but since the two are in series, the inductance value must be twice as big. The aerial or feeder should be tapped symmetrically about the centre of the coil with the link coil in the centre. The adjustment procedure is the same as for unbalanced lines, but in addition it is necessary to ensure that the feeder is working as a balanced system if the aerial is to operate correctly. Unbalance may be brought about by stray capacitance in the coupler.

There are certain points which should be borne in mind with such a circuit. It is usually desirable to use a low value of Q in order to reduce the need for circuit adjustment when changing frequency within a band, and to make the coupling between L1 and L2 variable although this is often difficult to achieve mechanically.

Where the inductive coupling between L1 and L2 is fixed it will be necessary to change the position of the taps and the value of C1.

The L/C ratio in the circuit L2/C1 is not very critical because by adjustment of the taps the value of the Q may be changed.

The values of L2 and C1 can be obtained by plotting the reactance (inductive) against the inductance and the reactance (capacitive) against capacitance, for the different amateur bands.

A reactance of around 500 ohms is usually satisfactory for the centre of the band for L2 and a similar value for C1 to bring the circuit to resonance.

The following table gives suitable values for L2 and C1 for the various H.F. Amateur Bands

Band (Mc/s.)	L2 (microhenries)	C1 (picofarads)
3·5	23	90
7·0	12	44
14·0	6	22
21·0	3·9	15
28·0	2·8	12

Note: The values for C1 are the actual value needed, the maximum value of the variable capacitor C1 should be a little higher than the figures quoted.

The ideal value for L1 is an inductance that results in a reactance equal to impedance of the line to the transmitter. For the purpose of construction it will be found that a value of about one tenth of L2 will be about optimum for L1 when the line to the transmitter is 50 ohms. Where the line is 75 ohms the value calculated for a 50 ohm line should be multiplied by 1·4.

It has already been stated that the coupling between L1 and L2 should ideally be variable but where this is not possible it will be found that the positioning of the taps for correct matching becomes quite critical.

Generally speaking transmitting type coils and capacitors should be used in L1. L2 and C1 i.e. similar to those used in the P.A. Tank Circuit.

It is desirable to use a split stator condensor in the C1 position, the rotor of which may be grounded or left in the air.

The Marconi Effect

If the total distance from the coupler to one end of the aerial is one half wavelength, there is a tendency for the two wires of the feeder to act in parallel, and produce a *T*-aerial effect. The transmitter end of the equivalent parallel system has a high impedance to earth: thus a small stray capacitance to a live part of the network may feed considerable power into this parallel or " phantom " circuit and the feeder can of course, radiate this power. The visible effect is that the feeder wires show unequal currents or voltages. This behaviour of the feeder is often called the *Marconi Effect*.

Resonant lengths of feeder should be avoided if possible but the following points may be noted. Earthing of the frame of the split-stator capacitor usually cures the unbalance, but if this fails it may be advisable to use a separate coupling coil for the feeder (instead of tapping it on to the tuning coil) wound either over, or in the centre of, the tuning coil. The centre point of the extra coil may be earthed direct or through a 200 ohm 2 watt resistor to damp the " phantom " circuit. In any case, at the first sign of serious unbalance (e.g., unusual currents or voltages on the line) the aerial system should be inspected to locate a possible fault; the presence of excessive harmonic output from the transmitter may also be suspected as a cause of apparent unbalance.

Tuning Compensation

Coupling the aerial or feeder to the aerial tuning unit sometimes makes it impossible to adjust the circuit for reasonably high loading. This occurs with tuned feeders with a high standing wave ratio when a heavy reactance load may be presented by the line. In such cases it is better to compensate the feeder reactance separately before transforming the resistance component, using a coil or capacitor connected in parallel with the feeder (see Fig. 13.57 (d).) If more capacitance than is available appears to be required for resonance, then capacitance loading is needed; if the converse, inductive loading is the solution. A loading component of the same value as the coil or capacitor of the tuning circuit may be tried first.

Adjusting Matching Units and Couplers

The most accurate way of adjusting units such as have just been described is with the use of an s.w.r. bridge or reflectometer. The layout of a typical set is up shown in Fig. 13.34, and a suitable reflectometer is described later in this Chapter.

The procedure to be followed is quite simple and should present no difficulties if followed carefully and in the correct sequence. The reflectometer should be capable of handling the power output of the transmitter and should be of the type which reads both forward and reflected power.

Connect up as shown in Fig. 13.34 and adjust the transmitter power or the reflectometer sensitivity so that in the forward reading position the meter indicates well up the scale with the taps set at a trial position equi-distant from the centre of the coil.

Now switch to read the reflected power. Vary C1 (Fig. 13.41) for the minimum possible reading on the meter.

If a low reading is not possible it will be necessary to adjust the tap positions always making certain that they are symmetrical about the centre of L2.

If the coupling between L1 and L2 is variable then after adjusting C1 for minimum current the coupling should be varied to obtain an even lower reading touching up C1 again. Once again if a low reading is not possible the position of the taps should be changed and the whole procedure repeated.

Where C2 is used and the coupling between L1 and L2 is fixed the procedure is the same but the varying of C2 should replace the varying of the coupling between the two coils.

Multiband Couplers

There are many variations of the aerial couplers described. For example, a pi-network similar to those described in Chapter 6 (*H.F. Transmitters*) for transmitter output circuits is often attractive because it can be arranged to cover as many as four amateur bands with the same set of components. The couplers of Figs. 13.57 and 13.59 can usually be arranged to cover two adjacent amateur bands but where complete coverage is needed it may be necessary to make up a coil together with its link and taps as a plug-in unit for each band, though there may still be some difficulty if the capacitor has too large a minimum capacitance for 28 Mc/s or insufficient maximum capacitance for 3·5 Mc/s.

A popular multiband aerial tuning unit for 3·5–28 Mc/s is shown in **Fig. 13.60** and is known as a *Z-match coupler*. It is a compound network using two pairs of windings and is

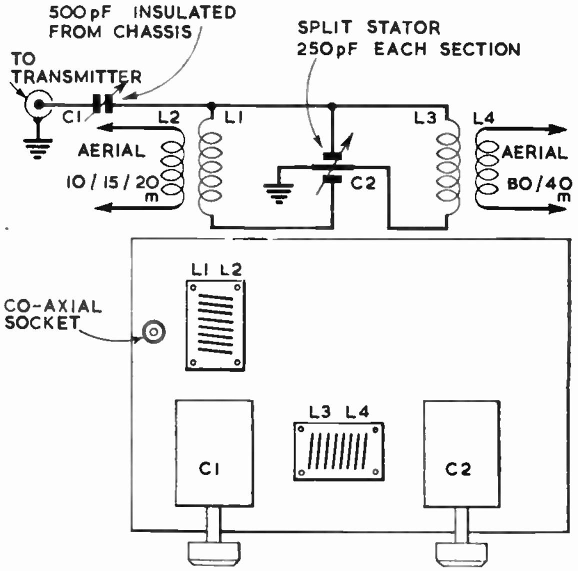


Fig. 13.60(Above). Circuit diagram of the Z-match coupler. (Below) Suggested layout. C1 is the series capacitor and C2 the split stator capacitor in the multiband tuning circuit.

capable of matching the wide range of impedances which may be presented by a tuned aerial line. Coils L_1 and L_2 may each be 5 turns tightly coupled. L_3 and L_4 are 8 and 6 turns respectively. L_1 and L_3 may be about 2½ in. diameter and L_1 and L_4 about 3 in. diameter (Fig. 13.58).

The series capacitor C_1 is 500 pF maximum and it should be noted that it is " live " on both sides: the frame should be connected to the transmitter link cable and an insulated extension shaft provided. The other capacitor, C_2, is the split stator type, 250 pF per section. C_2 tunes the coupler and C_1 adjusts the load to the transmitter. A standing wave indicator (reflectometer) is again an aid to tuning. C_1 and C_2 are adjusted for minimum reflection and the transmitter can then, if necessary, be trimmed for maximum output.

Power Indicators

It is necessary to be cautious in interpreting the power radiated in terms of meter readings, because the current and voltage vary from point to point in an aerial system. Some suggestions as to what to expect may therefore be helpful. A 100 watt d.c. input transmitter will deliver up to 1 amp. into an 80 ohm matched line, but only about 0·35 amp, into a matched 600 ohm feeder. In both cases this can be taken to represent true power. In a tuned line, however, the current may vary over a wide range and does not reveal the real radiated power, though it still serves for indication of tuning adjustments. R.f. thermoammeters are available with maximum readings of 0·5 to 5 amp., but usually have a rather cramped and limited scale at low readings so that they cannot be used for widely differing currents. Care is necessary in cases where the current is greatly different in the same aerial for different bands, since these meters are easily damaged by overload.

When the transmitter power is very low, the current may be below the range of robust meters and a low wattage lamp is the only practicable tuning indicator. On the other hand the corresponding voltage even from a low-power transmitter is easy to indicate and **Fig. 13.61** shows the circuit of a simple voltage indicator, employing a germanium diode.

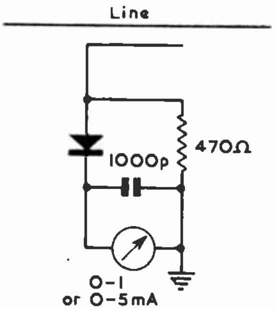


Fig. 13.61. Sensitive germanium diode voltmeter, for indicating aerial or line voltage, as an alternative to a current meter.

There is no need to make direct connection to the aerial line or feeder: sufficient energy can be picked up by a very small capacitance coupling, using as the capacitor an inch or so of wire fixed close to one wire of an open balanced line or inserted into a co-axial line through a slit cut in the outer sheath. The coupling wire should be held in position by adhesive tape. Sufficient coupling can usually be obtained to give several milliamperes of d.c. output: the germanium diode will pass 10 mA without damage. The meter connections can be led back to the operating position but the earth on the voltmeter should be to the aerial circuit earth or the cable sheath.

DIRECTIVE PROPERTIES OF AERIALS

Aerials do not radiate uniformly in all directions, first because there is no radiation from the ends of a straight wire, and second because the wave reflection from the ground produces an interfering effect which may augment or reduce the direct radiation in any given direction. The effect of the ground is mostly disadvantageous, as it generally reduces low angle radiation above the horizontal, though there are cases where it can be used effectively by placing the aerial at the correct height. It also helps to produce a little useful end-fire radiation from a horizontal wire which, in free space, has none.

If there is more than one half wave in an aerial, it is found that the total radiation can be regarded as the resultant of a number of components, one from each standing wave current loop. In any given direction these components may have to travel different distances, so that they do not arrive at a distant point in the same relative phases as they had in the wire. They can, therefore, augment or oppose each other. It is, however, possible to combine a number of elementary aerials so that their radiation accumulates in some favoured direction, at the expense of radiation in other directions. This gives a much stronger signal than a single aerial would give in the required direction for the same power input.

For these reasons aerials are said to be *directive* and to have *gain*.

The concept of gain was discussed on page 13.3 and the method of calculating the gain of simple multiple element aerials on page 13.6. The calculation of the gain of aerials of a more complex form, such as long wires with extended current distribution or Yagi aerials with parasitic elements, is very complicated; often it can only be determined by integration of the power flow in the solid polar diagram of the aerial.

From the point of view of the amateur, the gain of the more conventional types of beam aerial is already known, and it is practicable to construct beam aerials with moderate gain.

Before studying directive arrays, consideration must be given to the form of radiation pattern required by an amateur station. It is frequently desired to be able to work in any direction of the compass and this can be accomplished best by the use of a number of separate aerials. These require considerable space and more often it is necessary to use a compact rotating beam aerial. The directive pattern of such aerials not only gives gain and stronger signals, but in reception also helps to reduce interference from unwanted directions.

Beside directivity in compass direction (azimuth) it is also necessary to have the correct type of vertical radiation pattern. It has been shown in Chapter 12 (*Propagation*) how short wave signals travel round the world in a number of " hops." Since each hop introduces attenuation, a minimum number is desirable. This means that for long distance work low-angle radiation is most effective and experience indicates that the maximum radiated power should leave the aerial at an angle between 5° and 15° to the ground. For shorter ranges, say from the skip distance up to 1000 miles or so, a higher angle is needed. The most useful aerial for general work is therefore one which sends most of its energy at low elevation, but leaves a little at higher angles for short range work. The vertical aerial does just this, but has no directivity in azimuth and is said to be omnidirectional. With horizontal aerials it is not usually possible (except at 21 and 28 Mc/s) to place the aerials sufficiently high to achieve very low angle radiation, and directive aerials are therefore used to force the radiation down towards the horizon.

Long wire aerials have very definite radiation patterns, as will be seen later, but these patterns have several main beams or lobes and hence the gain is not so great as when most of the transmitter power is concentrated into one main beam. In order to achieve the latter condition, two or more elements, such as half wave dipoles, are combined in special ways and the combination is called an *array*. There are two general classes of array—the *broadside array* in which the main beam is at right angles to a plane containing the elements and the *end-fire array* in which the main beam is in the same direction as the row of elements; the two classes are illustrated in **Fig. 13.62.** It should be noted that the term end-fire refers to the layout of the array and not to the ends of the wires, although the term is sometimes applied to a long wire, because its direction of maximum radiation tends towards the direction of the wire.

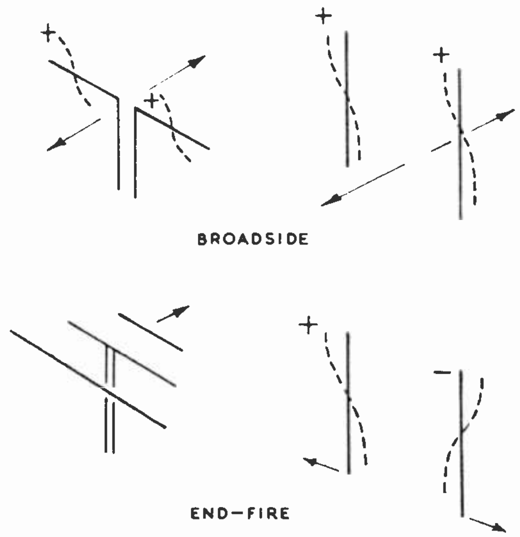


Fig. 13.62. Horizontal (left) and vertical (right) broadside and end-fire arrays, illustrating the terms. The arrows indicate directions of maximum radiation and the voltage standing waves on three of the aerials show their relative polarity or phase. The remaining aerial is a three element parasitic array and is unidirectional.

Radiation Patterns and Polar Diagrams

In order to send the maximum signal in a given direction horizontal plane directivity is required, but to ensure that the signal " reaches out " vertical plane directivity is necessary. Each of these contributes to the total gain of the aerial. To illustrate the radiation pattern of an aerial, *polar diagrams* are used in the form of curves, the radius of which in any direction represents the relative strength of signals in that direction.

The radiation from an aerial occurs in three dimensions and therefore the radiation pattern is really a solid surface; for example, the solid pattern of a half-wave dipole on page 13.4. A polar diagram is any section of the solid shape and a large number of sections may be necessary to detail completely an aerial radiation pattern. In practice it is necessary to be content with two polar diagrams taken in the principal planes, usually the horizontal and vertical, and giving the two cross-sections of the main beam.

Where the polar diagram has a definite directional form, the angle between the directions where the power radiated is half the value at the point of maximum gain (– 3db) is called the *beam width*. These points are marked on **Fig. 13.96** on page 13.56

To avoid confusion when discussing radiation, directions in the horizontal plane are referred to as *azimuth*; angles above the horizontal, in the vertical plane, are called *wave angles* or directions in *elevation*. Confusion often arises additionally when the expression horizontal (or vertical) polar diagram is used, unless it is made clear by a statement of the polarization of the aerial with respect to the Earth's surface. When reference is made to the polar diagram of an aerial in free space, the terms horizontal and vertical have no meaning, and the more precise descriptions of *E-plane* and *H-plane* polar diagrams are to be preferred. These are unambiguous, since the direction of the electric and magnetic fields around the aerial is a function only of the direction of current flow. The electric field (or *E*-plane) is parallel to the direction of the current and usually therefore parallel to the radiating wire. The magnetic field (or *H*-plane) is at right angles to the current and therefore normal to the radiating wire. The polar diagram of the half-wave dipole illustrated in Fig. 13.5 is then an *E*-plane diagram: the *H*-plane diagram of the dipole will be a circle. Such an aerial is then said to possess *E*-plane directivity, and is omni-directional in the *H*-plane.

No matter what name is used to describe the polar diagram, it should be remembered that these radiation patterns are for long distances and cannot be measured locally within a distance of many wavelengths from the aerial.

Construction of Polar Diagrams

Fig. 13.63 is a plan of two vertical aerials A and B, spaced by one half wavelength, and carrying equal in-phase currents. A receiver at a large distance in the direction a at right-angles to the line of the aerials (i.e. *broadside*) receives equally from both A and B. Since the two paths are equal the received components are in phase and directly additive, giving a relative signal strength of two units. In the direction b the signal from A travels one half-wavelength further than that from B. The two received components are therefore one half cycle different, or in antiphase, and so cancel to give zero signal. In an intermediate direction c the two paths are effectively parallel for the very distant receiver, but one component travels further by the distance x. In this case the components have a relative phase $\phi = 360\ x/\lambda$ degrees. The addition for this direction must be made *vectorially* using the " triangle of forces " also illustrated in Fig. 13.63 in which the two equal lines representing signal strength from A and B are set at the phase angle ϕ which corresponds to the path-difference x; the total or resultant signal is equal to the

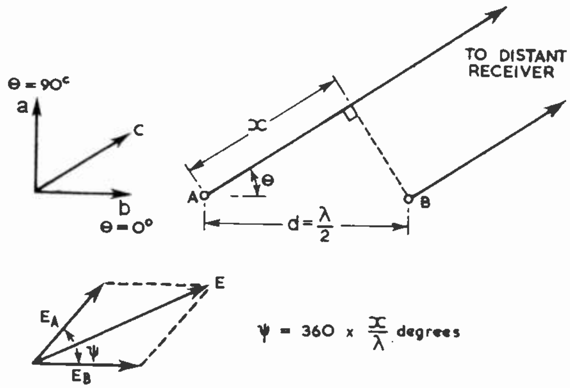


Fig. 13.63. The radiation pattern of two sources can be calculated by vector addition of the respective field contributions at various angles of azimuth. For simple arrays this can be converted into a trigometrical formula.

relative length of the line E_R which completes the triangle. This process could be carried out for all directions, but is tedious and it is usual therefore to convert the operations to a trigonometrical formula from which the result can be calculated more quickly.

In the general case of two aerials carrying equal but dephased currents and spaced a distance d apart, the appropriate formula for the H-plane polar diagram is:

$$E = 2E_0 \cos\left(\frac{\pi d}{\lambda}\cos\theta + \frac{\phi}{2}\right) \qquad \ldots\text{(i)}$$

where θ is the angle measured as shown in Fig. 13.63, ϕ is the electrical phase difference between the currents in the two aerials, and E_0 is the value of the field from one aerial alone in the direction θ. In the particular case illustrated in Fig. 13.63, $\phi = 0$, E_0 is constant for all directions of θ, and $d = \lambda/2$, so that the shape of the polar diagram can be plotted from the expression:

$$E = \cos\left(\frac{\pi}{2}\cos\theta\right)$$

The absolute magnitude of the field E will depend upon the actual value of the current flowing in each aerial and is of no direct importance when calculating the shape of the pattern. It comes into account only when the relative patterns of two aerial systems are being considered and the question of gain arises.

The polar diagram found in this way is a figure-of-eight, the curve marked $\lambda/2$ in **Fig. 13.64**(a). If, on the other hand the two aerials had been in antiphase, the resultant signal would have been zero in the direction a and maximum in direction b and the polar diagram as in Fig. 13.64(b).

In that case, $\phi = 180°$ (or π) in expression (i) above, and the formula for the polar diagram becomes:

$$E = \cos\left(\frac{\pi}{2}\cos\theta + \frac{\pi}{2}\right)$$
$$= \sin\left(\frac{\pi}{2}\cos\theta\right)$$

This expression for the anti-phase aerials is very similar in form to that for the in-phase aerials, and the change from cosine to sine reveals that the general pattern shape is the same, but turned through 90° of azimuth This is confirmed by the curves of Fig. 13.64.

A similar procedure is used if the aerials do not carry equal currents, or if they have an arbitrary phase relation. In this case the sides of the triangle would be drawn in lengths proportional to the two aerial currents and the relative phase of these currents would be added to the angle ϕ due to path differences.

The path difference x and phase ϕ of course are proportional to the spacing between the aerials, which in the example above, was one half wavelength. If the spacing is one wavelength then the phase difference changes twice as rapidly with change of direction, and so there are twice as many maxima and minima to the pattern, giving the curves marked λ in Fig. 13.64. These patterns are said to have four *lobes*.

With increased spacing between the two aerials, more lobes appear, two for each half-wave of spacing, and the pattern becomes like a flower with many petals. This type of pattern is not very useful, but if the intervening space is filled with aerials spaced $\lambda/2$, one pair of lobes grows at the expense of all the others, giving a sharp main beam with a number of relatively small *minor lobes*. This is the basis of stacked beam arrays. It should be noted, however, that the beam is only developed in the plane in which the array is extended; a broadside array of aerials all in phase produces a beam which

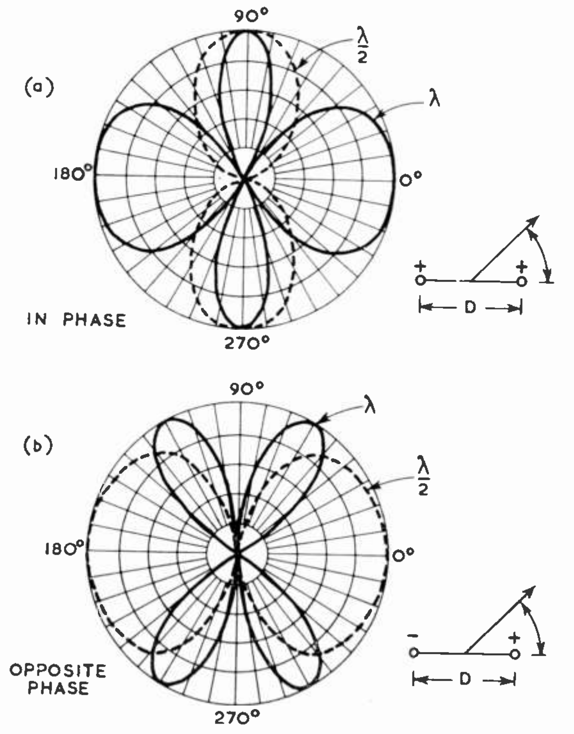


Fig. 13.64. Horizontal polar diagrams of two vertical aerials spaced λ/2 and λ. The upper diagram is for aerials in phase, the lower diagram for antiphase connection. If separate feeders are used the pattern can be changed by reversing polarity of one feeder. Directions 90°, 270° are broadside: 0°, 180° end-fire. Note that the end-fire diagrams are broader than the broadside ones.

is narrow in the horizontal plane, but its vertical pattern is the same as that of a single aerial. It is therefore necessary to extend a stacked array in the vertical direction to obtain a sharp vertical pattern. In an end-fire array where the aerials are spaced and phased to give the main lobe along the line of aerials, the vertical and horizontal patterns are developed simultaneously by the single row of radiators because the array is extended simultaneously in both of these planes. The azimuth pattern of an end-fire array is always broader than that of a broadside array of the same length; this effect can be noted in the patterns of Fig. 13.64.

Although the trigonometrical formula for the polar diagram of a large array may be very complex, and therefore laborious to plot in full, it is always readily possible to find the directions of minimum radiation or *nulls* by solving the equation for θ when E is made equal to zero.

Unidirectional Patterns

The patterns so far considered are symmetrical: a somewhat different combination of aerials gives a unidirectional pattern. The two dipoles in **Fig. 13.65** are connected a quarter-wave apart along a common feeder, with a wave entering from the left. In the direction away to the right it does not matter whether radiation leaves via the first aerial or the second; it takes the same time to reach its destination. Thus to the right, the components of radiation are additive because they are in phase. This is shown by the upper set of broken lines. To the left, however, the wave from the end dipole has to travel further than that from the first dipole by a distance equal to twice the spacing between them. This extra journey is one half wavelength and hence the two components differ in phase by 180° and cancel. Because of the $\lambda/4$ spacing the currents in the two aerials have a phase difference of 90° and are said to be in *quadrature*.

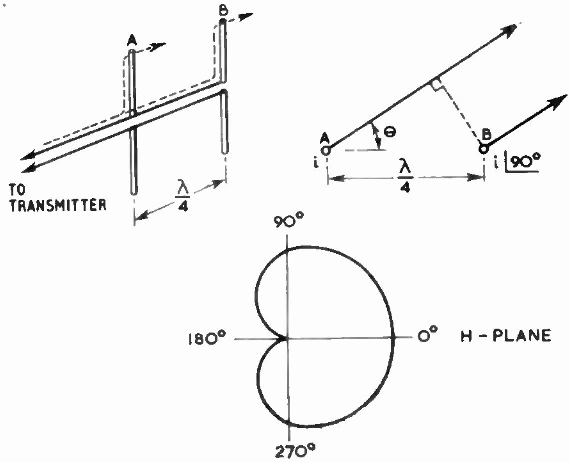


Fig. 13.65. Illustrating the fundamental principle of the driven reflector. The two wave components have equal path lengths in one direction, and therefore add, but in the other direction the paths differ by $\lambda/2$ and the components cancel. The polar diagram is the geometrical figure called a cardioid (heart shape).

The actual shape of the pattern can again be determined by substituting the appropriate values for d and ϕ in expression (i) given earlier, to give the formula:

$$E = \cos\left(\frac{\pi}{4}\cos\theta - \frac{\pi}{4}\right)$$

This fundamental principle is often used in broadside arrays in order to make the sharp radiated beam unidirectional by placing a second set of aerials a quarter-wave behind the main set. For the same reason, long end-fire arrays are built with quarter-wave spacing and progressive 90° phasing. In practice, however, it is extremely difficult to arrange the feeder so that the dipoles have the required currents and phase, and parasitic reflectors are more often used instead, as will be described later.

Array Factor

The radiation pattern of a large aerial array can often be calculated by breaking down the array into units of which the individual pattern is known, and then combining the patterns of the units together. For the purposes of combining the unit patterns, each is assumed to be a " point source " of radiation, located at its physical centre, such as the aerials A and B in Fig. 13.63, and an expression F_1 derived for the pattern of these sources. The final pattern F of the whole

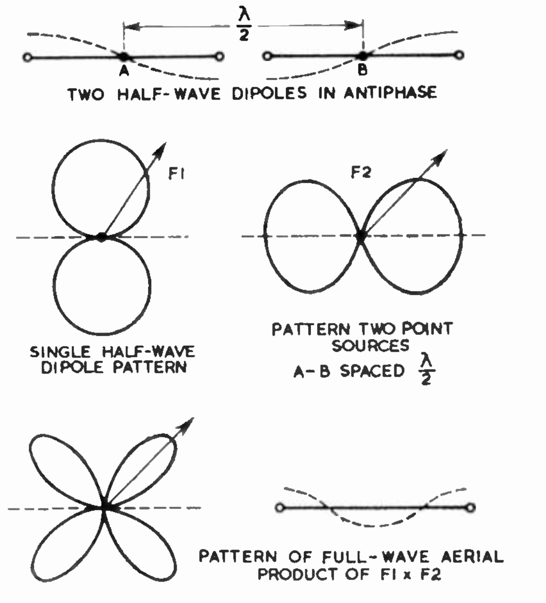


Fig. 13.66. The array factor. Each dipole has a pattern like F1. Two sources in opposite phase at A and B would have a pattern like F2. For any direction (e.g. the arrows) the chord F1 is multiplied by that of F2 for the same angle. When this is done for all angles the pattern shown at the bottom is obtained for the complete aerial.

aerial is then obtained by multiplying F_1 by the pattern of the individual units F_2 so that:

$$F = F_1 \times F_2$$

The expression F_2 is known as the *unit* pattern and F_1 as the *array factor*.

For very large arrays of dipoles this technique can be repeated by successive breaking down of the array into smaller and smaller units. The application of the array factor to the determination of the *E*-plane polar diagram for a pair of collinear antiphased half-wave dipoles is shown in **Fig. 13.66**, where *F* is the final pattern, F_1 that of the dipole and F_2 the array factor or pattern of the set of point sources. F_1 is unity for the azimuth pattern of vertical aerials. If each aerial were fitted with a reflector, then another multiplying factor would be included to represent the cardioid (heart shape) of Fig. 13.65.

A particular application of array factor is the conversion of *H*-plane to *E*-plane polar diagrams for arrays made up of parallel half-wave dipoles. Once the *H*-plane pattern has been established by measurement or calculation, the *E*-plane pattern is obtained merely by multiplying the *H*-plane pattern by the dipole factor F_d (the *E*-plane pattern of a single half-wave dipole) so that:

$$E\text{-plane} = H\text{-plane} \times F_d$$

Since the half-wave dipole possesses *E*-plane directivity normal to its axis the effect of converting the *H*-plane diagram is to sharpen it in the directions at right angles to the axis of the dipoles, at the expense of radiation in the directions along the axis of the dipoles. This is illustrated in **Fig. 13.67** which shows the *E* and *H*-plane diagrams for an array of two parallel dipoles with reflectors.

Vertical plane diagrams are found in the same way, by treating the aerial and its image in the ground as an array of two sources spaced by twice the height of the real aerial.

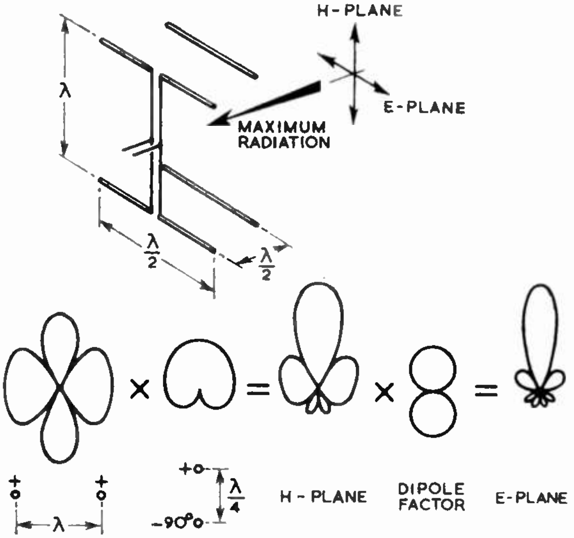


Fig. 13.67. Conversion of *H*-plane to *E*-plane polar diagrams using the principle of array factor.

Vertical Radiation Patterns over Earth

The relationship between the sign of the aerial and its image for the separate cases of horizontal and vertical polarization is shown in **Fig. 13.68.** In the case of the horizontally polarized aerial, the image is of opposite sign or phase, and the resultant field along the surface of the reflecting plane will always be zero. In the case of the vertically polarized aerial, the image is effectively of the same sign as the aerial, and the radiation is reinforced along the surface of the reflecting plane. For this reason only a vertically polarized aerial can give rise to a ground wave, there being little or no radiation at low angles from a horizontally polarized aerial.

A number of useful vertical patterns are given in **Fig. 13.70** from which it will be seen that choice of correct height is much more important for a single aerial than for arrays which themselves tend to radiate at low wave angles.

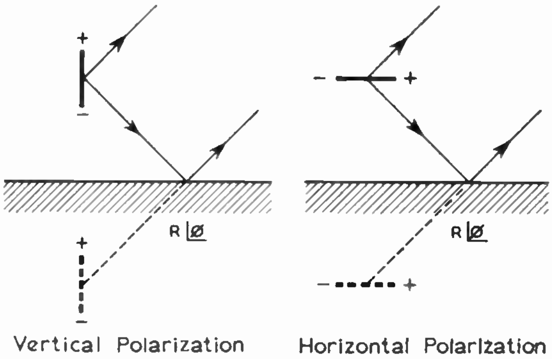


Fig. 13.68. Images of aerials above a reflecting ground plane. The vertically polarized aerial produces an image which is in phase and supports radiation along the earth's surface. The horizontally polarized image (reverse the image polarity shown) is in anti-phase and cancels the radiation along the earth's surface.

Effect of Earth Conductivity

When considering the effect of the ground on the radiation from an aerial, the ground was assumed to be a perfect conductor and, therefore, a perfect reflector. The vertical plane diagrams of Fig. 13.70 are based on this assumption, but in practice the soil is often far from perfect, behaving rather like a clouded mirror and giving a weak image.

This has the effect that the vertical plane polar diagrams differ from the ideal diagrams for perfect reflectors.

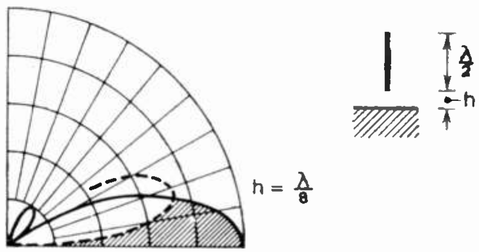


Fig. 13.69. Effect of ground conductivity and Brewster reflection on vertical radiation pattern of a vertically polarized aerial over a practical earth.

In Chapter 12, *Propagation*, reference was made to the Brewster effect and the resulting differences in the reflecting properties of the ground for horizontal and vertical polarization. In the former case the effect is small, and there is little difference between the ideal and actual vertical polar diagrams of horizontally polarized aerials over any but earth of the poorest conductivity. However, for vertically polarized aerials, the radiation at low angles in the vicinity of the Brewster region (10-15°) is greatly reduced over poor ground, and degraded even over ground of relatively high conductivity. Typical of the manner in which the actual vertical pattern departs from the idealised one is the diagram of **Fig. 13.69.** A more searching treatment of this aspect is to be found in Reference [5].

Total Radiation Patterns over Earth

All the preceding references to polar diagrams in the *E* or *H*-planes of an aerial, are derived for radiation of the appropriate polarization. For aerials in free space, the diagrams are entirely valid for there is no external influence to produce any effective change of polarization. In free space there is literally no radiation off the ends of a dipole, and the *E*-plane diagram correctly falls to zero in those directions.

If a dipole is disposed horizontally above an earth plane acting as a reflector, there will then be some radiation off the end at angles of elevation other than zero, and this radiation will be *vertically* polarized. This effect can be appreciated by holding a pencil parallel to the surface of a mirror and observing the image from the end-on direction at various heights above the mirror.

In Chapter 12, *Propagation*, it was explained that during h.f. propagation by means of ionospheric reflection, the waves undergo random changes of polarization in the ionosphere, and the resultant signal at the receiving end is the sum of the various components of differing polarization. Consequently, energy radiated at the sending end at the desired angles of elevation but with different polarization can also be considered useful since it will contribute to the total received signal at the remote end, and it is therefore necessary in practice to examine the effective radiation pattern in azimuth of the horizontally polarized aerial over the reflecting earth, taking into account the total energy radiated in each direction irrespective of polarization.

The free-space *E*-plane polar diagram of a horizontal half-wave dipole is shown in **Fig. 13.71(a).** This also represents the total horizontally polarized radiation in azimuth, for all angles of elevation above the ground. To this must now be added the contribution of vertically polarized radiation arising as a result of the presence of the reflecting ground. This is a function of the particular angle of elevation, and

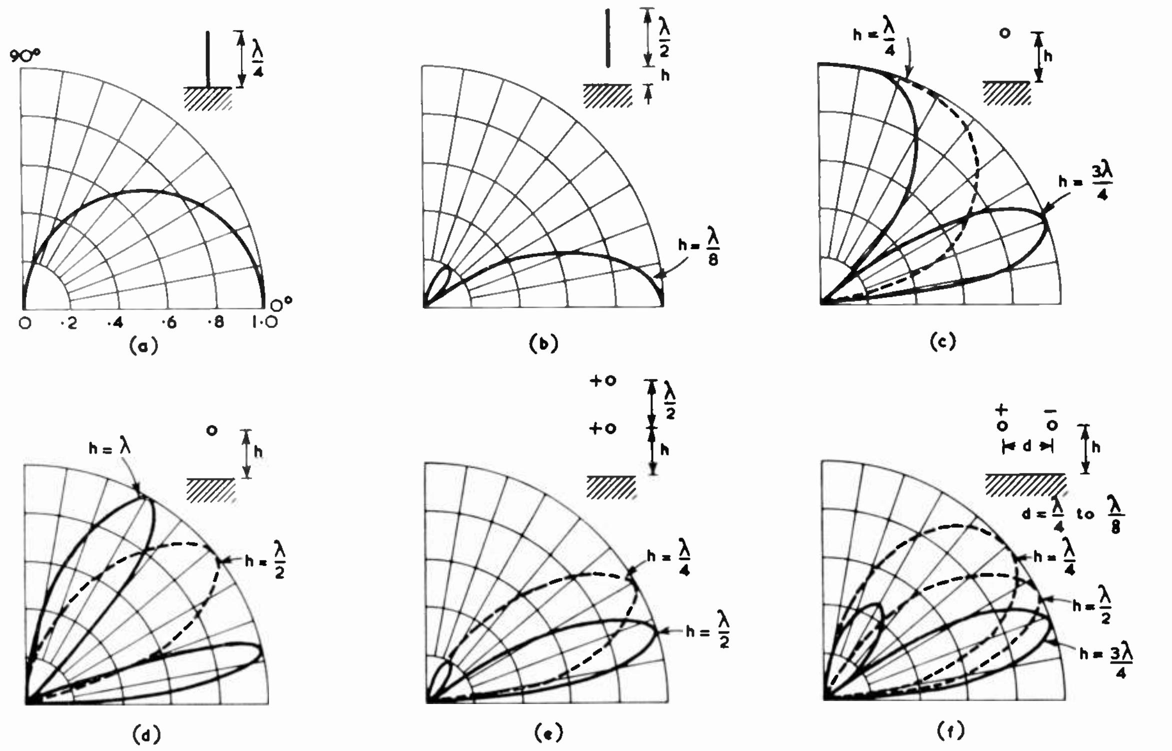


Fig. 13.70. Vertical plane radiation patterns. (a, b) Vertical aerials. (c, d) Horizontal dipoles and collinear arrays. (e) Broadside horizontal arrays. (f) End-fire horizontal systems—e.g. W8JK. Only half of each pattern is shown: they are symmetrical about the vertical axis, unless reflectors are also used. Diagrams (a) and (b) hold for all azimuth directions: the remainder are for broadside direction only. The beamwidth in azimuth depends on the length of the arrays, as shown in Fig. 13.96.

results in a variable degree of "filling in" of the nulls on the ideal *E*-plane diagram indicated by the broken line on Fig. 13.71(a). For any given aerial height above ground, and for any particular angle of elevation, it is possible to obtain the basic horizontal pattern from Fig. 13.71(a), apply a scale factor derived from the curves of **Fig. 13.70(c)(d)** for the appropriate wave angle, and add in the vertically polarized radiation off the ends to arrive at a total azimuthal radiation pattern *appropriate to that particular height and elevation.* Examples of such patterns are given in **Fig. 13.71(b)** for an angle of elevation of 30° above the horizon, and **Fig. 13.71(c)** for an angle of 15°. It is important to note that the dipole tends only to exhibit reasonable horizontal directivity at the lower wave angles, and is tending to become omni-directional for the higher angles associated with short skip propagation.

DESIGNS FOR PRACTICAL AERIALS

The first sections of this chapter have been devoted to the general aspects of aerials and transmission lines without any specific attempt to relate these to the particular requirements of any one amateur frequency band. The rest of the chapter deals with practical aerial designs and has been divided into those suitable for the low frequency bands (1·8 Mc/s and 3·5 Mc/s), and those which the average amateur can expect to realize for the higher frequency bands from 7 Mc/s to 28 Mc/s. A separate section is devoted to the use of vertical aerials, which behave rather as a class on their own, relative to the wide range of horizontally polarized aerials, and are a type of aerial which can be made to work fairly well on all bands from 3·5 Mc/s upwards, and can be contained within a reasonable physical size.

LOW FREQUENCY AERIALS

Marconi Type

The simple types of aerial described so far have been illustrated as horizontal aerials. On the lower frequencies, the 1·8–2·0 Mc/s and 3·5–3·8 Mc/s bands, there is often insufficient space to erect such an aerial and even if there were, it is unlikely that it would be possible to erect it sufficiently high to be really effective. In such cases some form of Marconi aerial is used. This is essentially a vertical aerial, working against an earth connection; the inverted *L* type is also usually considered to be a Marconi type aerial. In these aerials the image of the aerial in the ground is used to help make up the effective length to half-wave resonance.

The choice of a vertically polarized aerial for the lower frequency bands follows from the propagation characteristics explained in Chapter 12. On these bands there is little or no ionospheric propagation during the daylight hours, and a restricted amount during the period of darkness. The predominant mode for most of the time is ground-wave, and this can only be radiated by a vertically polarized aerial, which then represents a reasonable compromise between the conflicting day and night requirements.

The ideal length for a vertical aerial working against earth is one quarter-wave and since it is in effect a bisected half-wave aerial, its radiation resistance is about 35 ohms, i.e. half that of a half-wave dipole. Such a height is not usually possible, even for 3·5 Mc/s where it would be 66 ft., and hence it is necessary to use an electrically short aerial and to load it in such a way as to bring it to resonance, and if possible to increase the radiation efficiency. An aerial one tenth of a wavelength high (50 ft. for 1·8 Mc/s) has a radiation resistance of 4–5 ohms. This would not matter if the earth were a perfect conductor, but it is far from being so, and the resistance due to a most elaborate earth system, together with

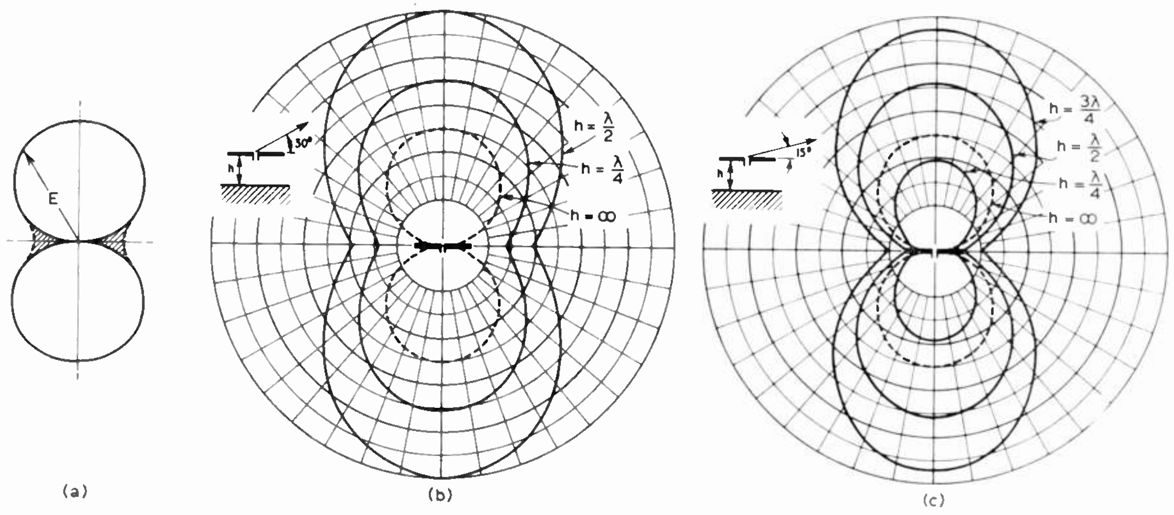


Fig. 13.71. Total azimuthal radiation pattern of a horizontally polarized half-wave dipole at varying heights above a perfectly reflecting earth. (b) At a wave angle of 30 . (c) At a wave angle of 15 . The curves are of relative field and are to the same scale. Poorer earth reflections will tend to emphasise the minima off the ends of the dipole.

the loss due to the current returning through the surface, may contribute 20 ohms or more. The efficiency is thus low. At higher frequencies where vertical aerials are used for their desirable low angle radiation properties, ground plane systems can be used to reduce earth loss, but at the lower frequencies they would be too large and other methods of improving the efficiency must be sought.

The relative efficiency of various combinations of earth resistance and aerial height are considered in more detail in the following section on Earth Systems.

It is the centre part of a tuned aerial, where the current is highest, that is responsible for most of its radiation. The ends (where voltage is high but current low) do not contribute much to the radiation, but are necessary in order to bring the aerial to resonance and build up the current at the centre. When the full resonant length cannot be erected it is necessary, in order to increase the efficiency, to load the remote end of the aerial in some way so that its electrical length is increased and so move the current maximum into the vertical part of the aerial. Another procedure is to increase the total effective length of the aerial to more than a quarter wave. This raises the impedance at the feed point, so that the power loss due to the resistance of the earth connection is relatively less.

The simplest form of loading is to add a horizontal top to make a *T* or inverted *L* aerial (**Fig. 13.72 (a)**). This loading should be arranged to create a current maximum in the centre of the vertical part. The distance from this point to the end of an *L* aerial is about 130 ft. for 1·8 Mc/s but somewhat less for a *T* aerial. If this cannot be done, the next alternative is to increase the loading effect of the top by making it into a cage or " flat top " of two or three wires joined in parallel; this increases the capacitance and effectively lengthens the aerial. If this procedure is not practicable, the aerial may be loaded by including a length of folded or coiled wire (see **Fig. 13.72(b)**) near the free end of the aerial. The coil must be arranged with care, using the same size wire and avoiding sudden bends so that the standing wave starts from the far end of the aerial and not from a reflection at the coil. It may be inconvenient to support the coil in the middle of the top, but if it is to be fitted to the mast at the far end of the aerial then it must be given a capacitance " platform " to work against. This can take the form of a large can or closed screen of wire enclosing the loading coil, the far end of which is connected to it. This loading is a very noticeable help where the total space only allows the erection of a wire one-eighth of a wavelength long and makes it possible to obtain useful results on the l.f. bands when the loft of a house is the only place for the aerial.

Other methods of end loading may be devised; for example, folding the aerial back on itself with spacing of, say 6 in., the aim being to bring the desired current maximum into the main part of the aerial.

Aerials for 1·8 Mc/s with relatively high input currents are preferably series-tuned, using about 250 pF capacitance with, say 30μH inductance. A suitable coil for 1·8 Mc/s could be made with 20 turns of 16 s.w.g. wire on a 3 in. diameter former, and spaced to occupy a length of 3 to 4 in. The transmitter link could be tapped across a few turns at the earthed end of the coil or the wire itself coupled to the transmitter output tank coil. About half the values given should be used for 3·5 Mc/s.

To reduce the deleterious effect of an imperfect earth connection the aerial may have a total length of about $\frac{3}{8}$ or $\frac{5}{8}$ of a wavelength while still maintaining the maximum current in the vertical part. A very effective aerial is shown in **Fig. 13.72(c)** where the main half-wave is folded into a *U* with one leg near the ground and adjusted in length until maximum current occurs at a point half way up the right-hand vertical section. The down-lead makes the total length up to about $\frac{5}{8}$ of a wavelength, so that the impedance at the feed point is a few hundred ohms and capacitive, whilst the earth current is relatively low. The coupling circuit for this aerial should be parallel tuned and use the components described for Fig. 13.72(b).

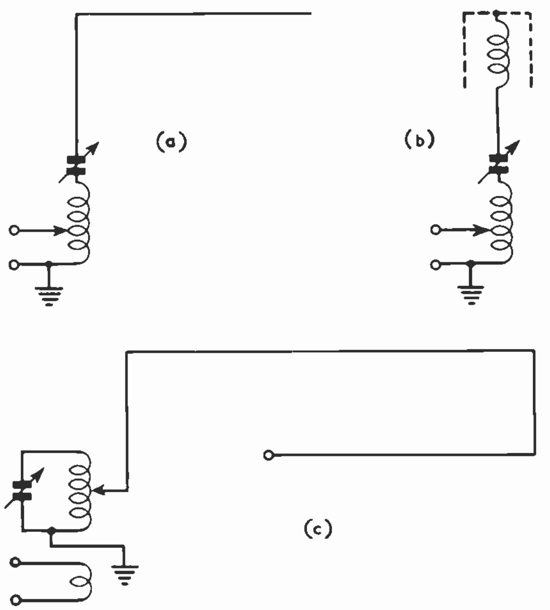


Fig. 13.72. Aerials for low frequency bands. (a) Quarter-wave (inverted L) aerial with series tuning. (b) End loading to raise the efficiency of a short aerial. (c) Special extended aerial with high efficiency, detailed in text. The input current in aerials (a) and (b) may be 0·5 amp. for 10 watts input; that in (c) is low, say 0·1 amp for 10 watts.

The 1·8 Mc/s aerials of Figs. 13.72(a) and (c) can also be used effectively on 3·5 Mc/s and higher frequencies, though it is difficult to predict their precise performance.

Earth Systems

The earth system employed has a considerable effect upon the overall efficiency of a low frequency aerial when the ground acts as the return path for the flow of r.f. currents. This is most commonly met at frequencies up to 2 Mc/s when the aerial system in use comprises an unbalanced radiator, often partly or wholly vertical, with an effective length up to a maximum of a quarter-wavelength or just over. In such cases the aerial is driven by a generator connected between the bottom of the radiator and the ground (**Fig. 13.73**).

The current distribution along the radiator is such as to be a maximum value at or near the ground, decreasing approximately sinusoidally to the end of the radiator at which point it must be zero. Because of the current in the radiator, charges are induced in the earth surrounding it, to give rise to a conduction or circulating current which flows back to modern practice is to use a plastic tubing which will negate the whole exercise.

The earth connection of the electricity supply is dangerous and may introduce noise into the receiver or be responsible for spreading interference to nearby television receivers. The transmitter earth connection should be separate from the safety earth for the equipment in the station, as it is part of the aerial system, carrying r.f. power.

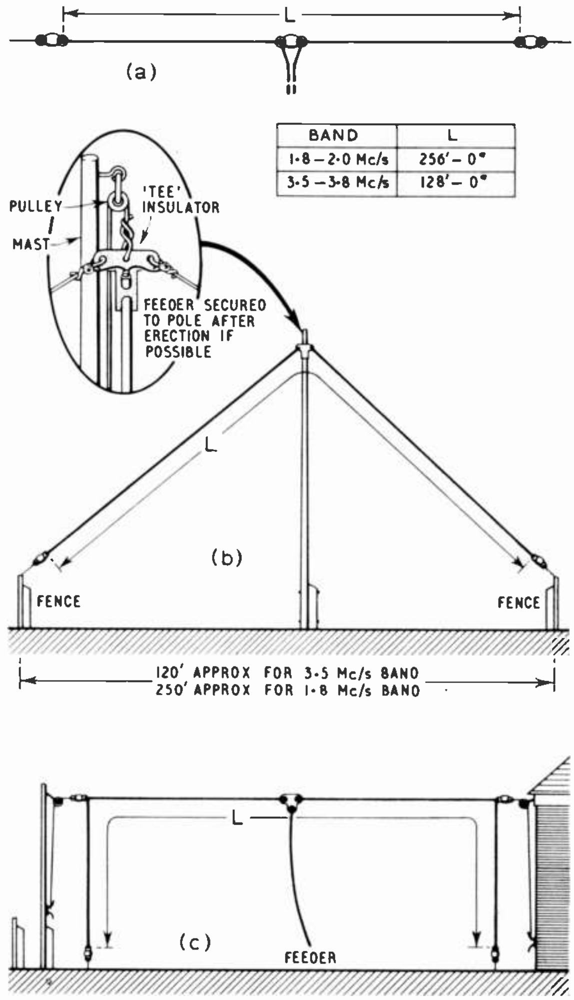
It is a mistake to bring a long " earth " lead into the radio room, which is often some distance from the point of connection to true earth, being in the limit on the upper floors of a building. The result of this is that the long earth lead will necessarily radiate since it carries the aerial feed current, and in consequence the equipment in the shack will be up-in-the-air to r.f. with many consequent problems of filtering and feedback. A far better arrangement is to install the a.t.u. in a box at ground level immediately adjacent to the earth mat connection, and to connect back to the shack with a low impedance co-axial line matched into the a.t.u. This will not only isolate the shack from the radiating part of the aerial system, but will permit the vital vertical section of the aerial to be installed clear of obstructions, which could otherwise well affect its performance.

Horizontal Dipoles

In the section on Marconi aerials it was explained that the predominantly ground wave propagation experienced on the 1·8 Mc/s band favours the use of a vertically polarized aerial. This is also true to a lesser extent for the 3·5 Mc/s band, but there is the possibility on both these bands of achieving ionospheric propagation particularly at night, with a consequent requirement for some radiation at higher angles (see also Chapter 12, *Propagation*). This is more readily achieved by the use of a horizontal half-wave dipole aerial, which is a balanced load and does not require the use of an earth return path for the transmitter connection. Such an aerial also offers a better compromise in cases when the available earth system is an extremely poor one which would result in excessive ground losses when used in conjunction with a Marconi aerial. The appropriate lengths for half-wave dipoles for the 1·8 Mc/s and 3·5 Mc/s bands are given in **Fig. 13.79.** Because of the large overall dimensions and the low frequencies in use, it is often impossible to erect the whole aerial at an ideal height (greater than a quarter wave: 132 ft. at 1·8 Mc/s; 66 ft. at 3·5 Mc/s), and in the clear.

The current distribution along the dipole is roughly sinusoidal and is concentrated in the middle with little or no radiation from the ends of the wire. The aerial should therefore be supported in such a way as to keep the centre region as high as possible, the ends being allowed to droop, or to hang down, depending upon the available space. Two possible arrangements for restricted gardens are shown in Fig. 13.79.

The radiation resistance of a half-wave dipole tends to be low when the aerial is less than a quarterwave above ground, being a function of height as shown in **Fig. 13.80**, and a standing wave ratio of as much as 3 : 1 may be expected when the dipole is fed with 75 ohm coaxial line, and a suitable balun. Such an s.w.r. may be regarded as an acceptable compromise on the low frequency bands involved.



BAND	L
1·8 – 2·0 Mc/s	256′ – 0″
3·5 – 3·8 Mc/s	128′ – 0″

Fig. 13.79. Dipole horizontal aerials for 1·8 and 3·5 Mc/s. The current is low at the ends and contributes little to the radiated signal.

The W3EDP aerial

Another approach to this problem which avoids the necessity for a direct earth connection bears the call-sign of its originator, W3EDP. In this design a short counterpoise wire is used in place of an earth, the aerial itself being suitably shortened to compensate for this addition.

A specific design which has been worked out experimentally for multiband use, employs an aerial cut to a length of approximately 84 ft., providing a simple and effective end-fed installation in cases where this length can be conveniently accommodated as a clear run from the radio room window. It is particularly useful when the radio room is on an upper

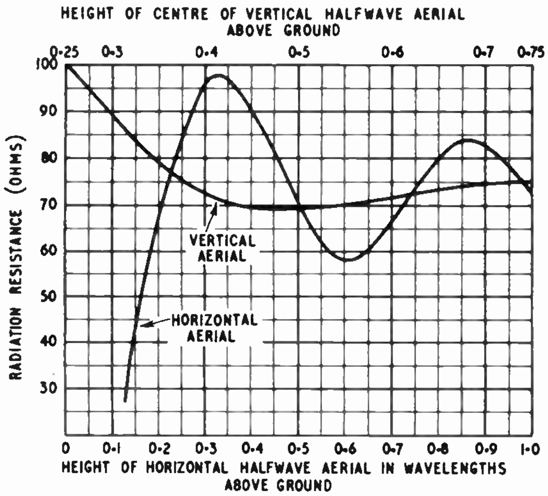


Fig. 13.80. Radiation resistance of half-wave horizontal and vertical dipoles as a function of height.

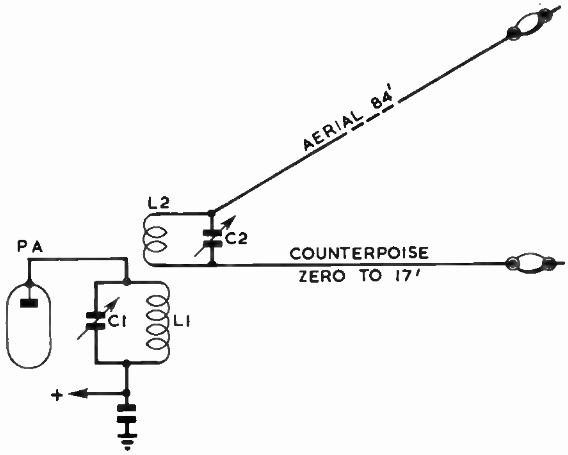


Fig. 13.81. The W3EDP end-fed aerial using a counterpoise wire instead of an earth connection.

floor and would involve a long and awkward earth lead. The counterpoise will vary from zero to 17 ft. in length, according to the frequency in use, as set out in the table of dimensions beneath **Fig. 13.81.** It can be dropped out of the station window, or even accommodated indoors, provided that the far end is well insulated and as far as possible clear of earthed obstructions, since this will be a high voltage point during use. Ideally the counterpoise should run at right angles to the aerial, which may be bent if necessary, or can be sloping if it is borne in mind that maximum radiation will then take place in a direction from the lower extremity.

The original W3EDP aerial was coupled directly to the transmitter tank coil, the counterpoise being connected to the end of the coupling coil nearest to the tank, since a certain amount of capacitive coupling at this point is stated to be desirable for best performance. The coupling coil which should be of plug-in construction must also be proportioned correctly, and capable of variable coupling to the tank so that the p.a. can be correctly loaded. If aerial current falls off before full loading can be attained, the balance between capacitive and inductive coupling may not be correct, or it may be necessary to adjust aerial length slightly to compensate for local circumstances. An aerial of this type may be expected to radiate harmonics if these are present in the transmitter at excessive strength, and there is no basic reason why it could not be link-coupled as shown in Fig. 13,72 for an earthed system. Some experiment would be necessary in this instance to arrive at the optimum aerial and counterpoise dimensions. Suitable harmonic filters can then be used in the coupling line.

HIGH FREQUENCY AERIALS

There is a wide range of choice of aerials for the higher frequency bands, ranging from the simple resonant wire types to the rotary Yagi beam. The selection is limited only by the individual choice of the amateur and the extent to which he wishes to elaborate for any particular band.

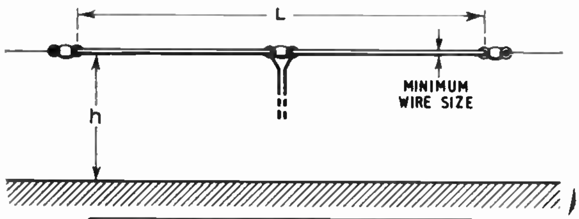
Simple Wire Aerials

These are broadly speaking made up as single lengths of wire cut to resonate at the appropriate band, or by careful adjustment of aerial and feeder, to achieve a compromise performance over several bands. In most cases the gain and directivity are modest, and their advantage lies in their cheap construction and inconspicuous nature when erected.

Half-Wave Dipole

This is the simplest single band aerial. The length appropriate to each band is given in **Fig. 13.82**, together with the optimum height above ground for general DX work. The latter is not critical to ± 10 per cent. The aerial may be constructed of any suitable wire, the gauge being determined by the span involved between the supporting points *including the horizontal portion of the halyard.* Recommended wire sizes for each band are also given in the table. The input impedance to the dipole is approximately 75 ohms balanced, and it may be fed by coaxial line and a suitable balun described on page 13.30. In order to avoid coupling to the outer sheath of the cable it should be allowed to hang vertically from the feed point connection for at least a length equal to a quarterwave in free space.

Simple dipole aerials (and indeed any halfwave single radiator fed by other means, e.g., Zepp) exhibit some degree of directivity at right angles to the wire. As explained on page 13.39 there is a significant amount of radiation off the ends depending upon the wave angle employed.



BAND	L	h 0·6λ	MINIMUM WIRE SIZE
7 Mc/s	66′ – 0″	86′ – 0″	7/0·029″
14 Mc/s	33′ – 0″	43′ – 0″	7/0·029″
21 Mc/s	22′ – 0″	28′ – 0″	16 swg
28 Mc/s	16′ – 0″	21′ – 6″	18 swg

Fig. 13.82 Half-wave dipole data for H.F. bands.

The "Windom" Aerial

When an aerial is resonant the " point impedance " (i.e. ratio of voltage to current) at any point on it is a pure resistance, varying from zero at the centre to several thousand ohms at the free end. There will therefore be an intermediate point that will match a single wire feeder, having for example, a characteristic impedance of about 1000 ohms. If this point can be found, the radiation from the feeder will be minimized. In the original Windom (**Fig. 13.83(a)**) the

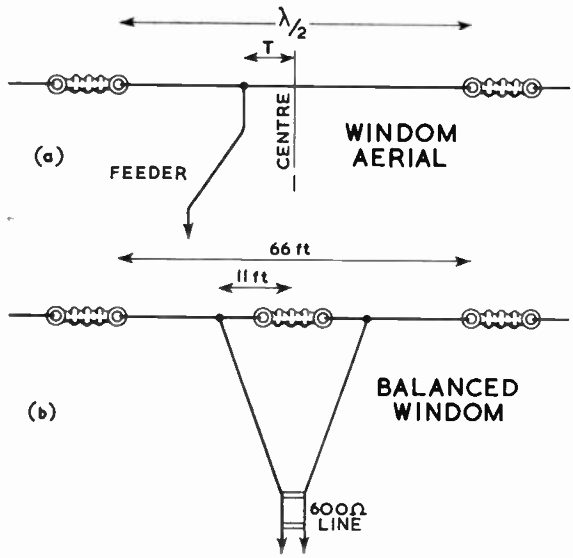


Fig. 13.83. (a) The Windom single wire fed aerial. (b) balanced version which minimizes feeder radiation and also works on multiple frequencies.

radiator is one half wavelength long, but in practice the system will work with any mutiple half-wave radiator. It is a simple and attractive arrangement but requires careful adjustment to achieve best performance: the radiator must be properly resonant before the correct tapping position can be found and the feeder should hang vertically from the radiator for at least a quarter wavelength.

The correct method of adjustment is first to choose an approximate position for the tap and then insert a pair of r.f. ammeters either side of the tap. Assuming the meters are alike (this may be checked by inter-changing them), they will read equal currents when the aerial has been adjusted to resonant length. The meters are then inserted in the feeder, one near the aerial and the other a quarter wavelength away and the tap moved until equal currents are again indicated. The dimension T in Fig. 13.83(a) is approximately one eighth of the half wavelength, but in practice this dimension will vary somewhat because of effects due to the environment of the aerial. **Table 13.4** gives typical adjustment factors. The coupling circuit at the feeder input may be similar to that described for end-fed aerials.

Although there will always be some radiation from the feeder the Windom aerial can give excellent results. Its success depends on having good conductivity soil beneath it and for this reason it is more suitable for damp locations. It is essentially a one-band aerial, but for bands other than one for which it is resonant it can be used as an end-fed aerial (aerial and feeder as one long wire). Radiation will then be partly horizontally polarized and partly vertically polarized.

TABLE 13.4

Single Wire Feeders

Band and Aerial	Radiator Length (feet)	Tap Distance (feet)
3·5–14 Mc/s half-wave	L = 470/f (Mc/s)	T = 66/f (Mc/s)
21–28 Mc/s halfwave	L = 460/f (Mc/s)	T = 66/f (Mc/s)
14–28 Mc/s full-wave	L = 960/f (Mc/s)	T = 170/f (Mc/s)

The values given are recommended for aerials constructed from 14 or 16 s.w.g. wire.

The VSIAA Aerial

It will be seen from the figures of **Table 13.4** and from the formula for half-wave aerials that a given length of wire does not resonate at true harmonic multiple frequencies. A 66 ft. wire may resonate on 7 Mc/s but would have to be nearer 68 ft. long for full-wave resonance on 14 Mc/s. At higher frequencies, as the wire accommodates more and more half-waves, the difference becomes less significant because the resonance of the aerial becomes more heavily damped by radiation. It thus becomes possible to construct a form of multi-band Windom, with the tap in an approximately correct position on all even harmonic bands. If the tap is placed one third of the way along a standing wave current loop on the lowest frequency it will always be in the same position on a current loop on all even harmonic frequencies. This system is most suitable for very long wires, the tap being about 22½ ft. from the end for all frequencies down to 7 Mc/s, and even works quite well with long aerials on 21 Mc/s (where the " one third " rule breaks down), because the long wire places a substantial load on the end of the feeder.

The Double Windom

The VS1AA approximation has the disadvantage that there is considerable radiation from the feeder. This can be overcome by making it into the balanced system shown in Fig. 13.83(b) so that the feeder radiation is cancelled. This aerial can alternatively be regarded as a development of the centre-fed aerial with tuned feeders, the spacing of the taps helping to reduce the v.s.w.r. on the feeder. The length of the arms of the *V* section of the line should be at least equal to the separation between the taps, preferably more, and the overall length from one end of the aerial into the station should be chosen as for the centre-fed aerial in order to give reasonable tuning arrangements; it may present some difficulty on 21 Mc/s because of the high standing-wave ratio.

Asymmetrical Twin Feed

It will be realized that a long-wire aerial, i.e. one which is two or more half wavelengths long, can be fed with low-impedance twin line at any current-maximum position, such as the point a quarter-wave from one end (**Fig. 13.84**). In such a case 70–80 ohm twin line could be used and would match the aerial well enough, though the impedance at such a point in the aerial is somewhat greater than that of a single half-wave. This could be done with a 7 Mc/s aerial, and then at 21 Mc/s the feeder would again be at a current maximum position. Other bands have an even harmonic relationship, and the feeder would be badly mismatched.

One way in which this problem has been solved is to connect the feeder into the aerial at an intermediate position at about one third of the way along the wire where the impedance is higher, and to use 300 ohm twin line. This method introduces a fresh difficulty in that, although the theoretical v.s.w.r. on the line may not be too high, the feeder is no longer balanced because it is connected into an asymmetrical point of the aerial. The difference between the currents in the two wires of the feeder is equivalent to a single vertical radiating component, so that there is some radiation from the feeder. Nevertheless this system is an improvement over the single-wire feeder.

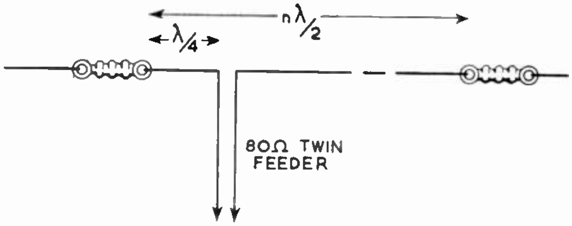


Fig. 13.84. Asymmetrical twin line feed for a harmonic aerial.

Long Wire Horizontal Aerials

The basic patterns of aerials of various harmonic lengths are shown in **Fig. 13.85** and further information is given in **Table 13.5**. It will be seen that there are two lobes in the diagram for each half-wavelength of wire, and that those nearest the end-on direction are always strongest. It should be remembered that these diagrams are sections, and the reader should try to visualize the lobes as sections of cones about the aerial.

Polar diagrams represent effectively the azimuth directivity of these aerials, but their general assessment must include the effect of height, and this can be done by considering them in relation to the vertical plane diagrams of Fig. 13.70. There is, however, one very important feature: in conjunction with the ground these aerials produce some radiation off the ends.

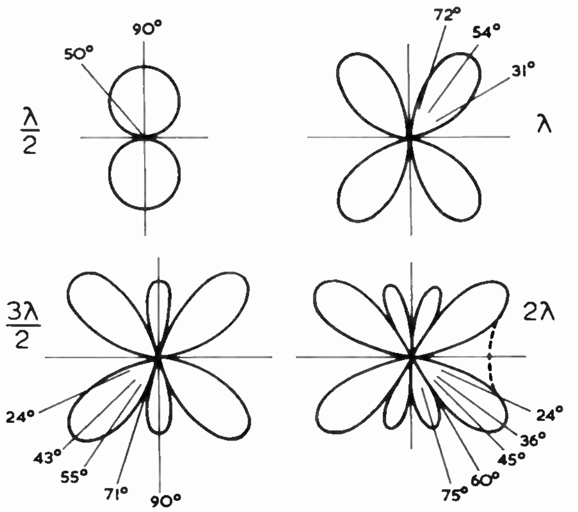


Fig. 13.85. Theoretical polar diagrams for wire aerials up to 2λ in length. The angles of the main lobes and crevasses are shown, also the angles at which the loss is 3db in the main lobe. The lobes should be visualized as cones about the wire. When the aerials are horizontal, end-fire radiation can take place at useful wave angles, especially from very long wires, e.g., the broken line on the aerial. Details of long wire aerials are given in Table 13.5.

For a dipole one half-wavelength high, the end signal may be only 3db (one S point) less than the broadside at 30° elevation, but the effect is most useful in longer radiators, where the end pattern almost fills in for elevations of 15–30°, as shown on the 2λ line in Fig. 13.85. In the case of a three wavelength long aerial (which would be about 100 ft. for 28 Mc/s) the end radiation at a 15° wave-angle may be much better than the broadside signal from a dipole at the same height.

TABLE 13.5
Properties of Long Wire Radiators

Length	Angle of main lobe to wire	Gain of main lobe over half-wave dipole	Radiation resistance
1 λ	54° (90°)	0·4db	90 ohms
1½ λ	42°	1·0db	100 ohms
2 λ	36° (58°)	1·5db	110 ohms
2½ λ	33°	1·8db	115 ohms
3 λ	30° (46°)	2·3db	120 ohms
4 λ	26° (39°)	3·3db	130 ohms
5 λ	22° (35°)	4·2db	140 ohms
6 λ	20° (31°)	5·0db	147 ohms
8 λ	18° (26°)	6·4db	153 ohms
10 λ	16° (23°)	7·4db	160 ohms

The number of complete conical lobes (see Fig. 13.86) is equal to the number of half-waves in the aerial. The main lobe is the one nearest to the direction of the wire, and the figures in this table gives its direction and gain. When a multiple full-wave aerial is centre-fed the pattern is like that of one half, but with more gain in the main lobe. The angles in brackets correspond to this case. When the aerial is terminated, or self-terminating, the radiation resistance is 30 to 50 per cent greater, and the main lobe slightly nearer to the wire.

The long wire aerial is the simplest form of directive aerial and is very popular because it is at the same time a good all-round aerial. Although the notch in the pattern in the broadside direction is quite noticeable, in practice on the even-multiple aerials the effect of the other crevasses is not so marked, chiefly because their direction varies with wave angle, so there is always a way through for signals.

Table 13.5 gives important properties of wires up to 10 wave lengths long. It will be seen that a 10λ aerial has a main lobe at only 16° angle to the wire, with a gain of over 7db. Remembering that these lobes exist all round the aerial, e.g. in the vertical as well as the horizontal plane, it will be realized that in the direction of the wire there is a vertical polar diagram with a lobe maximum at 16° elevation. It is found that the ground reflection effect is small for this direction, and although the polarization is vertical, a long wire aerial gives good radiation off its end, and can be considerably better than a half-wave dipole at the same height above ground. (It is this effect which sometimes causes confusion over the use of the term *end-fire* which refers to a particular type of array and not to the direction of a main lobe with respect to a wire.)

The radiation resistance figures in Table 13.5 are the resistance at any one current maximum, say $\lambda/4$ from the end, and are representative free space values, fluctuating somewhat with height as in Fig. 13.80. For example, the radiation resistance of a dipole is very low when it is close to the ground but rises rapidly to over 90 ohms at a height of $0{\cdot}3\lambda$, drops to about 60 ohms at 0·64 and then settles about the free space value of 73 ohms for greater heights. The longer aerials behave in a similar way. From Table 13.5 it will be seen that the resistance increases steadily as the size of the radiating system increases; this is a general rule for all large directive aerial systems.

Effect of Feed Position

The general subject of how to feed energy to a long wire has already been discussed and it was indicated that it could be fed with a single wire at one end or near one end or by tuned feeders in the centre. In each case the system is in fact a multi-band aerial. A balanced line can also be used a quarter wavelength from one end but in this arrangement the aerial is essentially a one-band device. The main distinction to be considered here is whether the feed is at a current or voltage maximum, because these alternatives may give two entirely different aerials.

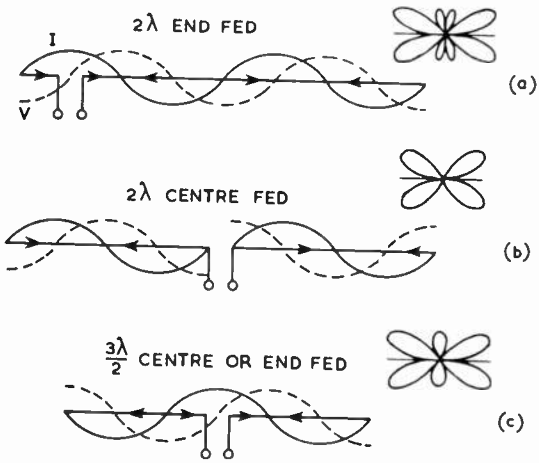


Fig. 13.86. Illustrating the difference between feeding at a current, or at a voltage maximum, and how centre-feed affects even, and odd-multiple half-wave aerial differently. When the feed enters a current loop the pattern is the same as with end-feed, but when a voltage loop is entered, the pattern is like that of one half only of the aerial.

The long wires described above were continuous with alternate positive and negative current loops along them. This is the situation with end-feed. It is also the same with centre-feeding to an odd number of half-waves because the feed enters a current loop. But if a wire an even number of half wavelengths long is broken at the centre at a voltage maximum and fed with balanced line (**Fig. 13.86**) an extra phase reversal is introduced at this point, as can be seen by sketching the standing waves. Thus a centre-fed three wavelength aerial becomes an array of two $3\lambda/2$ aerials and has a pattern which is basically that of the $3\lambda/2$ aerial multiplied by an array factor corresponding to the $3\lambda/2$ spacing between the centres of the two halves. Since this particular array factor is very nearly the same as the pattern of each half, the final diagram is very like that of a single $3\lambda/2$ aerial, but with emphasis on the broadside lobe. By comparison the 3λ end-fed aerial has a null in the broadside direction.

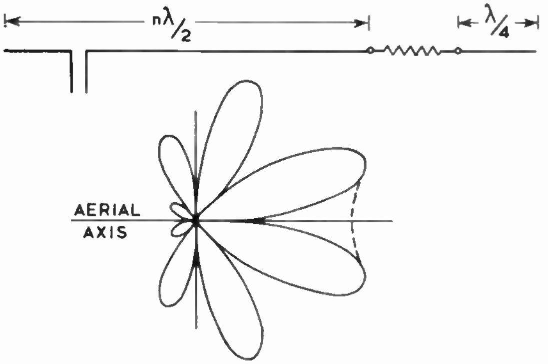


Fig. 13.87. A 2λ terminated aerial and its radiation pattern, showing end-fire effect. The end-to-end ratio depends on several factors, but may be 3 to 5db for a full-wave unterminated aerial and more than 10db for a terminated aerial. The 500 ohm terminating resistor is earthed to a $\lambda/4$ artificial earth. A multiple " earth " fan may be used to cover two or three frequency bands. These aerials do not need critical length adjustment.

End-fire Effect

In practice a long wire fed at one end tends to radiate best from the opposite end, the pattern of a 2λ aerial tending to become like that of **Fig. 13.87** (these patterns are ideal patterns). This tendency is greater the longer the wire, and occurs because the lobes to the right (in Fig. 13.87) are due to the radiation from the forward wave along the wire, whilst those toward the left can only be due to the wave reflected from the far end. Because of loss by radiation, the reflected wave is weaker than the forward wave; thus a long wire tends to behave like a lossy transmission line. The one-way effect can be enhanced by joining the far end of the aerial to a quarter wave " artificial earth " wire through a 500 ohm (matching) resistor. This resistor must be able to absorb about 25 per cent of the transmitter power if the aerial is only 2λ long, but only about 10 per cent if it is very much longer. The power lost in this resistor is not wasted because it would all have gone in the opposite direction. Such an aerial is of course only correctly terminated on one band, though a fan of earth wires can be used, one for each band. It can be fed (for one band use) by means of an 80 ohm feeder one quarter wavelength from the free end as in Fig. 13.87. For multi-band working tuned feeders at the free end may be used.

The " Zepp " Aerial

A simple wire aerial can be made to work satisfactorily on a number of harmonically related bands by feeding it at one end. This will always present a high impedance and can be connected to open wire tuned lines. The operation of such lines is explained in the section dealing with Transmission Lines. Such an aerial is known as a Zeppelin or " Zepp," and is illustrated in **Fig. 13.88**, which gives suitable dimensions for both the aerial and feeder.

The system is based on the concept that if the two feeder wires carry equal and opposite currents, they will not radiate. In practice it is difficult to secure this condition with the same aerial on all bands. Multiple quarter- or half-wavelength long feeder lines are unsuitable for the same reason and also because they make adjustment of the aerial coupler difficult. A length of 45 ft. will provide an intermediate impedance on all bands, and will work with close or medium spaced taps or series tuning on all bands except 21 Mc/s. At the latter frequency a high impedance will be presented by the feeder. Open wire 600 ohm line is preferable, because the v.s.w.r. on the line is high. The aerial should be made just over one half wavelength long at the lowest frequency to be used, say 67 ft. for frequencies from 7 Mc/s upwards (or 135 ft. if 3·5 Mc/s is to be included) and adjusted in length until the feeder currents are equal at the *lowest* frequency. If the half wavelength top cannot be erected for the lowest frequency band the free wire of the feeder (the one not connected directly to the aerial) may be disconnected at the transmitter end and the remaining wire plus the aerial used as an end-fed arrangement.

If a feeder length other than 45 ft. is employed it may be necessary to modify the tuning arrangements on each band to accommodate the changed impedance presented to the coupler. The required form of tuning (series or parallel) can be obtained from Fig. 13.54 on page 13.32, where the length L refers to the *feeder only*. This follows from the knowledge that the *end* impedance of the aerial will always be high, or a voltage maximum, irrespective of the length or band. For the same reason it is permissible to alter the length of the radiating top by $\pm$ 10 per cent and still resonate the system by adjustments to the coupler, although best operation will result when the top itself is resonant, i.e., an exact multiple of a halfwave at each operating frequency.

For the parallel coupler, the capacitor C should have a maximum value of at least 100 pF for operation down to 3·5 Mc/s, and 25 pF if the lowest band is 14 Mc/s. The value of the corresponding capacitors in the series coupler is approximately one half of these figures. The inductance value required can then be determined from a suitable Abac relating frequency, L and C. The link coupling should be approximately one tenth of the main coil.

The Extended Double Zepp

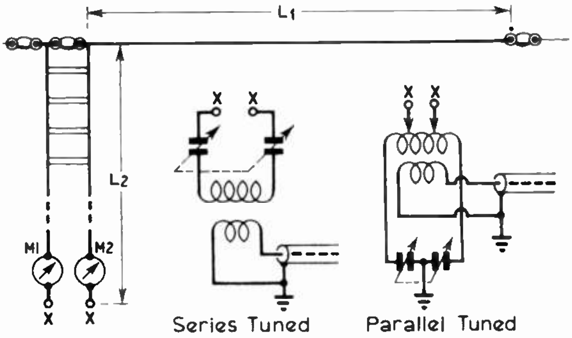
This aerial system is simple to erect and adjust and gives a gain of approximately 3db over a halfwave dipole.

The horizontal polar diagram is similar to that of the dipole but with somewhat sharper main lobes which are at 90° to the run of the wire. In addition there are four minor lobes at 30° to the wire. At twice the frequency or more the pattern becomes similar to that of one and two wavelength aerials.

The system is a type of collinear aerial (see page 13.6). The layout is shown diagrammatically in **Fig. 13.89** and comprises two lengths of wire each 0·64 λ long, fed in the centre with an open wire line. For low power, and where the line length is short, 300 ohm ribbon can be used.

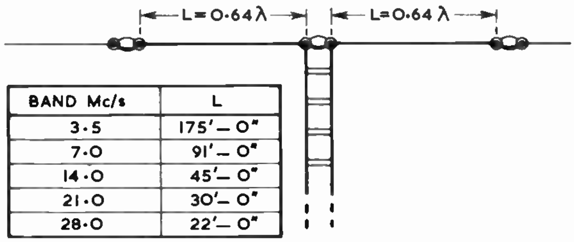
The table of Fig. 13.89 gives the design length for the centre of different bands.

A feeder length of 45 ft. is again a good compromise and the coupling arrangements described for the ordinary Zepp are equally suitable. The form of coupling (series or parallel) can again be found from Fig. 13.54. In this case the length L should include the half-top of the aerial.



AERIAL LENGTH L1	FEEDER LENGTH L2	BAND Mc/s	TUNED
135' – 0"	45' – 0"	3·5, 7, 14, 28	SERIES
135' – 0"	45' – 0"	21	PARALLEL
67' – 0"	45' – 0"	7, 14, 28	SERIES
67' – 0"	45' – 0"	21	PARALLEL

Fig. 13.88. The end-fed Zepp aerial employs tuned feeders and an aerial coupler to achieve correct operation on each band. Currents in the meters M1 and M2 should be equal within $\pm$ 10 per cent for balanced operation on the line.



BAND Mc/s	L
3·5	175'– 0"
7·0	91'– 0"
14·0	45'– 0"
21·0	30'– 0"
28·0	22'– 0"

Fig. 13.89. The Extended Double Zepp aerial employs centre feed with tuned lines.

The Multee Two Band Aerial

This aerial was developed by W6BCX and works on two bands. On the lower band it performs as a top loaded vertical and on the higher frequency band the top section forms a quarterwave folded dipole with the vertical section forming a quarterwave matching transformer.

Two important points should be noted. If an earth is used it must be a good one and the vertical portion should be as near vertical as possible. The layout is shown in Fig. 13.90 and the dimensions are as follows:

	1·8/3·5 Mc/s	3·5/7·0 Mc/s	7/14 Mc/s
L1	65 ft.	33 ft.	17 ft.
L2	54 ft.	27 ft.	13 ft. 6 in.
L3	56 ft.	25 ft.	12 ft.

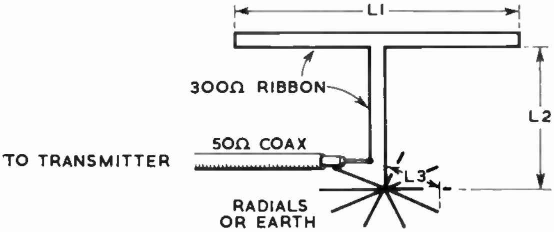


Fig. 13.90. The W6BCX Multee two band aerial.

300 ohm Ribbon Aerials

A simple multiband aerial is shown in Fig. 13.91 and can be fed with 75 ohm flat twin or 75 ohm coaxial cable. L1 acts as a halfwave dipole on 7 Mc/s and as three halfwaves on 21 Mc/s; L2 functions as a halfwave dipole on 14 Mc/s.

The lengths of L1 and L2 can be calculated from the formula 468/F (Mc/s), the answer being in feet. Dimensions for L1 and L2 are as follows when cut for mid-band operation:

L1 Length	Band	L2 Length	Band
246 ft.	1·8 Mc/s	128 ft.	3·5 Mc/s
128 ft.	3·5 Mc/s	66 ft. 5 in.	7 Mc/s
66 ft. 5 in.	7 & 21 Mc/s	33 ft.	14 Mc/s
33 ft.	14 Mc/s	22 ft.	21 Mc/s
22 ft.	21 Mc/s	16 ft. 3 in.	28 Mc/s

Any combination of L1 and L2 can be chosen to cover the particular bands required.

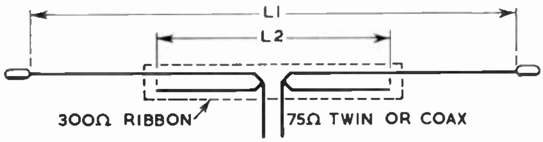


Fig. 13.91. A two band 300 ohm ribbon aerial. Dimensions are given in the text.

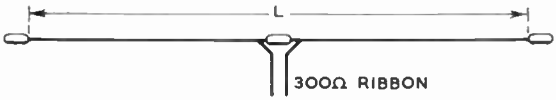


Fig. 13.92. The DJ2ZF multiband dipole.

The DJ2ZF Multiband Dipole

This is a simple three band aerial the layout of which is shown in Fig. 13.92 and was originally described by DJ2ZF. The dimension *L* can be 167 ft. and when fed with a 300 ohm line gives satisfactory match on 3·5, 7 and 14 Mc/s.

If *L* is reduced to 83 ft. 6in., then this aerial can be used on 7, 14 and 28 Mc/s.

This type of aerial can be fed with a Z match coupler or a balun as shown in Fig. 13.93.

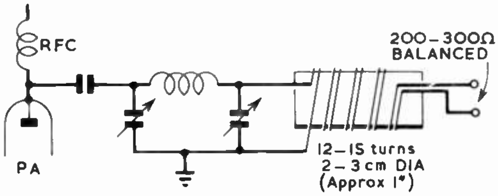


Fig. 13.93. The aerial coupler for the DJ2ZF aerial of Fig. 13.92. This provided a balance-to-balance connection for single ended pi-net-work circuits.

F7FE All Band Dipoles

A way of connecting and laying out four separate dipoles to cover 3·5, 7, 14, 21 and 28 Mc/s was worked out by F7FE and used by him with great success. Its great merit is that it only uses two supporting masts and is fed by a single 75 ohm flat twin feeder or coaxial cable.

Only four dipoles are used because the one cut for 7 Mc/s works well on 21 Mc/s.

The layout is as shown in Fig. 13.94 and is more or less self-explanatory; each dipole is cut to mid-band using the formula 468/F (Mc/s).

The centre insulating plate is made up in polystyrene sheet and has six holes drilled in it. The top pair take the 28 and 7 Mc/s dipoles and the centre two the 3·5 Mc/s dipole and the bottom two take the 14 Mc/s dipole.

A short length of wire connects all the left-hand quarter-waves in a similar manner. The 75 ohm feed is connected to the bottom end of these pieces of wire.

BEAM AERIALS

Directive or beam aerials may be divided into two main classes: those using arrays of half wave dipoles and those based on the known directive properties of long wires. In all cases the elements are arranged so that radiation is cumulative in the favoured direction but is minimized in other directions. The first type can be sub-divided into broadside arrays with the main beam at right angles to the plane of the array and end-fire systems where the main beam is projected along the plane of the elements.

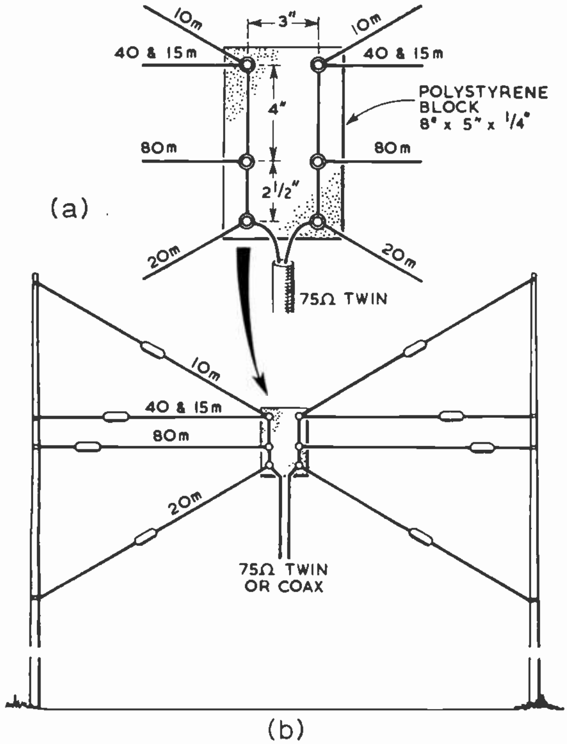


Fig. 13.94. The F7FE all-band dipole aerial.

In broadside arrays the elements are connected together by phasing lines, so that they are all in one phase, but in the end-fire aerials the elements may be all connected to give a progressive phase change along the array (driven arrays) or there may be one driven dipole together with a number of nearly resonant free *parasitic* elements which modify the local field of the radiator so that a unidirectional end-fire pattern is produced (*parasitic arrays*). Parasitic reflectors can also be added to broadside arrays to produce a unidirectional beam instead of the fore-and-aft pattern of a single row of elements.

Although these two general classes of array are both effective in producing directivity and gain, their construction and properties are quite distinct. The arrays are resonant and interconnected by tuned lines and are therefore usually limited to one-band use, although special arrangements can be devised which will operate well on two or even three bands, at the cost of having to use tuned feeders. On the other hand their patterns are well defined and relatively free from spurious lobes; in particular the vertical and horizontal patterns of broadside arrays can be controlled separately, since they depend on the height and width respectively. In end-fire arrays this is not possible because the vertical and horizontal patterns both depend on the length. Arrays are, however, fairly complex mechanical and electrical structures and are best suited for fixed direction working; amateurs can usually only build simple arrays for h.f. work, such as the *lazy-H* or the two and three element *Yagi* arrays and although these can be made rotatable, it is in the v.h.f. field where the greatest advantage can be taken of multi-element arrays or stacks.

Long wire aerial arrays, described on p. 13.51, are much easier to construct and to make work, and can be erected in situations quite unsuitable for multielement arrays because they require few supports. They also have the advantage that they can be used over wide frequency ranges, say four amateur bands, though the optimum performance is limited to a 2 : 1 frequency range. Their patterns are not so well defined as other arrays, a difference very noticeable in reception, and, like the end-fire arrays, their vertical and azimuthal patterns are determined simultaneously by the height and length. They can give a remarkable performance in the main beam direction, but tend also to radiate a good signal in all directions, which is to some extent an advantage in transmission but a disadvantage in reception. Long wire arrays do not use reflectors because they can be made unidirectional by terminating them with matching resistances.

In selecting arrangements for amateur work, first consideration should be given to acquiring a low wave-angle for improved long-distance propagation; the effective gain which can be obtained in this way is much greater than is apparent from the radiation patterns alone. The azimuthal pattern should never be too sharp for reception because the amateur requires to search. This leads to a preference for end-fire aerials; where there is space to erect them, long wire types are best because of their simplicity and effectiveness, but where space is limited two-, three- or four-element Yagis are used.

Broadside Arrays—General

Fig. 13.95 gives a selection of simple broadside arrays together with gain and average radiation resistance figures. A variety of feed connections are shown, which are interchangeable as discussed below, and it is apparent how to extend the arrays beyond the number of elements illustrated. For all these arrays the azimuth patterns can be estimated as a function of the length of the array from **Fig. 13.96** and the broadside vertical patterns, as a function of height from Fig. 13.70.

In broadside arrays, the elements are all in phase and the interconnecting *phasing lines* or stubs must be adjusted to secure this condition. The spacing between elements and between the centre of the elements need not be one half-wave, but can vary up to $\frac{3}{4}\lambda$, beyond which minor lobes in the pattern become too large to ignore. The choice of spacing is, however, in practice dependent on the type of phasing line used. The position of the feed-point depends on the input impedance required, or on the fact that some of the simpler arrays will operate on more than one band. A centre feed position should be used if possible, especially in long arrays, because power is being radiated as the currents travel along the array, and the more distant elements may not receive their proper share. Uneven power distribution can cause the beam to broaden and " squint."

Collinear Arrays

The collinear array is the simplest method of obtaining a sharp azimuthal pattern and is simply a row of half-wave radiators strung end-to-end. To bring all elements into phase it is necessary to provide a phase reversing stub between the high voltage ends of each pair, except where the position is occupied by the feeder.

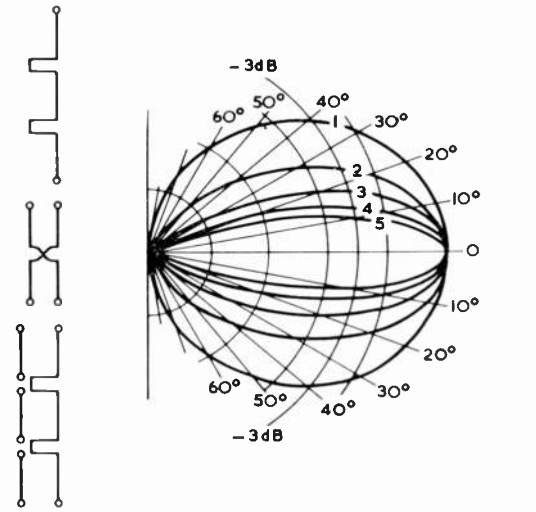


Fig. 13.96. Horizontal polar diagrams of collinear and similar arrays, for one to five half-waves overall length, showing half-power points (−3 db). Without a reflector the patterns are bi-directional. With a reflector, or in the case of a W8JK aerial, the patterns are only very slightly sharper, because the forward pattern of a dipole and reflector in this plane is very little different from that of a dipole. Minor lobes are not shown but should be −14db or lower in a well adjusted array. The pattern of a 1·25λ dipole is slightly sharper than that of a full-wave dipole, but has minor lobes at −10 db level at about 60°. Vertical patterns of vertically stacked dipoles are also represented by the upper half of the above curves.

The simplest form of array is the centre-fed full-wave or *double Zepp* with a high impedance tuned line feed at the centre. In this form it can be used on other bands; when the total length is only one half-wave the impedance is between 60 and 100 ohms, but in the full-wave condition it is 5000 ohms or more. When the frequency is raised to the value giving three half-waves the impedance is about 100 ohms, whilst at two full-waves it is about 3000 ohms. The high impedances can be lowered to between 1000 and 2000 ohms by using a flat top of twin wires 2–3 ft. apart, joined in parallel, in order to improve the v.s.w.r. on a 300 ohm line; this does not alter the low impedance value. The broadside beam would only occur on the bands corresponding to half- and full-wave; on other bands the pattern would be that of the long wire type.

Tuning and Matching Collinear Arrays

In order to match a full wave aerial, a closed quarter wave stub may be added at the centre and 80 or 150 ohm twin connected into the centre of one dipole, or 300 or 600 ohm line tapped into the stub (Fig. 13.44(b)). The array needs tuning when a stub is used, and for this purpose the stub is made a little too long and a moving short-circuit provided. It may be possible to couple a grid dip oscillator into the bottom of the stub to find resonance. Approximate dimensions (in feet) are given by $470/f$ (Mc/s) for the radiators and $240/f$ for each leg of the stub.

The full-wave radiator can be extended to about 1¼ wavelengths (*extended double Zepp*) in which case the gain is increased to 3db and the v.s.w.r. reduced to about 6 on a 600 ohm line, and may be reduced further if a tuning stub, about ⅛ wave long is added in parallel at the feed point. **Fig. 13.95(b)** is a useful variation in which the two halves are separated, because the two equal low impedance feed lines can be connected in parallel, in or out of phase by means of a plug connector in the station, giving broadside or full-wave pattern at will. When the gap is $\lambda/2$ the pattern may be found by multiplying the λ-patterns of Fig. 13.64 by $\sin\phi$ where ϕ is the angle to the wire (see *Array Factors*).

Longer arrays may use ½ to ⅝ wavelength elements with tuning stubs and a feed point either at the centre of one element (current maximum) or at a phase reversal point (voltage maximum). The impedance at the centre of any element rises rapidly with the number of elements but falls rapidly at the phase reversal position, the two meeting at about 1200 ohms which is the characteristic impedance of a balanced aerial. These longer aerials should be tuned up section by section as described above, the total length of wire in fact between centres of shorting bars being $950/f$ (Mc/s).

Horizontal Broadside Arrays

These aerials, consisting of two horizontal arrays one above the other, give more gain than the collinear arrays for the same number of dipoles. At low wave angles their azimuthal patterns are the same as for collinear arrays, but the vertical pattern has only one main lobe whatever the height. By comparing the pattern of Fig. 13.70(d) for a dipole one wavelength high with Fig. 13.70(e) for $h = \lambda/2$ (total height λ), it will be seen that all the energy is in the low angle lobe, and this gives greater effective gain and improvement of signal stability at long distances.

Three Band Array

The basic model using two half-wave dipoles (**Fig. 13.95(c)**) can be used as a broadside array over a 2 : 1 frequency band —e.g. 14, 21 and 28 Mc/s. For this purpose the vertical spacing at the lowest frequency should be ⅜ wavelength so that at the upper frequency it will be ¾ wavelength. The phasing line should be 600 ohms (not crossed) with the feed point at its centre. Since it is not practicable to match over three bands it will be necessary to use tuned lines. The v.s.w.r. on 300 or 600 ohm main feeder will be between 6 and 10 and cannot easily be improved. There is nothing critical about any dimensions on this aerial. Methods of matching it to a line on any one frequency are considered later in this section. Arrays of four or more elements are, of course, one-band aerials.

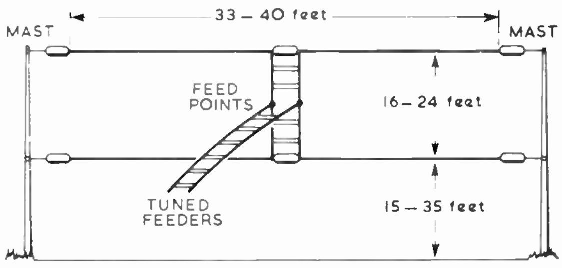


Fig. 13.97. A Lazy H array for 14, 21 and 28 Mc/s.

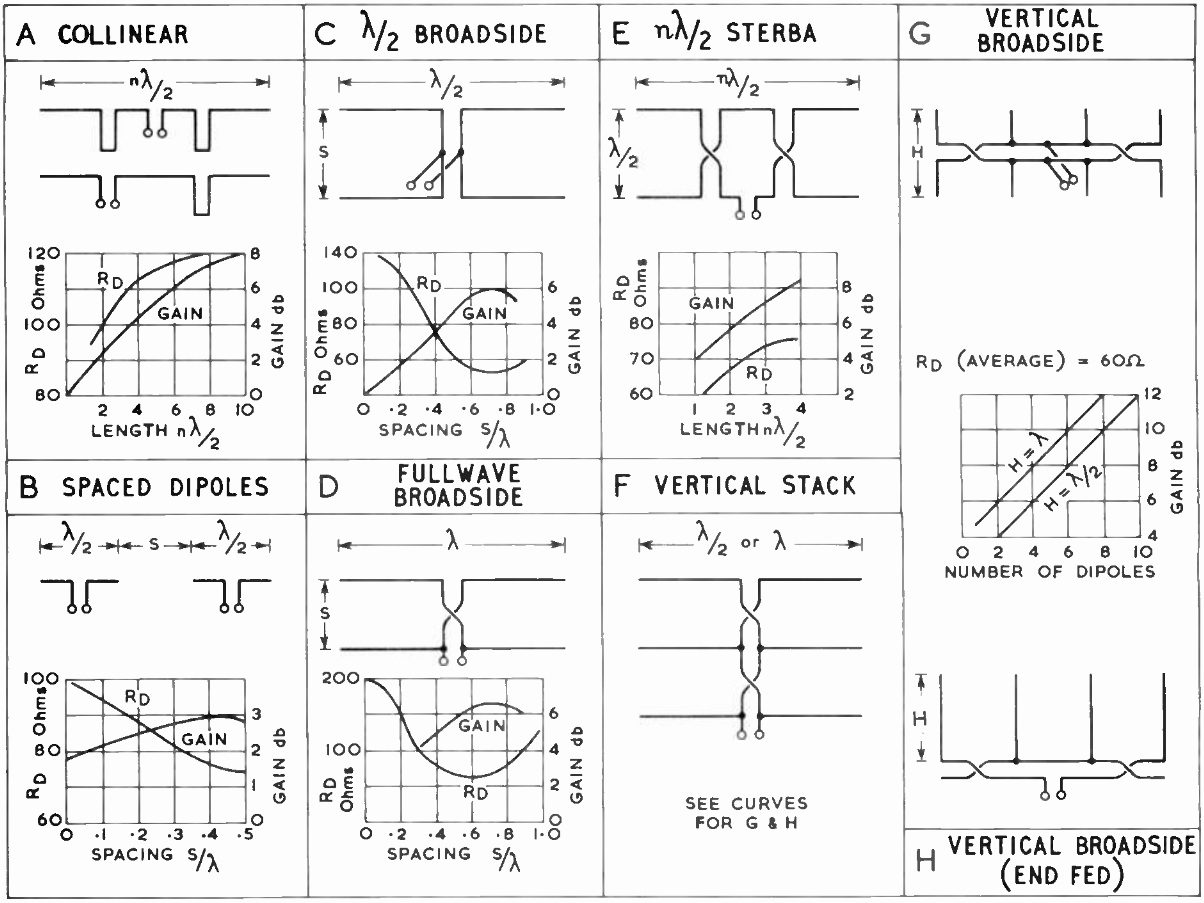


Fig. 13.95. Four general types of broadside array. (a) Collinear arrays. (b) End-spaced dipoles. (c, d, e) Two-tier, Sterba or Barrage arrays. (f) Pine tree or Koomans, stacked horizontal λ/2 or λ dipoles. (g, h) Vertically polarized broadside arrays. Gain figures are with reference to a free-space dipole, in terms of spacing or total length in half-waves. Resistance figures are average over the array, and are added in series or parallel according to the feed arrangements, as described in the text. Various feed positions are shown, and details are given in the text. The aerial in (c) can be arranged to give a broadside beam over a 2 : 1 frequency range, e.g. 14, 21 and 28 Mc/s.

Lazy-H

When the array is one wavelength long it is called a *lazy-H* and the effective gain is quite high especially if the upper wire can be placed at a height of one wavelength. With the phasing line connected as in Fig. 13.95(d) it must be electrically a half wavelength long and crossed to restore phase. The impedance across the bottom end of the phasing line is then of the order of 3000 ohms and a λ/4 stub transformer would be the best matching arrangement (Fig. 13.95(b)). On the other hand, if it is connected as in Fig. 13.95(c) with a centre tap feed, there is no need to cross the line as its two branches are always in phase. This connection makes use of the quarter-wave transformer effect in the 600 ohm phase line and will reduce the feed point resistance to around 70 ohms so that a low impedance feeder can be used.

A version of the lazy H which can be used on 14, 21 and 28 Mc/s is shown in **Fig. 13.97.** The gain will not be as high as in the basic version but it is nevertheless a worthwhile bi-directional fixed array.

The length *L* should be 33 to 40 ft. and the spacing 16 to 24 ft. The larger dimensions will give slightly more gain on all bands. The height of the lower wire should not be less than 15 ft. and best results will be obtained if it is 30 to 35 ft. above ground.

Sterba Curtain

Longer arrays have, of course, sharper patterns and greater gain, generally up to 4db more than a collinear array of the same length. The loading due to radiation is also high so that a six-element array with series feed (Fig. 13.95(e)) would have effectively the input impedance of six dipoles in series, say 500 ohms, whilst an eight element array fed at the base of the centre phase line would be like four full-wave centre fed aerials in parallel, about 800 ohms. In either case, the v.s.w.r. on a 600 ohm line would be low enough without extra matching. The element length (feet) should be about 470/*f* (Mc/s) and the phasing line on these larger aerials may be 600 ohms open wire or 300 ohms ribbon and of resonant length.

Phasing Lines

When the phasing lines are part of a series connection (e.g. Fig. 13.95(d) and (e)) they must be electrically one or more complete half wavelengths long. When the length is $\lambda/2$ they must be transposed to recover the phase reversal due to the wave travelling along the length of line, but if they are one wavelength long the phase is restored and the lines are not crossed. However, if the feeder is tapped into the middle of the phasing line as in Fig. 13.95(c) the current divides in phase regardless of its length and no cross-over is needed in that phasing line.

The velocity factor of open wire lines is nearly unity and the length factor for a half-wave is 0.48 0.49 λ; thus it is usually necessary to space the upper and lower rows by $\lambda/2$ in order to achieve a practical construction. When 300 ohm ribbon is used advantage can be taken of its velocity factor of 0·8 and the aerials spaced vertically by $\frac{3}{8}$ wave with an electrical half-wave of ribbon ($365/f$) or $\frac{3}{4}$ wave with a full-wave of ribbon. On the other hand 300 ohm ribbon is not so good for small arrays because in such applications it carries a high v.s.w.r.

Broadside Verticals and Stacks

The vertical broadside arrays are attractive because they have the low angle radiating properties of vertical aerials, but they require elaborate supports and are thus not likely to be used except on 21 or 28 Mc/s. Vertical patterns of such arrays are given in Fig. 13.70(a) (b), and horizontal patterns for two elements in Fig. 13.64. The patterns are broader than those of the equivalent collinear arrays. The half-wave dipole impedances with mutual effects tend to average 60 ohms.

The stack of horizontal dipoles is really the same aerial rotated. Its azimuth pattern is that of a single horizontal element; the vertical pattern is not illustrated, but improves on that given in Fig. 13.70(c), as the number of elements increases. In these arrays advantage can be taken of the velocity factor of 80 ohm line (0·67) or 300 ohm lines (0·8) to make the phasing lines one wavelength long (uncrossed) and increase the spacing of the elements to $2/3\lambda$ or $3/4\lambda$, but the 80 ohm line should not be used with full wave dipoles because of the high v.s.w.r. it would carry.

"Flat Top" or W8JK End-fire Array

These aerials use two anti-phase radiators side by side **(Fig. 13.98)** and force the broadside radiation to low angles by cancelling the radiation in upward directions. The vertical patterns are given in Fig. 13.70(f) It will be seen that the wave angle drops as the height is raised, but the pattern does not break up at certain heights, as does that of a single wire, and so the height is not critical.

The radiators may be separated by light wooden spreaders, hanging from two supports, and may be half wave, full wave, or extended like a collinear array. With a two or four element flat top array, the elements are half wavelength long and not critical, but in a longer array the distance between cross-over centres or between cross over centre and free end must be a resonant length. Alternatively, the phasing line can be made half-wave long (uncrossed) using 600 ohm line, and left to hang **(Fig. 13.99)**. The dipole lengths are slightly affected by mutual impedances: a suitable length (feet) is

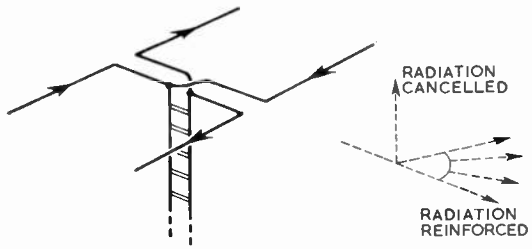


Fig. 13.98. The basic W8JK flat-top array. The antiphase currents in the elements cancel any vertical radiation and reinforce in the plane of the array.

$460/f$ (Mc/s) for dipoles $480/f$ for full wave aerials in these arrays.

The theoretical gain relative to a half-wave dipole is 2·5db at half-wave spacing between elements rising to nearly 4db at very small spacings. These gains are added to the collinear gain if more than two elements are used. Close spacing is attractive because apart from the higher gain the aerial is easier to support and mechanically more stable, but it will be seen from **Table 13.6** that radiation resistance drops to very low values at small spacings. This means large standing waves, high currents and voltages, and intense local field, which all result in power loss through conductor resistance or dielectric losses in insulators and spreaders. As a result the theoretical gain is never reached, but 3·5db can be obtained if precautions are taken.

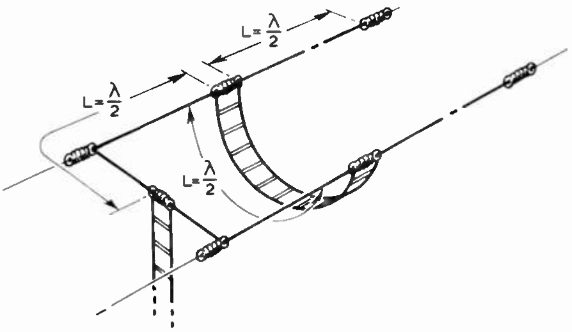


Fig. 13.99. An end-fed single section W8JK using a slack half-wave-Length phasing line which is not transposed, unlike the short line of Fig. 13.98(d).

The directivity of the W8JK is also a function of the spacing, being greatest at close spacings. If the array is fed exactly at the centre of the phasing line as shown in **Fig. 13.100**, the horizontal pattern will be bi-directional. If, however, the feed point is moved to the centre of one of the elements, the array will become uni-directive, with maximum radiation in the direction of the "driven" element. This arises because the current in the other element then lags from the ideal 180° phase differing by an amount corresponding to the electrical length of the phasing line.

A flat top beam comprising two half-wave dipoles spaced

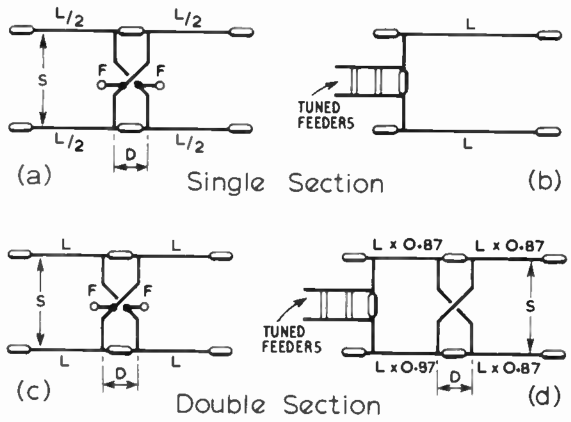


Fig. 13.100. Single and double section W8JK arrays using centre and end feeding. Dimensions are given in Table 13.8 for different bands.

0·15 to 0·2λ at 14 Mc/s can be used as a full wave array on 28 Mc/s, where spacing would become 0·3 to 0·4λ and although the v.s.w.r. on the line may be high on one or more bands the aerial will at the same time perform as a beam on 21 Mc/s.

Table 13.6 gives dipole centre- and end-impedances for half- and full-wave arrays; it will be seen that with spacing less than 0·2λ power loss can easily occur, due to large currents in the wire and high voltages across the insulators and matching to a line may be difficult. One way to deal with the extreme impedances is to remove them by folding the radiators two or three times. This multiplies the low centre impedance by four or nine, and reduces the high impedance by the same factor. The two branch feed lines can then be used as quarter-wave transformers using suitable cable to achieve the required feed point impedance. A few examples are given in Table 13.7. An alternative method of matching using transformers in the line would be found difficult due to the high v.s.w.r.

TABLE 13.6

Impedance Values for W8JK Aerials

Spacing S λ	L = λ/2		L = λ	
	Rd	Re	Rd	Re
0·1	6	40,000	10	50,000
0·15	12	20,000	20	25,000
0·2	20	12,000	30	16,000
0·25	33	7,500	50	10,000
0·3	46	5,500	65	8,000
0·4	64	4,000	100	5,000
0·5	85	3,000	125	4,000

Approximate theoretical impedances in ohms at two points of a W8JK array. Rd is the impedance at the centre of any half-wave element and Re the impedance to earth of any free end. Figures are based on an aerial characteristic impedance of 500 ohms.

TABLE 13.7

Input Impedance of W8JK Aerials

S/λ	λ/4 Branch Z_0	Input Impedance λ/2, 1-2-3-Fold			Fullwave Input Impedance
		1	2	3	
0·15	80	250	70	30	4,000
	150	900	250	100	
	300	—	900	400	
	600	—	—	1,600	
0·2	80	160	40	—	2,500
	150	600	150	60	
	300	—	600	250	
	600	—	—	1,000	
0·25	80	100	—	—	2,000
	150	300	80	—	
	300	—	300	150	
	600	—	—	600	

A few phasing-line transformer arrangements for half-wave and full-wave W8JK aerials using single or folded elements. Figures are only approximate, and are based on an aerial characteristic impedance of 500 ohms. The full wave input impedances are for single wire elements.

Suitable dimensions for one, two and four section W8JK arrays are given in **Table 13.8** in conjunction with **Fig. 13.101.** The design includes a matching stub (see page 13.25) for feeding the array with aperiodic balanced 600 ohm twin wire line.

A simple practical design for single band single section arrays is shown in **Fig. 13.102** and **Table 13.9.** This utilizes 300 ohm twin wire line for the phasing sections. By taking advantage of the impedance step-up provided by three-wire folded radiators, the array is arranged to present an acceptable match to a main feeder also made of 300 ohm flat twin.

The ZL Special

A convenient and simple modified form of the W8JK can be constructed from 300 ohm ribbon on lightweight spreaders and fed with 75 ohm balanced line at the centre of one of the elements as shown in **Fig. 13.103.** The centre-to-centre spacing of the folded elements should be 0·15-0·2 wavelength to achieve a reasonable match to 75 ohm line. Coaxial cable can be used to feed this aerial if a suitable balun is connected at the feed point. The phasing line can conveniently be constructed also of 300 ohm ribbon, with connections transposed between the two folded elements, and taped along the centre support which together with the booms must, of course, be non-metallic (wood or bamboo is suitable). The elements are not fed in exact antiphase and since the spacing is significant, compared with a wavelength, the beam will possess some directivity, as shown by the arrow. The beam can be reversed, either by connecting the feed line to the centre of the other element (Fig. 13.103(a)), or by spinning the whole aerial over (Fig. 13.103(b)).

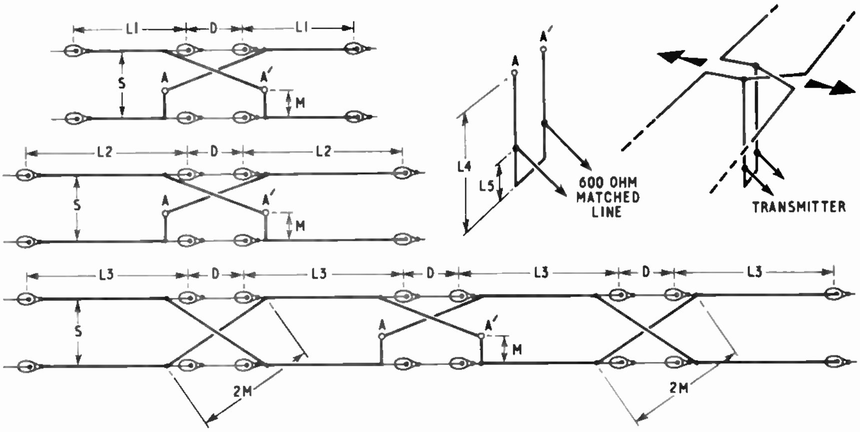


Fig. 13.101 Layout of 1, 2 and 4 section centre fed W8JK arrays. Inset is a quarter wave section enabling the array to be matched the untuned 600 ohm balanced line. The matching section must hang vertically down from the feed points A–A′. Dimensions are given in Table 13.8.

TABLE 13.8

Frequency	Spacing	S	L1	L2	L3	D	M	L4	L5
7·1 Mc/s	0·125λ	17 ft. 7 in.	34 ft.	60 ft.	52 ft. 7 in.	4 ft.	8 ft. 10 in.	33 ft. 8 in.	4 ft.
14·2 Mc/s	0·125λ	8 ft. 8 in.	17 ft.	30 ft.	26 ft. 3 in.	2 ft.	4 ft. 5 in.	16 ft. 10 in.	2 ft.
14·2 Mc/s	0·15λ	10 ft. 5 in.	17 ft.	30 ft.	25 ft. 1 in.	2 ft.	5 ft. 4 in.	16 ft. 10 in.	2 ft.
14·2 Mc/s	0·2λ	13 ft. 10 in.	17 ft.	30 ft.	22 ft. 10 in.	2 ft.	7 ft. 2 in.	16 ft. 10 in.	3 ft.
28·2 Mc/s	0·15λ	5 ft. 2 in.	8 ft. 6 in.	15 ft.	12 ft. 7 in.	1 ft. 6 in.	2 ft. 8 in.	8 ft. 6 in.	1 ft.
28·2 Mc/s	0·25λ	8 ft. 5 in.	8 ft. 6 in.	15 ft.	10 ft. 4 in.	1 ft. 6 in.	4 ft. 5 in.	8 ft. 1 in.	2 ft.

Note: L4 and L5 are approximate values. In particular L5 should be adjusted for minimum s.w.r. on the main feed line.

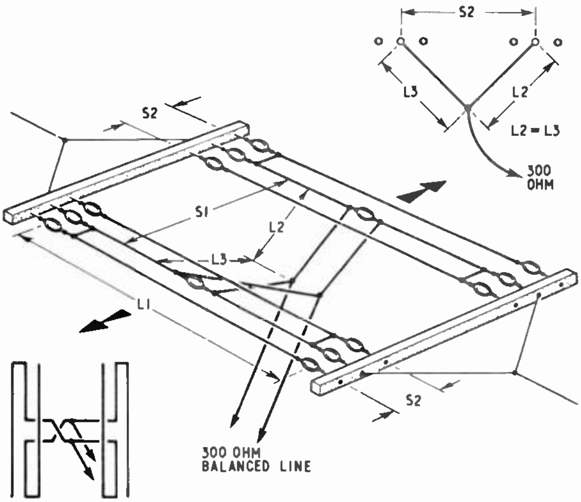


Fig. 13.102. A practical single section W8JK design. The phasing lines can be made from 300 ohm flat twin polythene cable if allowance is made for the velocity factor. Dimensions are given Table 13.9

TABLE 13.9

Frequency	S1	S2	L1	L2, L3 *
7·1 Mc/s	20 ft. 9 in.	14 in.	65 ft. 10 in.	27 ft. 7 in.
14·2 Mc/s	10 ft. 4 in.	12 in.	32 ft. 11 in.	13 ft. 9 in.
21·2 Mc/s	6 ft. 4 in.	12 in.	22 ft.	9 ft. 3 in.
28·2 Mc/s	5 ft. 2 in.	9 in.	16 ft. 7 in.	6 ft. 11 in.

* For air-spaced line. If flat polythene 300 ohm twin is used, these lengths must be multiplied by 0·82 (velocity factor)

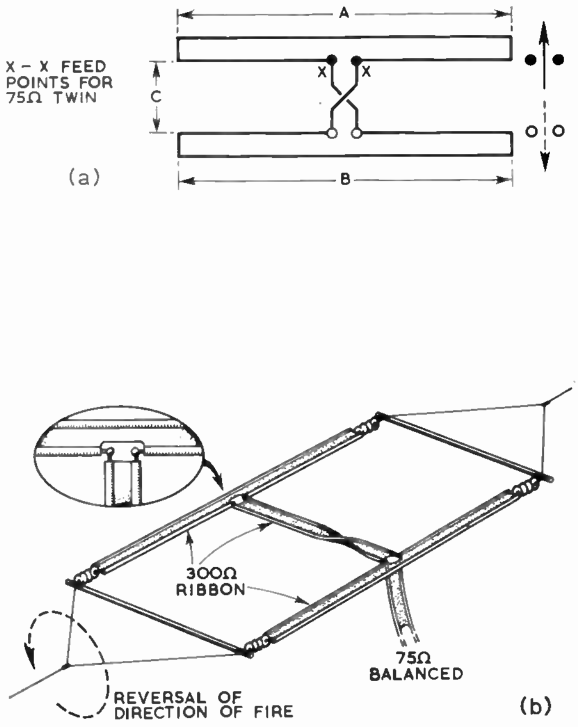


Fig. 13.103. The ZL Special. Reversal of direction can be obtained either by feeding at the opposite dipole or by physical rotation of the aerial about a horizontal axis.

LONG WIRE BEAM AERIALS

The V Beam Aerial

A long wire aerial two wavelengths long has a lobe of maximum radiation at an angle of 36° to the wire. If two such aerials are erected horizontally in the form of a *V* with an included angle of 72°, and if the phasing between them is correct, the two pairs of lobes will add fore and aft along a line in the plane of the aerial and bisecting the *V*. Remaining lobes do not act in this way and so this provides what is essentially a bi-directional beam, although minor lobes will occur away from the main beam.

Fig. 13.104(a) illustrates the principle. If the waves on the wires were visible, they would be seen to flow in the directions of the arrows, to appear in phase from the front of the array; hence an anti-phase or balanced feed line is necessary at the apex.

The directivity and gain of *V*-beams depend on the length of the legs and the correct choice of the angle at the apex of the *V*.

This angle also depends on the length of the legs so that in order to design a *V*-beam the length of the legs should perhaps be the first consideration as this is likely to be the limiting factor in most amateur installations.

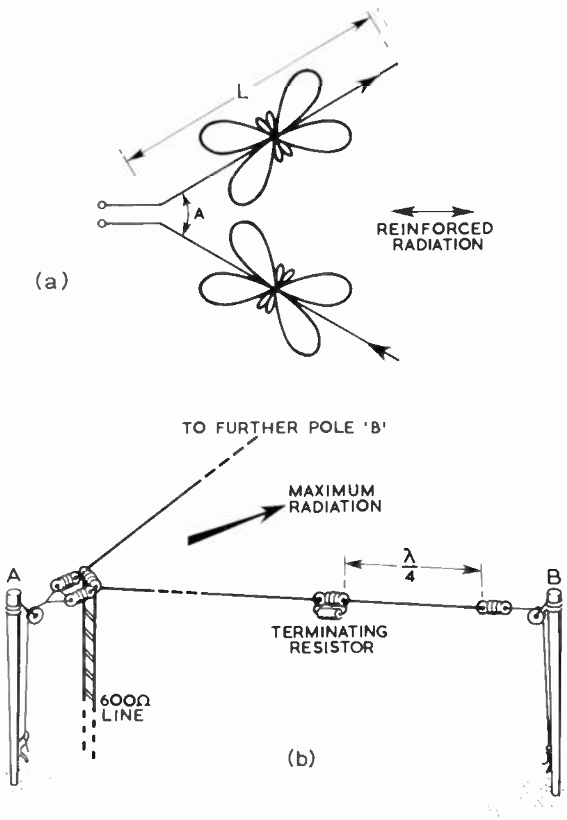


Fig. 13.104. The Vee beam aerial derived from two long wires at an acute angle A. The polar diagrams reinforce along the axis of the aerial and tend to cancel in other directions.

Assuming that the correct angle is used the gain to be expected in the most favourable direction is given in **Table 13.10.**

TABLE 13.10

Leg Length in Wavelengths	Gain	Apex angle
1	3db	108°
2	4·5db	70°
3	5·5db	57°
4	6·5db	47°
5	7·5db	43°
6	8·5db	37°
7	9·3db	34°
8	10·0db	32°

The horizontal polar diagrams of such beams varies with the leg length and is virtually bi-directional along a line bisecting the apex. The shorter lengths show quite a broad lobe which becomes increasingly narrower as the number of wavelengths in each leg is increased.

The general layout of *V*-beams is shown in **Fig.13.104(b).**

Probably the best way of feeding a *V*-beam is by the use of tuned feeders as the impedance is high. If a *V*-beam is designed for a particular frequency it can then be used successfully at higher frequencies. Ideally such a beam should be erected at a height of one wavelength above ground but as in the case of a 14 Mc/s beam this is difficult to achieve. A height of half wavelength will give satisfactory results although the gain will be reduced at low vertical angles of radiation.

This aerial is not one which can be used by the majority of amateurs because of its size but where space is available it is very useful in view of its simplicity. Its other main limitation is the narrowness of the main lobes which limits its all round coverage unless a number can be erected and so orientated so as to cover all the main land masses.

The *V* exhibits some degree of end-fire effect and can be made unidirectional if it is terminated, for example, with the artificial earth described in connection with Fig. 13.87. A suitable value of resistor would be 500 ohms for each leg. The input impedance, in the resonant condition, may rise to 2000 ohms in a short *V* but will be between 800 and 1000 ohms in a longer or terminated aerial and thus 600 ohm feed lines can be used. Alternatively a non-terminated *V* can be driven like the balanced Windom shown in Fig. 13.83(b).

The Rhombic

Early difficulties of terminating a wire high in the air led to the development of the *rhombic* aerial in which a second *V* is added, so that the ends can be brought together. The same lobe addition principle is used but there is an additional complication, because the lobes from the front and rear halves must also add in phase at the required elevation angle. This introduces an extra degree of control in the design so that considerable variation of pattern can be obtained by choosing various apex angles and heights above ground.

The rhombic aerial, a development of the *V*-beam, gives an increased gain but takes up even more room and requires an extra support. There are two forms of the rhombic; the resonant rhombic which exhibits a bi-directional pattern and the terminated rhombic which is non-resonant, unidirectional and has a higher gain.

The layout of the two forms are shown in **Fig. 13.105.**

As can be seen the resonant rhombic can be considered as two acute angle *V*-beams placed end on to each other. The main advantage of the resonant rhombic over a *V* is that it gives somewhat greater gain for the same total wire length, and its directional pattern is less affected by frequency when used over a wide range.

Feeding the rhombic is best done by tuned feeders as for the *V*-beam, as this allows the system to be used on several amateur bands.

The average gain of a rhombic is 3db greater than that of a *V* aerial with the same legs and this is to be expected because the rhombic is twice as large.

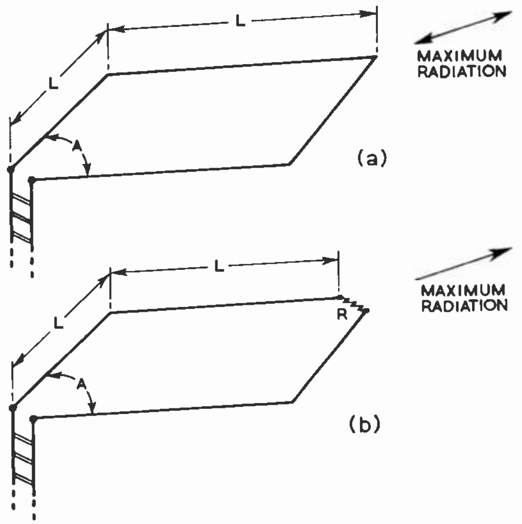


Fig. 13.105. Terminated and unterminated rhombic aerials.

The non-resonant rhombic layout is similar to that of the resonant type, the only real difference being that it is terminated at the far end from the feed point by a non-inductive resistor slightly higher in value than the characteristic impedance of the ordinary or unterminated version. This allows for the energy loss through radiation by the time the far end is reached. An average termination will have a value of approximately 800 ohms. It is very important that the terminating resistor be as near a pure resistance as possible, i.e. without inductance or capacitance. For this reason the use of ordinary wire wound resistors should not be considered as they have far too much inductance and distributed capacitance. The power rating of the terminating resistor should not be less than one third of the power input to the aerial. Generally the rating used is one half for safety margin. Such resistors are not readily available for power outputs of more than say 20 watts but a suitable arrangement can be made up by using a suitable number of small resistors wired in parallel in the same way as is used in construction of dummy loads.

The terminating resistor may be mounted at the extreme ends of the rhombic at the top of the supporting mast or an open line of 800 ohms can be brought down from the top and the terminating resistor be connected across this at near ground level. The impedance at the feed point of a terminated rhombic is 700-800 ohms and a suitable feeder to match this can be made up of 16 s.w.g. wire spaced 12 in. apart. Heavier gauge wire will need somewhat wider spacing, which can be determined from the curves of Fig. 13.12.

The design of both resonant and non-resonant rhombic aerials can be taken from Table 13.10, considering them to be two *V*-beams joined at the free ends. These figures will give a wave angle of the main lobe of approximately 15° in all cases, when the aerial height is one wavelength.

The design of *V* and rhombic aerials is quite flexible and both types will work over a 2 : 1 frequency range or even more, provided the legs are at least 2λ at the lowest frequency. For such wideband use the angle is chosen to suit the length *L* at the mid-range frequency. Generally the beamwidth and wave angle increase at the lower frequency and decrease at the upper frequency, even though the apex angle is not ideal over the whole range.

A suitable height above ground is one wavelength, although aerials with legs as short as 2λ may produce an extra high-angle lobe at this height. A short-leg aerial may be more satisfactory at a height of $\lambda/2$ even though this raises the wave angle somewhat.

A comparison with the figures given for tuned arrays in Fig. 13.95 will show that, for the same total length of radiator, *V* and rhombic aerials give somewhat lower gain, but against this must be set the simplicity of construction, wideband properties and ease of feeding. They do require a great deal of space but where there is no room to erect the rhombic, the gain and low wave angle can, to a large extent, be obtained by placing one *V* over another, spaced $\lambda/2$ apart at the mean working frequency, with the lower leg not less than $\lambda/4$ high. The two feed points in this case can be joined by open wire lines. If the feed line (300 to 400 ohms) is joined to the centre point (as **Fig. 13.106**) multiband operation is maintained; this, however, is practically difficult. The alternative feed limits operation to one band, since the two *V*s must be kept in phase with each other.

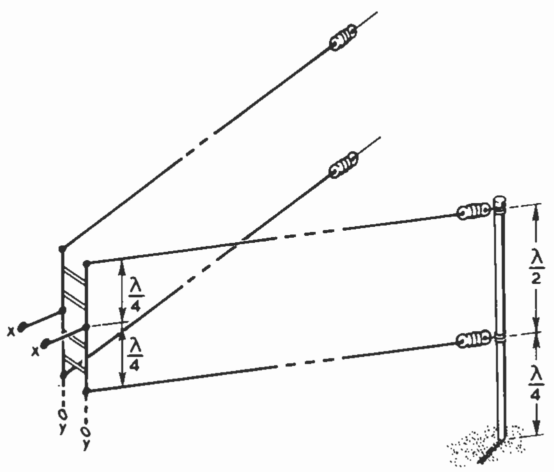


Fig. 13.106. Stacked Vee beam aerials provide lower angles of radiation and increased gain. Feed at points X-X for multiband or Y-Y for single band operation.

PARASITIC ARRAYS

The broadside arrays of the previous section radiate equally fore and aft. In order to make them unidirectional, reflectors are added behind each dipole of the array. In theory, driven reflectors can raise the forward gain of an array by 3db, or that of a single dipole by 4db and at the same time produce a high front-to-back signal ratio. In practice it is very difficult to adjust arrays with driven reflectors so that the currents are equal in amplitude and 90° in phase as required by theory, because the effect of mutual impedance is to produce unequal radiation resistances in aerial and reflector.

It has been found much simpler and just as effective to use a parasitic *reflector*, which picks up some of the power of the driven dipole and re-radiates it in the required phase. At a quarter-wave spacing the coupling between aerials is however too low, and better performance is obtained by reducing the spacing and at the same time increasing the length of the reflector to make it inductive, i.e. lagging in phase: this adjusts the phase of the induced current to suit the reduced time of travel between the closer elements. The resulting array has more gain than a driven pair: at the optimum spacing of $0{\cdot}15\lambda$ the gain is 5db, and the front-to-back ratio about 12db, but the input resistance falls to 25 ohms.

The parasitic element may also be cut shorter than resonant length, so that it is capacitive or leading in phase. In this case it draws the wave forward, and at the same time cancels the radiation in the opposite direction. Such a parasitic element, called a *director*, gives a little more gain than a reflector, with a much better front-to-back ratio, especially at close spacings. It is therefore more commonly used. The optimum spacing is $0{\cdot}1\lambda$ where the maximum gain is just over 5db, with the director tuned as described later. The best front-to-back ratio, 15 to 17db, occurs with a somewhat shorter director, and the gain is then slightly less. If the spacing is reduced to $0{\cdot}05\lambda$, a front-to-back ratio of over 25db can be obtained with the same gain but the aerial impedance is then very low, and fluctuates violently with movement of the conductors in the wind so that such close spacing is avoided in practice.

Adjustments in these aerials are usually made for best back-to-front ratio, because in these aerials the gain does not change very rapidly as the parasite length is adjusted. and because of the advantage of interference reduction, particularly in reception. This may be done by comparing received signals from local stations, or if the aerial can be rotated, by transmitting to or receiving from a fixed station.

A field strength indicator may help in the adjustment of these arrays, but does not give the best result, because the local field of any aerial is very different from the distant field, especially near the ground. If a field strength meter is used it should be connected to a short dipole, set up parallel to the array, at the same height and as far away as possible. An important point is that the impedance of the driven aerial changes rapidly when the parasitic element is tuned, so that the power drawn from the transmitter also varies with this tuning. Thus, unless a power-flow monitor, not an ammeter but a reflectometer is used the optimum aerial gain adjustment will not be obtained.

The adjustment for best front-to-back ratio will give a little less than maximum gain, but once the advantage of a good ratio has been experienced in reception, it will be realized that this slight loss is well justified. It will be found that the best reflector for this aim is somewhat longer, or the best director somewhat shorter than the length giving maximum gain.

Two Element Arrays

Two-element aerials of these types can be constructed for 14 Mc/s and higher frequencies, in fixed or rotating form, and as suspended wire aerials for 7 Mc/s. Rotating arrays are usually made of rigid metal tubing. When the array is fixed, for example supported between spreaders, good insulators should be used at the ends of the wires as the voltages there are relatively high, and serious loss can therefore easily occur. In the case of fixed arrays, it is convenient to be able to reverse direction, and this may be arranged by changing the parasitic element from reflector to director, using a *U*-section or stub in the centre which can be shorted out to change from the length of reflector to that of director. This stub can, in fact, be extended by one half or one full wavelength using 600 or 300 ohm twin line, so that the position of the shorting link can be changed at ground level without lowering the aerial.

TABLE 13.11
Properties of Two-element Parasitic Arrays

Spacing	Reflector				Director			
s/λ	X	Gain	F/B	X	X	Gain	F/B	R_d
	ohms	db	db	ohms	ohms	db	db	ohms
$0{\cdot}05\lambda$	—	—	—	—	−10	4·4	20	13
					+10	5·6	6·6	4
$0{\cdot}1\lambda$	+30	5	6	16	−10	5·2	7·8	14
	+50	4	10	26	−40	4·2	17	33
$0{\cdot}15\lambda$	+20	5	6	35	—	—	—	—
	+50	4	10	50	—	—	—	—

Theoretical properties of two-element parasitic arrays, which are largely realized in practice. The tuning of the parasitic element is given in terms of reactance X (see Fig. 13.107). Gain and front/back ratios are given in db, and feed resistance R_d in ohms. Of the two sets of figures given for each spacing, one is for maximum gain and the other maximum front/back ratio.

From **Table 13.11**, it will be seen that $0{\cdot}1\lambda$ is a good value of spacing for a combined reflector/director system; this is 7 ft. for 14 Mc/s. At this spacing the feed point impedance at the centre of the driven aerial is between 15 and 30 ohms, and the aerial is fairly sharply tuned, but can still effectively cover the 14 Mc/s band. To match this impedance to available transmission lines, a transformer is needed, and this should be introduced as near the aerial as possible in order to preserve bandwidth. The best way to do this is in the aerial itself, by using a folded radiator. A two-fold type, such as the 300 ohm ribbon folded dipole, will bring the feed impedance to a value suitable for 70–80 ohm twin feeder. When the array is rotatable and self-supporting, folding is not really practicable and open wire twin line is not suitable because it changes its characteristics as it wraps itself round the supporting mast. However, the transformers given later for three-element arrays may be used, the adjustments being practically the same.

Tuning the Parasitic Elements

Formulae for element lengths are frequently quoted, but are not useful unless the diameter of the elements is also specified. What controls the performance is the reactance of the parasitic element; that is, the amount by which the element is detuned from resonance, and this, in terms of length, depends on the length/diameter ratio. **Fig. 13.107** shows reactance as a function of percentage change of length for various relative conductor diameters. The crossover point of the curves occurs when the radiator is a true free-space half wavelength long calculated as $492/f$ (ft. Mc/s). At this length the conductors all have a reactance of about + 40 ohms, and must be *end corrected* by the percentage which brings the reactance to zero to obtain resonance. For example, the correction for a relatively " fat " aerial, with L/d = 100, is 5·5 per cent so that for 28·0 Mc/s the full half-wave would be 17 ft. 6 in. and the end correction 1 ft. making the resonant length 16 ft. 6 in.

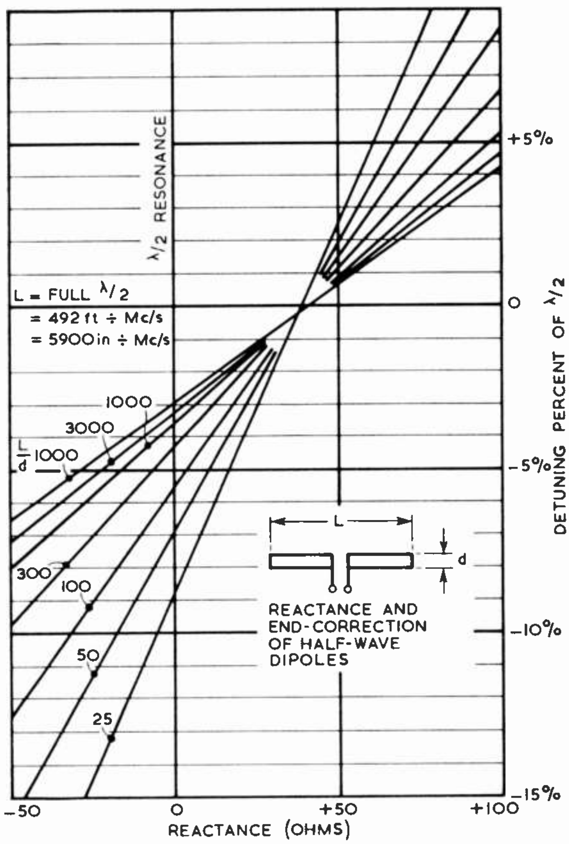


Fig. 13.107. Tuning and reactance chart for half-wave dipoles, as a function of the length-diameter ratio. A radiator exactly half-wave long is " over tuned " by 42 ohms, and " end correction," given as a percentage of the length, is necessary to bring it to resonance (zero reactance). The chart is useful for the construction of parasitic arrays and v.h.f. dipoles. Each 1 per cent of length corresponds to 5 units in the factor 492/f or 60 in 5900/f (f is in Mc/s).

The best reflector has a positive reactance of about 40 ohms, which brings its length to almost a true half-wave, independently of diameter. For 0·1λ spacing, best director action occurs at −10 to −40 ohms reactance with a length some 5 to 10 per cent shorter than the full half wavelength. For example, at 21·0 Mc/s, $\lambda/2 = 492/f = 23{\cdot}4$ ft. A 1 in. diameter half-wave radiator would have a L/d ratio of about 300, and an end-correction of −4·4 per cent, giving a resonant length of 22 ft. 4 in., whilst a director of −30 ohms reactance would be 7.4 per cent short of true half-wave, or 21 ft. 7 in.

In practice it is always advisable to find the best adjustment experimentally, using the above theory as a guide, because the environment of the aerial has some effect. The parasitic elements can be tuned first, independently of the driven element, which may then be adjusted for optimum tuning and impedance match, as described later.

The following figures will be found satisfactory for *two element* beams in most locations where the spacing using a director is 0·1 wavelength or where the reflector spacing is 0·15 wavelength.

Reflector length in feet $\dfrac{500}{f(\text{Mc/s})}$

Driven element in feet $\dfrac{475}{f(\text{Mc/s})}$

Director length in feet $\dfrac{455}{f(\text{Mc/s})}$

From this it will be seen that the driven element is a little longer than the normal half wavelength. The director is 5 per cent shorter than the driven element and the reflector is 5 per cent longer than the driven element.

Three- and Four-element Parasitic Arrays

Improved performance can be obtained by a combination of reflector and one or more directors, the usual arrangement having all the elements in one plane. Such aerials are called *Yagi* arrays, after their inventor, and give as much gain as can practicably be obtained in the space they occupy. They have a relatively compact beam of radiation in the direction of the directors. There is no advantage in using more than one reflector in an in-line array, though many directors may be used with advantage. For h.f. work one, or at the most, two, directors are the limit because of the size of the array, but in the v.h.f. bands longer arrays with more than three elements may be built, and are described in Chapter 14.

The characteristics of two-element arrays can be calculated quite well, but three or more elements lead to such complexity that their analysis requires a computer. A vast amount of experiment by amateurs and others has, however, resulted in comprehensive knowledge, and an infinite variety of prescriptions for element lengths and spacings.

The vertical and horizontal radiation patterns of a three element Yagi array are given in **Fig. 13.108.**

The maximum theoretical gain obtainable is 7·5db for three elements and nearly 9db for four elements. These patterns have their half-power points at a beamwidth of about 55° in the *E*-plane (the plane of the elements) and 65–75° in the *H*-plane (at right angles to the elements). These high gains are, unfortunately, associated with a feed-point resistance of only a few ohms, so that in practice the loss due to conductor resistance and environment may be appreciable in proportion. In addition, the tuning is very sharp, so the array will only work usefully over a small part of, say, the 14 Mc/s band, whilst the line matching for a low v.s.w.r. is very difficult.

The power loss due to conductor resistance, or high feeder v.s.w.r., is such that these high gains are never realized in practice. It is better, therefore, to design the aerial for less gain, but better general performance. How this can be done is best seen with the aid of **Fig. 13.109**. This is a " contour " map of gain and impedance as a function of parasitic tuning,

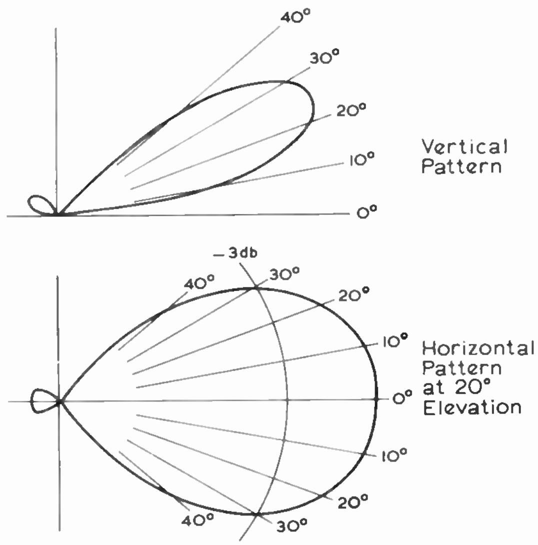


Fig. 13.108. Vertical and horizontal radiation patterns of a three-element Yagi array spaced 0·15 + 0·15λ, and erected at a height of one half wavelength. These patterns, which were obtained from a scale model, do not vary much with other spacings of the elements. The vertical pattern is quite narrow compared to the free-space pattern and this is due to choice of correct height. At a height of λ the pattern is also good but at 3λ/4 an extra lobe appears in the vertical pattern. Note that the front/back ratio along the ground may be very different from that at 20° elevation.

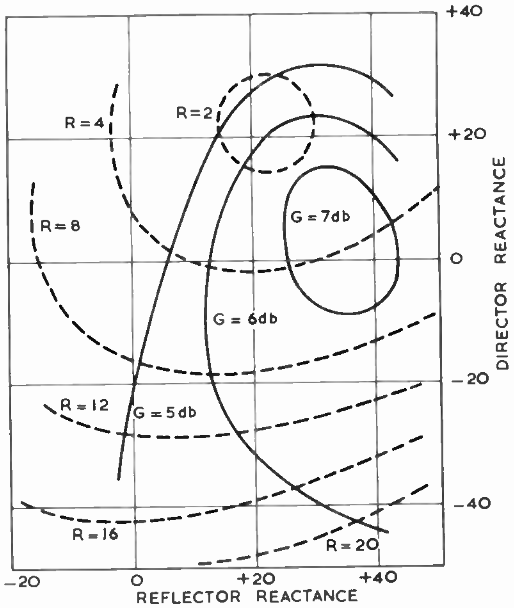


Fig. 13.109. Contour chart showing contours of gain (solid lines) and input resistance (broken lines) as a function of the tuning of the parasitic elements. This chart is for a spacing of 0·15λ between elements, but is also typical of arrays using 0·2 + 0·1 spacings.

for a Yagi with element spacings of 0·15λ. The lower right-hand corner of the chart is clearly the region to aim for where the gain is still over 6db (or 7db for four elements) whilst the feed resistance approaches 20 ohms. This region of the chart corresponds to rather long reflectors (+ 40 to + 50 ohms reactance) and rather shorter than optimum directors (− 30 to − 40 ohms) and also is found to be the region of best front-to-back ratio.

Although the chart is for aerials with both spacings equal to 0·15λ, the resistance and bandwidth are both somewhat improved, with the gain remaining over 6db, if the reflector spacing is raised to 0·2λ and the director spacing reduced to 0·1λ. Such an array would operate satisfactorily over at least half the 28–29·7 Mc/s band or the whole of any other band on which it could be used.

The overall array length of 0·3λ is practicable on all bands from 14 Mc/s upwards, though the 20 ft. boom required for a 14 Mc/s array is rather heavy for a rotary array, and there is a temptation to shorten it. Spacings should not be reduced below 0·1λ for both reflector and director as the tuning will again become too critical and the transmitter load unstable as the elements or the feeder move in the wind.

The optimum tunings do not vary appreciably for the different spacings, and it is therefore possible to construct a practical design chart (**Fig. 13.110**). An array, made with its element lengths falling in the shaded regions of the diagram, will give good performance without further adjustment. The length of the director may be decided in advance. The reflector may then be adjusted experimentally to improve the front-to-back ratio. It will be seen that the radiator is somewhat longer than a normal dipole; this is because the parasitic elements have a detuning effect on it. The addition of a second director also spaced 0·1 to 0·15λ will not

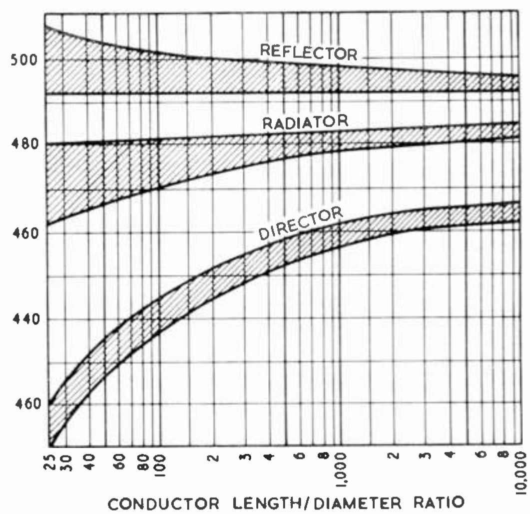


Fig. 13.110. Design chart for Yagi arrays, giving element lengths as a function of conductor length-to-diameter ratio. The tuning factor L is divided by the frequency in Mc/s to give the lengths in feet. These curves are for arrays of overall length 0·3λ, with reflector reactance + 40 to + 60 ohms and director −30 to −40 ohms, and give arrays of input impedance between 15 and 20 ohms. Element lengths which fall within the shaded areas will give an array which can be used without further adjustment, though the front-back ratio may be improved by adjusting the reflector.

materially affect the above recommendations, and can be expected to increase the gain by 1db.

The adjustment of the array can be carried out as indicated for two element arrays, using the front-to-back ratio as the criterion. For this purpose the elements can be made with sliding tube extensions or, alternatively, a short variable stub can be fitted at the centre. The tuning-up can be carried out without reference to impedance matching, which is always the last operation. Adjustments should be made with the array as high and as clear as possible, since tuning alters near the ground, and performance should be checked with the array in its final position. It should be emphasized that attempts to adjust for maximum gain lead inevitably into the top right-hand corner of Fig. 13.109 where bandwidth is low and matching difficult.

Impedance Matching of Yagis

The 15–20 ohm aerial impedance resulting from the foregoing recommendations can conveniently be matched to 80 ohm line by a single fold of the driven element. Since the impedance rises on either side of the optimum frequency, it is better not to match at this frequency, but to bring the impedance up to say, 60 ohms (v.s.w.r. = 1·3 with maximum voltage at the aerial). The aerial will then hold within a v.s.w.r. of 2 over a somewhat wider band.

It is possible to multiple-fold the driven element using, for example, a rigid central tube surrounded by a cage of wires joined to radial struts a few inches long at the free ends of the tubes. In this way the radiation resistance can be raised to 300 or 600 ohms. When there are several thin wires surrounding a tube, the step-up ratio is approximately the square of the number of conductors added, so that four wires would bring an 18 ohm aerial up to 300 ohms. Most of the examples which have been published are for maximum-gain Yagis, with a basic feed impedance of 6 to 8 ohms, and use up to nine wires. It is claimed that this multi-folding restores the bandwidth, but this is doubtful since it is the parasitic elements which are responsible for the high Q.

It will be appreciated that two adjustments are necessary for matching to the feed line, namely, tuning and transformer ratio. Most of the known matching arrangements provide only one of these, and the operation of adjusting the length of the radiator at the same time as the transformer is tedious and difficult. The *T* and *Gamma* transformers of Fig. 13.40 are suitable for this purpose.

The Cubical Quad

This aerial is very popular with DX operators on the 14, 21 and 28 Mc/s bands and was originally developed by W9LZX in 1942.

It consists basically of a driven element in the shape of a square, each side of which is a quarter wavelength long. Behind this driven element is placed a closed square of wire to act as a parasitic reflector. The layout is shown in **Fig. 13.111** and in this configuration it is horizontally polarized. Feeding in the middle of one of the vertical sides would change the polarization to the other sense in the main direction.

The quad can be considered as two closely stacked two-element parasitic aerials which are end-fed. The currents in the top and bottom wires are in phase and add to give broadside horizontally polarised radiation. Those in the vertical sides are in anti-phase and cancel in the broadside direction, giving rise only to a small amount of vertically polarized radiation off the side.

The radiation resistance of the cubical quad varies with its height above ground and also the spacing between the driven element and the parasitic reflector.

In order to achieve a radiation resistance of 75 ohms the spacing between the two elements should be 0·15 wavelength and the height half or one wavelength above ground. It is convenient therefore that the optimum spacing between the elements for maximum gain approaches 0·15 wavelength. The gain obtained in practice from a cubical quad depends upon its use. When measured against a reference dipole at fairly close range the gain is 5·7db. However, because the main lobe of radiation in the vertical plane is quite low the signals received at a DX station will show an apparent gain over a dipole a good deal greater than the free-space gain of 5·7db.

The actual angles of radiation in the vertical plane depend of course on the height of the aerial above ground. At a half wavelength high the vertical angle of the main lobe will be approximately 25° and at a height of one wavelength as low as 12° which means that it is a very satisfactory aerial for DX working. Because of its comparatively small size it is not difficult to make up in the form of a rotating beam.

There are certain points of importance to note in the electrical construction of a cubical quad. Because the elements of a quad consist of a continuous loop there is no "end" effect and the usual formulae are not accurate. In fact the overall length is slightly longer than it would be for an aerial in free space. For the driven element, each side of the square should be $248/f$(Mc/s) and fed with 75 ohm flat twin, or coaxial with a balun. If an unbalanced feed by coaxial cable is used there will be a distortion of the lobes caused by the unbalance.

The size of the parasitic reflector element is the same as that of the driven element but it is made electrically longer than the driven element by means of a short circuited stub which can be adjusted for maximum gain or front to back ratio. It will be found quite easy to adjust the stub for a front-to-back ratio better than 20db.

Some amateurs have avoided using stubs to tune the reflector by making the overall length of the reflector 5 per cent more than the driven element. While this method does not permit adjustment for optimum results it has often worked quite well in practice when a single quad has been mounted.

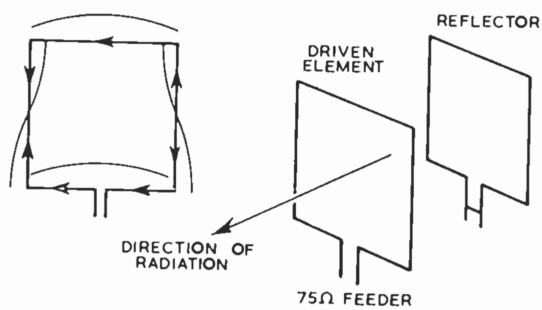


Fig. 13.111. The basic cubical quad showing the current distribution and direction around the driven element. Radiation from the vertical sides tends to cancel as the currents are in opposition.

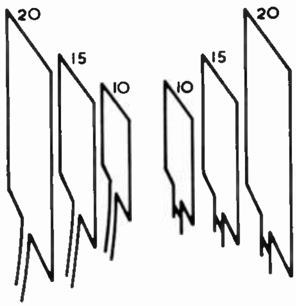


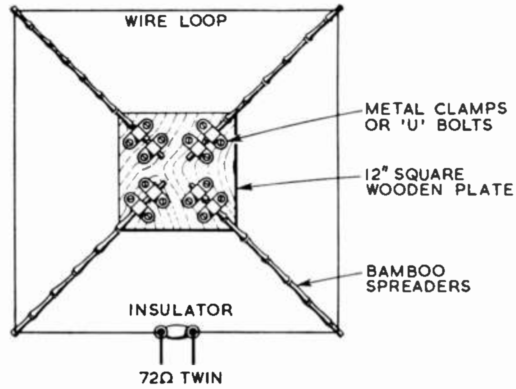
Fig. 13.112. A three-band nest of cubical quads maintaining correct spacing for each band.

The following table will give dimensions suitable for the DX bands.

Band	Side of Quad	Spacing	Approximate Stub Length	Side of Quad Reflector without Stub
14 Mc/s	17 ft. 6 in.	8 ft. 5 in.	36 in.	18 ft. 4 in.
21 Mc/s	11 ft. 8 in.	5 ft. 7 in.	20 in.	12 ft. 4 in.
28 Mc/s	8 ft. 8 in.	4 ft. 2 in.	16 in.	9 ft. 1 in.

The foregoing description is for a single band quad but a two or three band quad can be made up by mounting the elements concentrically on a bamboo spider. The dimensions will be the same as in the above table if three quads are mounted as shown in **Fig. 13.112.** If, however, the driven elements are mounted in the same plane and the reflectors similarly, there will be a change in the gain and the radiation resistance due to the increased spacing used on the higher frequencies. If the spacing is optimum for the 14 Mc/s then the gain will be lowered and the radiation resistance will be increased for the 21 and 28 Mc/s quads. At 28 Mc/s the gain will drop to 5·0db and the radiation resistance will increase to 140 ohms.

A suggested constructional layout for a single band quad is shown in **Fig. 13.113.** If a three band quad is to be constructed, the length of the horizontal boom should be reduced, and the bamboo arms angled outwards, so that the correct element-to-element spacing is maintained for each band.



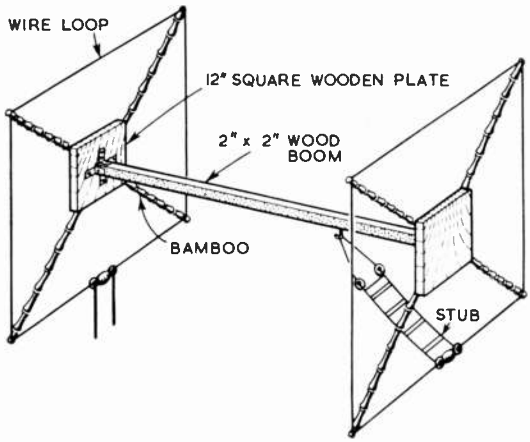


Fig. 13.113. A suggested method of construction for a lightweight single band cubical quad.

The G4ZU "Bird-cage" Aerial

Popular as the quad aerial has become, it presents some practical difficulties in construction. The fact that the wire elements are supported at points of relatively high r.f. potential can lead to insulation losses in wet weather. A multi-band quad offers considerable wind resistance, and the use of substantial bamboo cross-arms results in a prominent structure which may not be acceptable in all locations.

An aerial developed on parallel principles by G4ZU is claimed to overcome these disadvantages. While two quarter-wave loops are again used in an electrically similar manner, the upper and lower arms of each are bent into a "*V*" as shown in **Fig. 13.114**, bringing the low-potential feed and stub points near to the central supporting mast which may now be of metal as in the case of many parasitic arrays having earthed elements. These upper and lower arms which are individually only 0·125 wavelength long, are made from

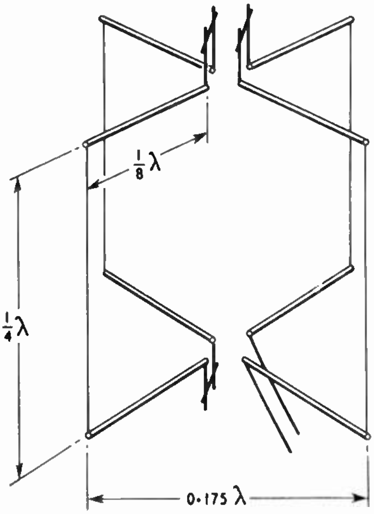


Fig. 13.114. The G4ZU Bird Cage modification for the cubical quad aerial.

dural tubing, and themselves form the supporting structure, no insulated beams being required. The vertical elements joining their tips are of wire. It still remains necessary to insulate the radial arms from the central dural mast, but as these are points of *low* r. f. potential they can be supported by blocks of insulating materials such as Tufnol, Bakelite or wood impregnated with wax, and leakage due to weather will not give rise to serious losses.

For purposes of adjustment, stubs will be required at the centre of each loop and these can be conveniently supported along or even inside the mast. The feed point impedance is in the order of 40-50 ohms, but higher impedances up to 600 ohms can be obtained by closing the lower element, and tapping the feed points out symmetrically along the driven element in a form of " *Y* " match. It has been pointed out also that the use of an upper stub of 300 ohm ribbon, $\lambda/2$ long for the working frequency of, say, 14 Mc/s, and accommodated inside the mast, will enable the aerial to operate on the next lower band of 7 Mc/s, when the stub will become $\lambda/4$, and the whole driven loop a half-wave element.

The use of tubing for the centre portion of each equivalent dipole gives the G4ZU a broader band-width than is typical of wire beams, whilst the " V " formation results in somewhat higher gain than the quad aerial, and a marked reduction in side-lobes. Typically 9db is claimed, with the advantages of the front-to-back ratio, and very low angle radiation when erected a half wavelength above ground. For single-band operation a simplified construction has been proposed in which the height of the array is increased to somewhat more than a quarter-wave in order to resonate outside the low end of the desired band without the use of stubs. Variable capacitors are then inserted into each loop near to the feed-point, and adjusted to reduce the electrical length of each loop to resonance; that in the reflector loop being set for maximum gain at the working frequency, and that in the driven loop for minimum s.w.r. on the feeder. The setting up of this form of the G4ZU is stated to be exceedingly simple, and can often be carried out in the ideal manner after erection.

VERTICAL H.F. AERIALS

In recent years vertical aerials have gained greatly in popularity, since their attractive properties of omnidirectional low-angle radiation provide an excellent solution to DX working in locations where space is not available for the larger beam structures, or in cases where the latter are ruled out by aesthetic considerations, or by local Council regulations. The uses of such aerials for multiband and for mobile applications is considered in a later section.

In its simplest form the vertical aerial will be approximately a quarter wavelength long, and used against earth as shown in Fig. 13.115(a). A 50 ohm cable may be employed with the sheathing joined to a good earth, but there will always be a significant loss of power in the earth resistance, and matching to the cable will not be exact since the feedpoint impedance will in general be lower than 50 ohms. Such aerials can prove effective in wet locations having exceptional earth conductivity, in coastal regions, or in maritime-mobile applications where the hull of a vessel can provide low " earth " resistance to the water. They have also been used with very fair results when the aerial can be mounted well in the clear above a tall building, or perhaps protruding through

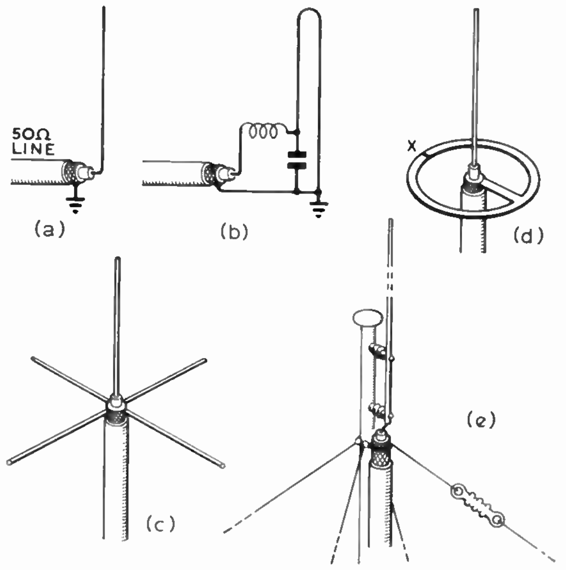


Fig. 13.115. Quarter wave vertical aerials. (a) Single wire using earth connection. (b) Folded wire with transformer (see text dealing with impedance matching). (c) Ground plane using artificial earth. (d) Compact ground plane. (e) Guy wires used as a ground plane. The cable size is exaggerated for clarity.

the tiles from inside the loft-space of a house, and earthed to the network of domestic water pipes which in many modern systems are of copper. A copper rising-main can form an excellent r.f. earth, but beware of plastic piping systems, and do not omit a high-voltage series capacitor such as a 0·01 μF mica or 1000 pF ceramic component in series with the earth lead to guard against any possible short circuits to the supply mains at the equipment end.

A method which may be used to reduce earth losses is that of folding the radiator as shown in Fig. 13.115(b) which has the effect of raising the radiation resistance and feed-point impedance by a factor of four, thus reducing the proportional earth resistance. The feed impedance of between 150 to 200 ohms is difficult to match directly, and a transformer or its equivalent may be needed as described later. Such aerials are difficult to construct as self-supporting structures, but are excellent if a wooden mast is available to which they can be attached. A convenient feature is that the aerial forms a closed loop across the feeder cable, so that continuity can be checked from the station by the use of a d.c. resistance meter.

The Ground Plane

Problems of earth resistance can be virtually eliminated by erecting the vertical over a perfectly-conducting surface, such as a large sheet of copper, which gives rise to the term ground plane aerials. The radiation pattern is also greatly improved under such conditions. In practice however, it is usual to simulate the ideal ground-plane by a system of rods or wires having similar properties, as this is very much simpler to carry out. The usual approach employs a number of radial wires approximately a quarter wavelength long. At high frequencies four or six of these may be self supporting rods at the top of a vertical mast, and bonded to the cable sheath on the lines shown in **Fig. 13.115(c).** Another construction favoured for physical reasons is the ring groundplane illustrated in **Fig. 13.115 (d).** At the lower amateur frequencies where such self-supporting construction is not practicable, the radials may consist of wires buried a few inches below ground, up to 16 being desirable for a really efficient system (see also page 13.31). The vertical ground-plane has an excellent radiation characteristic even at ground level, and there is little advantage in raising it except when surrounding objects such as buildings introduce screening losses. It may then be supported by a short insulating mast; the radials can form guy-wires as indicated in **Fig. 13.115(e).**

Matching the impedance of a ground-plane to the widely used 50 or 75 ohm cables presents a problem which is considered more fully in the section on Impedance Matching. Whereas the feed-point impedance of an earthed quarter-wave aerial may lie between 35 and 50 ohms, that of a true ground-plane will be 20 ohms or less, and feeder cables of this impedance are not available. A solution to this problem can be found by tilting the radials downwards as illustrated in Fig. 13.115, since this raises the feed-point impedance towards that of the dipole, which the aerials would become if the radials were considered to be " tilted " directly downwards at 90 degrees. Four guy-wires may be used, fitted with insulators at a quarter wavelength from the base of the radiator, and spaced as nearly as possible at 90° from each other. If the angle of tilt is 45° the system will be a good match to 50 ohm cable. Standing waves along the feeder can be measured as described later, the angle of tilt being adjusted slightly to arrive at a minimum s.w.r. It is possible by the use of still steeper angles to match into 75 ohm cables, but such an aerial becomes very nearly a vertical dipole and demands considerable height, since the radials will extend downwards for nearly a quarter wavelength. In addition, approximately another quarter wavelength clearance above ground is desirable if the aerial is to exhibit good low angle radiation. If 75 ohm cable is to be used, it is better practice to employ a horizontal ground-plane and fold the radiator as in Fig. 13.115(b), or to use a form of matching network.

In some installations it may be found that the outside of the cable is " live " after it has left the ground-plane. When this occurs, it may be necessary to coil up a few feet of the cable to act as a form of choke effective on the outer surface only. Where there is insufficient space for radials the ground-plane can be folded as in Fig. 13.115(d). To understand how this works, imagine that a single radial only, has been provided, and that after a certain distance it is split and the ends folded into a circle. The two free ends are at the same potential and may therefore be joined together. The length of the radial part is 0·07 wavelength and the circumference 0·43 wavelength. This aerial is as effective as the previous one, but tuning is somewhat more critical and the input impedance low.

The Extended Ground-Plane

The aerials so far described are assumed to be very close to a quarter wavelength long, and as has been stated, the feed-point impedance is invariably low. As the aerial is lengthened however, the feed-point impedance increases, and can be expressed in the form $R + jX$ where R is a pure resistance and X an inductive reactance which would represent a mismatch. Increasing the aerial length from a quarter to approximately one-third wavelength raises the resistive component to 75 ohms, whilst the reactance X rises at the same time to between 300 and 400 ohms. It is thus possible to obtain a good match directly into any feeder of impedance between 50 and 75 ohms by lengthening the aerial by a suitable amount, whilst at the same time tuning out the additional inductive reactance thus introduced by means of an adjustable series-capacitor. The adjustment will, of course, only be correct for one frequency band, where it provides a simple and effective alternative to the more sophisticated methods of impedance matching set out in the following section.

Arriving at the best length for the extended ground-plane aerial cannot be done reliably by calculation, since it is related to the form of earthing, the number and length of the radials, etc, in actual use, and is dependent to a lesser extent upon the height of the system above ground, and the effect of surrounding objects. 0·375 wavelength is frequently regarded as a good compromise or starting point. However, if a standing-wave meter or reflectometer of the type described in Chapter 19 is available the optimum length can be simply arrived at by trial and error. The standing-wave ratio is measured with the capacitor adjusted to give minimum reading. A small adjustment to aerial length is then made, and the ratio checked to see if it is better or worse; preferably the aerial should initially be too long, and the reduction should lead to a reduced s.w.r. It is then only necessary to reduce further the length until by adjusting the capacitor the s.w.r. passes through, or near to zero. If sloping or movable radials are in use, the final adjustment can be made even more easily by repositioning these, or even shortening one of them slightly. Increasing the radial angle from the horizontal, or even reducing the length of one radial wire, is equivalent to slightly lengthening the aerial itself.

The Loaded Vertical

Vertical or ground-plane aerials present little problem at the higher frequency bands where a self-supporting radiator up to 0·375 wavelength long, or a wire supported by a mast can be erected. Not infrequently a metal mast or lattice tower insulated from earth at its base can be used effectively, as is common practice at broadcast frequencies. To reap the advantages of vertical aerials at 7 Mc/s or lower frequencies however, it may be necessary to load the aerial to resonance by inductance, on the lines described in the preceding section for the 1·8 Mc/s band, and this is common practice in mobile installations where the length of whip-aerial is strictly limited by practical considerations.

There is no general principle in relation to the inductive loading of aerials to permit their use at lower frequencies, since the added inductance can be of any value from a few turns used mainly to trim the aerial length to resonance, to a coil in which the greater part of the total inductance is concentrated. Two broad approaches to the problem are recognized however, generally referred to as base loading, and centre or top loading. Both are represented in Fig. 13.72(b). The guiding principle is that for best performance from a vertical aerial maximum current is essential in the vertical portion, particularly towards the top where screening is likely to be least serious. In a short aerial however, maximum current will always occur at the feed-point, and will decrease progressively towards the free end. Base loading can be thought of as coiling-up the bottom part of the aerial in which maximum current is flowing, thus reducing radiation and encouraging coil losses, but the effects are to some extent offset by the fact that less inductance is needed to load an aerial to a given frequency at the base than at any other point along its length.

Base loading is clearly the most convenient since the loading coils are accessible for changing bands, and can also be used as a coupling transformer on the lines shown in Fig. 13.72. For the reasons mentioned, however, base loading is generally confined to adapting an aerial for use on the next lowest frequency band, because the inductance will not then be large enough to upset the current distribution seriously, or to introduce excessive losses. Thus a vertical aerial designed for quarter wave or 0·375 wavelength on 28 or 21 Mc/s be large enough to upset the current distribution seriously, or to introduce excessive losses. Thus a vertical aerial designed for quarter wave or ⅜ wavelength on 28 or 21 Mc/s can be simply adapted for use on 14 Mc/s; and more commonly a 14 Mc/s vertical may be loaded for use on the 7 Mc/s band. A 32 ft. vertical resonant at 7 Mc/s is not unduly difficult to erect, and may be base-loaded effectively for 3·5 Mc/s. The exact size of a base-loading inductance coil cannot be precisely stated, since in effect it is tuned by the self-capacitance of the aerial above it to earth, and this capacitance is dependent upon local conditions. As a guide however, it can be stated that a coil containing an equal length of wire to the aerial itself will prove rather too large for resonance into the next lowest band, and the aerial can be tapped down such a coil until resonance is achieved, and the unused portion then discarded. A roller-coaster coil is popular for base loading purposes. Whatever type be used, the coil should be of low-loss construction, and protected against the weather by a suitable waterproof housing.

The marked increase in mobile activity in recent years has led to the introduction of numerous loaded whip aerials having demountable coils specially suitable to this work. Such coils are ideal as a basis for more permanent loaded vertical aerials for the h.f. bands. Mobile aerials are in many instances centre-loaded, which implies that the coil will be located part-way up the mobile system, and may be designed for mounting upon a section of 1 in. diameter aluminium tubing. The section of whip above this coil determines its resonance, and often takes the form of a 4 ft. to 5 ft. section of ⅜ in. diameter tubing. Coils are available in interchangeable form for the various amateur bands.

Since in a mobile system of this kind maximum current is carried in the section below the loading-coil, there will be a very large increase in radiation if this section is greatly lengthened. Making this change however has only a slight effect upon the resonant frequency, which tends to rise as the capacitance of the top section to earth is reduced, but is simultaneously lowered by the added inductance of the lengthened lower section. Excellent loaded verticals can therefore be constructed by mounting a standard coil and top whip section for the band required upon a mast of 1 in. diameter tubing, which may itself be of resonant length for a higher amateur band. The mast must be insulated at its base, either by the aid of a mounting insulator, or a short section of wood or bakelite rod or tube, and will be fed as already described. There is no objection to a small trimming-inductance at the base, since the frequency in the loaded-mode is likely to be a little high; should it be otherwise, the top section may be shortened with caution, as this length is very critical. An aerial of this kind will require to be guyed at a point just below the loading-coil, and at other points according to the actual construction and overall height chosen.

The aerial described is effectively top-loaded for the lowest frequency band, and will give low-angle radiation with good efficiency. Bearing in mind that a vertical radiator need not necessarily be quarter wave, but that half wave aerials equivalent to dipoles can be used with equal effectiveness, it is possible to excite such a system on several bands by altering the feeder termination. Where space is limited, the feeder line reasonably short, and the base of the aerial accessible for adjustment, a vertical compromise system on these lines can be very useful for all-round DX operating.

As a practical example, the vertical aerial may consist of from 32 ft. to 40 ft. of 1 in. diameter tubing, terminating at the base in a coil similar to that described in Fig. 13.72(b) with provision for tapping the feeder up or down the inductance, but without the series capacitor shown in that illustration. At the top this 1 in. tubing carries a loading-coil of the type used in 1·9 Mc/s mobile installations, and carrying above it the usual whip section which will resonate towards the h.f. end of the 1·8 to 2·0 Mc/s band. To operate on this band, the feeder is taken very nearly to the top of the base loading-coil, the remaining few turns of which resonate the system to the required part of the band, whilst the larger portion of the coil forms a high-inductance shunt across the feeder termination.

For operation as a vertical half wave on the 3·5 Mc/s band, the feed-point is moved down the coil until the electrical length of the whole aerial becomes three-quarter wavelength at the working frequency, when the system will be voltage-fed at the bottom of the vertical section. The inductance of the base loading-coil must be sufficient to assure this condition. Where this is not so it will be of assistance to add a 100 pF variable tank capacitor across the whole of this coil, and to tune it to resonance as indicated by maximum brightness from a neon lamp placed adjacent to the top of the loading coil. Maximum current will then occur at the top of the vertical section, immediately below the top-loading coil, and in a favourable position for effective radiation.

Operation on the 7 Mc/s band makes use of the vertical section only as an extended ground plane, the top loading coil serving as a choke. The feedpoint will be very high on the base-loading coil, much as for 1·9 Mc/s operation. The system can be made to resonate in the 14 Mc/s band as a vertical half-wave by bringing down the feed-point as before until the electrical length to this point is three-quarter wavelength, the portion of the base loading-coil above the feed-point resonating at 14 Mc/s with its own self-capacitance. Somewhat more advanced designs based on this conception have been described in which the optimum feed point is selected by switching. It is recommended that the feeder line be taken under-ground in order to minimize stray radiation which may disrupt the low-angle pattern. In setting up such a system it is very advantageous to make use of an s.w.r. meter so that the best matching can be quickly found, but providing the feeder is short and preferably near to an electrical half wavelength in the 1·9 or 3·5 Mc/s bands, and that the transmitter loads up well, results are unlikely to prove unsatisfactory.

MULTIBAND AERIALS

In the average location, an amateur will not have space for many aerials, and thus the problem of making one array work usefully on more than one amateur band is an important one. Throughout this chapter, care has been taken to indicate the multiband possibilities and limitations of the various types of aerial. This section puts the question the other way round and answers it with a few aerials designed specially to work on more than one band. The simpler aerials, such as dipoles, or long wires fed in various ways, will do this but there are always disadvantages—the radiation pattern changes, and usually there are high v.s.w.r. problems on the feeder.

Multiband Dipoles and Ground-planes

The first approach to a multiband dipole is shown in **Fig. 13.116(a)**. Dipoles are cut for the various bands and supported about one foot apart in the same plane, and then joined to a single 80 ohm twin line. The theory is that the aerial which is in tune at half-wave resonance takes all the power, but this simple theory ignores the coupling between the dipoles, and so the match to the line may not always be very good, and the multi-lobe patterns of the longer wires tend to appear when the shorter ones are active.

The second approach is to connect reactances into the aerial with one of two objects—(a) to cut off the aerial progressively for each frequency band, so that it is a dipole for each band, or (b) to use the reactance as a phase changer, so that at the higher frequencies the extensions behave like a collinear array. Method (a) will be described but method (b) is difficult to apply. **Fig. 13.116(b)** shows one way to do this, using resonant *LC* units (*traps*). Starting from the feeder, the radiator is cut to length for the highest frequency, say f_3. Parallel circuits, resonant at f_3 are then inserted, one for a ground-plane aerial, two for a dipole, and the aerial is then extended till it resonates at f_2, the next lower frequency. The procedure is repeated for a third frequency if required. Finally the lengths of the sections are re-trimmed for each band in the same order. The ideal value of $\sqrt{L/C}$ ratio for the traps would be equal to the characteristic impedance, Z_a, of the half-aerial, which makes $2\pi fL = Z_a$, but this would cause too much shortening of the next lower frequency section of aerial, and suitable values for reactance of L and C are $\frac{1}{2}Z_a$. At half their resonant frequency the traps provide together an inductive reactance of $Z_a/3$ which is equivalent to about one-sixth of the length of the lower frequency quarter-wave. Thus for a dipole working on 14 and 7 Mc/s, with a Z_a of 500 ohms, the first section would be 16 ft. long, the coil would be approximately 2μH and the capacitance 50 pF. The residue for 14 Mc/s would then be about 10 ft. Values of Z_a are taken for the full length of the aerial from **Table 13.12** and reactances can be translated into coils and capacitors.

A suitable coil for the above example would comprise eight turns of 16 s.w.g. wire 2 in. diameter and 2 in. long supported by polystyrene strips cemented in position. The coil should be mounted over a long insulator and the capacitor, which should be of the stacked mica type and rated at not less than 500 volts, mounted along the insulator. The whole assembly should be tuned to resonance by adjusting the coil, using a grid dip oscillator and then sealed into a pelythene bag. The aerial described above, will also work quite well on 21 Mc/s.

The theory of operation described is approximate, and does not allow for the coupling which exists between aerial sections. It is therefore advisable to adjust the aerial during construction.

A suitable trap design for multiband rod aerials.
(By courtesy of Mosley Electronics Ltd.)

TABLE 13.12
Characteristic and Anti-resonant Dipole Impedances

Ratio L/D	Char. Imp. Z_a	End R_e (λ2)	Centre Rc (λ)	Centre Rc (2λ)	Typical Aerial
15,000	500	4,200	5,000	3,500	66 ft. of 14 s.w.g.
10,000	480	3,800	4,600	3,200	33 ft. of 14 s.w.g.
5,000	450	3,400	4,000	2,700	16 ft. 6 in. of 14 s.w.g.
2,500	400	2,700	3,200	2,200	—
1,000	350	2,000	2,400	1,600	66 ft. 1¼ in. dia. *
500	300	1,500	1,800	1,250	33 ft. 1¼ in. dia. *
200	250	1,000	1,300	800	—
100	210	750	900	550	12½ in. by ¼ in. dia.

This table gives the quarter-wave characteristic impedance, Z_a, the end-impedance R_e of half-wave dipoles and the centre impedance R_e of full wave and 2λ aerials, in terms of conductor length/diameter (L/D) ratio. R_e is based on a radiation resistance of 60 ohms, and the full wave R_e on 100 ohm per half-wave. Values may vary by 20 per cent in practice due to environment.
* Cage radiator.

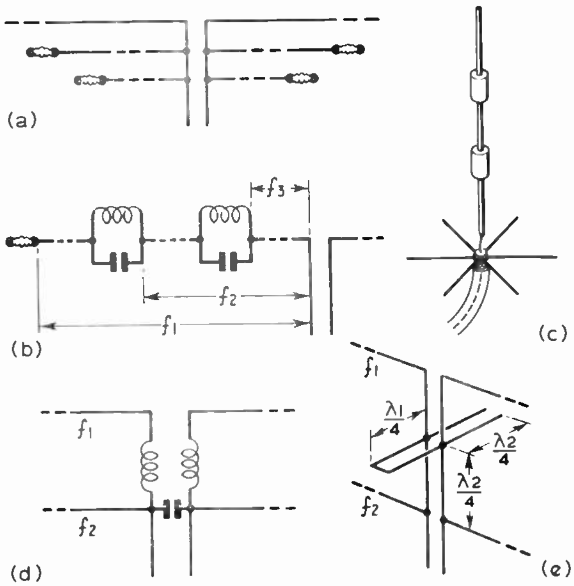


Fig. 13.116. Multiband arrangements. (a) Parallel dipoles. (b) Multiband dipole. (c) Ground plane equivalent of (b). (d and e) Two "frequency switch" filters for feeding two aerials over one transmission line.

When the process is applied to a ground-plane aerial, the physical construction is simpler, because the aerial can be made on a self-supporting tube. The tube can be broken and supported by rod insulators over which the coil is wound; the trap circuits can then be well protected by tape, or even encased in Fibreglass. A " whip " can be used for the top section. The ground-plane radials may not work well on all three bands, and some longer wires should be provided, if possible, for the lower frequencies.

When it is required to use two aerials on different bands with a common feeder, this can be done by inserting a low pass filter immediately after the high frequency aerial (**Fig. 13.116(d)**). The filter has a cut-off midway between the two frequencies: at this cut-off frequency the reactance of the capacitor is equal to the impedance (Z_0) of the line, whilst each inductor has half this reactance. An alternative form of filter (**Fig. 13.116(e)**) uses stubs connected a distance $\frac{1}{4}\lambda$ from the first (f_2) aerial. One stub is a $\frac{1}{2}\lambda_2$ (i.e. $2X\lambda_2/4$) and is open-circuit; this therefore presents a short-circuit on to the line at f_2 and thus the second stub has no effect on the first. The short circuit becomes a high impedance at the second aerial (f_1), by transformation. The second (shorted) stub tunes out the first at f_1 when the total length of the two stubs is $\lambda_1/4$. The length of line from the second stub to the first aerial is not important.

The G5RV 102 ft. Dipole

The 102 ft. dipole has been found an excellent compromise suitable for all h.f. bands and can be fed either of the ways illustrated in **Fig. 13.117**. In the upper diagram, tuned feeders (300 to 600 ohm) are used all the way: in the lower version the high impedance feeder is 34 ft. long and is connected into a 72 ohm twin or coax line. At this junction, the aerial impedance due to standing waves is low on most bands, as can be checked with the aid of Fig. 13.54 for a length of 34 + 51 = 85 ft.

The aerial should be supported at the optimum height for the band which is considered most important for DX working; that is, a half or full wavelength above ground. It is perhaps better to arrange this for 14 Mc/s at which frequency the aerial is designed to present a fairly close impedance match to 72 ohm coax or twin-lead via the 34 ft. stub which, in this case, acts as a one-to-one impedance transformer.

On 1·8 Mc/s the two feeder wires at the transmitter end are connected together or the inner and outer of the coax joined and the top plus " feeder " used as a Marconi aerial with a series-tuned coupling circuit and a good earth connection.

On the 3·5 Mc/s band, the electrical centre of the aerial commences about 15 ft. down the open line (in other words, the middle 30 ft. of the dipole is folded up). The aerial functions as two half-waves in phase on 7 Mc/s with a portion " folded " at the centre. Although the 72 ohm feeder " sees " a somewhat reactive termination it loads satisfactorily and radiates effectively.

At 14 Mc/s the aerial functions as a three half-wavelength aerial with an effective all-round low-angle polar diagram. Since the impedance at the centre is about 100 ohms, a satisfactory match to the 72 ohm feeder is obtained via the 34 ft. of half-wave stub. By making the height a half-wave or a full-wave above ground at 14 Mc/s and then raising and lowering the aerial slightly while observing the standing wave ratio on the 72 ohm twin-lead or coax feeder by means of an s.w.r. bridge, an excellent impedance match may be obtained on this band.

On 21 Mc/s, the aerial works as a slightly extended two-wavelength system or two full waves in phase and is capable of very good results especially if tuned feeders are used to reduce loss. On 28 Mc/s it consists of two one-and-a-half wavelength in-line aerials fed in phase. Here again, results are better with a tuned feeder to minimize losses although it works satisfactorily with the 34 ft. stub and 72 ohm feeder.

When using tuned feeders, it is recommended that the feeder taps should be adjusted experimentally to obtain

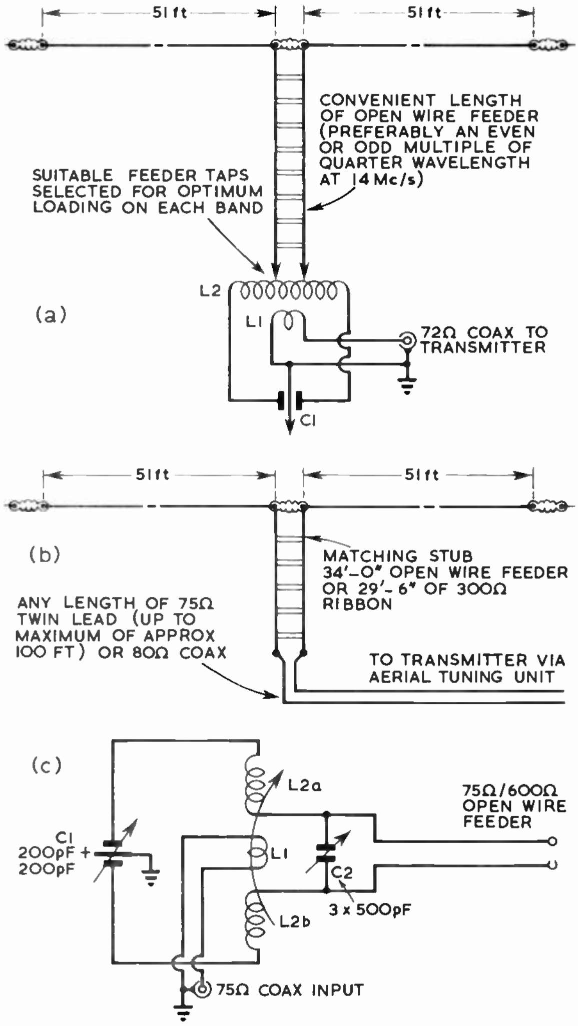


Fig. 13.117. Two versions of a simple but effective multiband aerial for 1·8–30 Mc/s. L1 is the coupling coil and C1 L2 form a resonant circuit at the operating frequency.

optimum loading on each band using separate plug-in or switched coils. Connection from the a.t.u. to the transmitter should be made with 72 ohm coaxial cable in which a suitable TVI suppression (low pass) filter may be inserted.

Multiband Parasitic Arrays: Frequency Switches

The principle of a stub or network used as a frequency switch, as in Fig. 13.116, can be applied to parasitic arrays to enable them to work on more than one band. The range of frequency is limited to 2 : 1 because, in terms of wavelength, the spacing will change, and at only one frequency can the optimum spacing be employed. For example a spacing of $0{\cdot}2\lambda$ at 28 Mc/s, which is rather higher than optimum, becomes $0{\cdot}1\lambda$ at 14 Mc/s, a spacing which gives good performance but critical tuning.

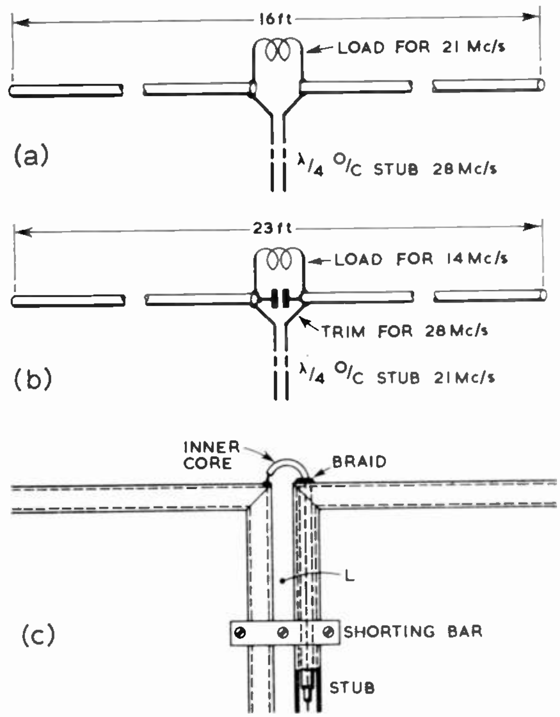


Fig. 13.118. Multiband parasitic elements. (a) is a director which operates on 21 and 28 Mc/s. (b) is a reflector which operates near half-wave resonance on 14 and 21 Mc/s and as a pair of reflectors on 28 Mc/s. (c) Cut-out showing how the twin support boom of the array can be used with shorting bars to provide the low frequency loading inductance, and how the stub, in this case concentric, can be stowed inside one leg of the boom.

Fig. 13.118 shows multiband elements. The radiator is made to a length suitable for one frequency, cut in the centre, and then closed by means of an open-circuit quarter wave stub. The gap is then loaded with inductance or capacitance which, together with the stub, loads the element to tune at a second frequency. At the first frequency the stub has zero impedance, and the loading reactance is therefore shorted out.

The director in Fig. 13.119(a) is made for 29 Mc/s with a quarter-wave stub at this frequency. The loading coil, which may be a few turns of wire about 1 in. diameter, is adjusted to give a second resonance at approximately 22 Mc/s. The tuning can be checked with a grid dip oscillator coupled to the coil or stub, but final adjustments should be made with the complete array in position. Allowing for velocity factor, the stub will be about 6 ft. 6 in. of 300 ohm twin, or 5 ft. 6 in. of 80 ohm twin feeder or concentric cable. The three-band reflector in Fig. 13.119(b) is adjusted to resonate, with its stub, at approximately 20 Mc/s, and loaded with a coil to work on 14 Mc/s. At 28 Mc/s the stub is longer than a quarter wave, and acts as an inductance, which is in parallel with the coil. The addition of a small capacity (a few picofarads) will produce parallel resonance with the two inductances, and can be adjusted to separate the two halves of the radiator into two separate reflectors for 28 Mc/s; it is necessary to retrim the coil for 14 Mc/s as this adjustment is made.

Fig. 13.119(c) shows how the loading inductance can be formed by connecting a shorting bar across a double boom which supports the array. The stub (concentric) is stowed inside one leg of the boom, with its inner and outer conductors connected across the gap in the reflector.

The Minibeam

A variety of arrays for use on more than one band, both in the h.f. and v.h.f. ranges have been developed by G. A. Bird (G4ZU), a pioneer in this field. The " Minibeam " model outlined in **Fig. 13.119** used the parasitic elements shown in Fig. 13.118, and operates on 14, 21, and 28 Mc/s. The radiator is resonated a little below 21 Mc/s: on 14 Mc/s it acts as a short radiator, and on 28 Mc/s as a " short " full wave element. The aerial is a five-element array on 28 Mc/s, with high gain: on 21 Mc/s it is a three-element array and the gain is claimed to be little more than for a normal Yagi, because of the extended radiator, whilst on 14 Mc/s it is a two element array with 3 to 4db gain, the director being ineffective.

The impedance of this array is very different from that of a normal three-element array, because of the untuned radiator, and the increased number of elements, and 300 ohm twin is recommended for feeding it. Adjustment follows the general lines indicated for two- and three-element arrays, but is of course more complex, as it involves three frequencies, and is mainly made to the loading devices, not the radiator. For

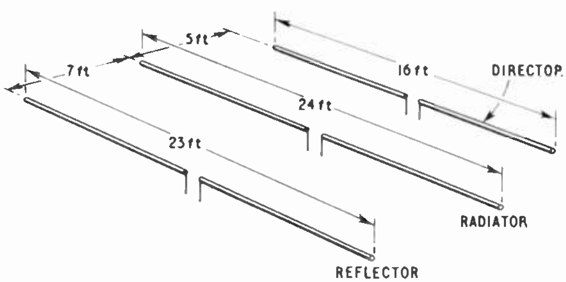


Fig. 13.119. The G4ZU " Minibeam " array. The radiator acts as a short dipole on 14 Mc/s, as a half-wave dipole on 21 Mc/s and as a full-wave dipole on 28 Mc/s. The aerial is equivalent to a five-element array on 28 Mc/s, a three-element array on 21 Mc/s, and a close-spaced two-element array on 14 Mc/s. The tuning of the parasitic elements is described in the text.

1 in. diameter elements, the free director resonance is about 5 per cent higher than the working frequency, the reflector about 5 per cent lower. Tuning should be carried out with the array as high as is practicable, since the elements may tune 1 per cent higher in frequency when erected than when near the ground. It is not too difficult to obtain good gain and front-to-back ratio on 21 and 28 Mc/s, but adjustment for 14 Mc/s is critical because it is a close-spaced beam on this band.

LIGHTNING PROTECTION

In order to achieve the most efficient radiation from an aerial, it should in general be erected in the clear and as high as possible. In such conditions it then becomes a potential hazard since it represents a very good lightning conductor. A strike on an unguarded aerial can have disastrous results in the radio room, causing damage which may be serious and even fatal. It is therefore prudent to give some thought to the question of lightning protection when erecting an aerial system.

Adequate protection can be obtained in two basic ways, which can loosely be described as *executive* and *automatic*. The first requires a positive action to achieve the necessary protection, and usually renders the station non-operational. The second is designed to give full protection at all times without materially affecting normal operation. It is ultimately prudent to cease operation during violent storms, but the flexibility of the automatic type of protection is generally highly desirable. It also gives a degree of suppression of noise arising from the general static discharges on the aerial common during the Summer, or from charged precipitation (rain static).

All lightning protection is based upon a low resistance d.c. path from the exposed conductor down to a true earth. This can be achieved in a number of ways. It is however of paramount importance to note that a good low resistance earth return is required, of the quality described in the earlier section on Earth Systems. *Under no circumstances* should the public electricity supply mains earth or a water-pipe system be used for lightning protection.

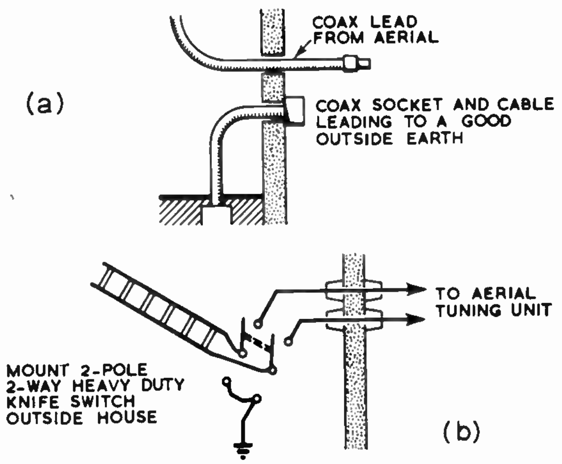


Fig. 13.120. Lightning protection using a switch for open wire line or an earthed socket for coaxial cables.

Executive Protection

This is achieved by the physical making of an earth connection to an otherwise live feeder or aerial, by means of suitable switches. Its very nature is such as to short out the correct operation of the aerial and feeder system, so that the operator must use his own discretion when deciding at what stage to protect the equipment. The most convenient way is to earth the downcoming feed line at the lowest point *outside* the building. This can be achieved with an open knife switch on balanced line, or by a special earthed socket for coaxial lines. The latter are equally a potential hazard even when the outer is firmly bonded to earth, unless the aerial design is such as to present a good low resistance d.c. connection across the upper end of the coaxial line. These two arrangements are shown in **Fig. 13.120.**

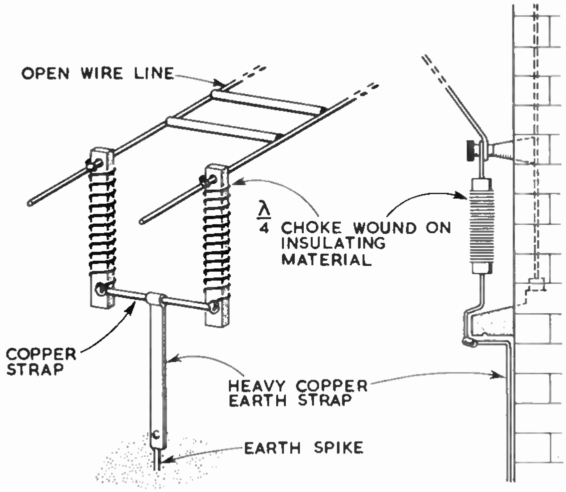


Fig. 13.121. Lightning protection using static leak chokes.

Automatic Protection

This can be obtained by connecting the d.c. earth to the feeder (or aerial) in such a way that it represents an open-circuit connection at r.f. The devices employed are often called *static leaks*. For the lower frequencies, two large chokes constructed from heavy gauge wire (16 s.w.g. or larger) can be connected in parallel across balanced lines, and their common connection earthed. For unbalanced systems, such as Marconi and coaxially fed aerials, a single choke from the live conductor to earth will suffice. The reactance of such chokes cannot be made infinite, and they will have some effect upon the line performance. This can usually be tuned out in the aerial coupler, and to that extent their operation can be considered independent of frequency. They should however always be designed to have a reactance equal to at least 4 to 5 times the line impedance at the lowest operating frequency, and preferably as high as a factor of 10, e.g. $X_L = 6000$ ohms, for open wire line, or 1000 ohms for low impedance coaxial cable. Typical layouts are shown in

Fig. 13.121. Where the inductive reactance is embarrassingly low, attempts can be made to parallel resonate it with a suitable capacitor to improve the isolation at radio frequencies. A more elegant arrangement for single frequency working can be constructed using the principle of the " metal insulator " described on page 13.16. A short-circuited quarter wave stub can be connected across a line, and have virtually no effect since its input impedance is very high: in fact it looks like a parallel resonant circuit of extremely high Q. A suggested layout is illustrated in **Fig. 13.122**, for both balanced and coaxial lines. This technique is of course limited to the higher frequency bands where the physical length of the stub is not unwieldy: it is particularly suited to the protection of balanced open line arrays at v.h.f. It should be repeated here that although automatic static leaks are useful for conditions of high static charge level and during minor storms, it is still strongly recommended that the more positive isolation of the executive type be *employed at all times* when a severe storm is in progress and a substantial strike is possible. The ultimate isolation can of course be obtained by disconnecting all feeders completely outside the building and leaving the free ends well clear in the garden.

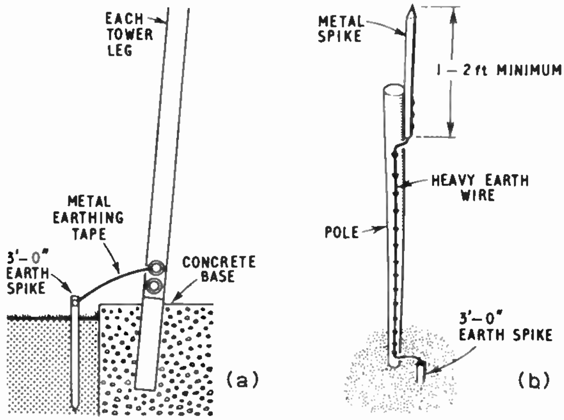


Fig. 13.123. Earthing of towers and poles is important as a safeguard against lightning. Aluminium tape should be used if the tower leg is galvanized steel and the bolted joint painted over to avoid electrolytic corrosion.

Protection of Masts and Towers

In addition to protecting the equipment in the station from lightning damage via the feed lines, some attention should also be paid to masts and towers, which, if struck violently, could cause damage by collapse. All metal poles and towers should be positively earthed at their base to a proper earth connection (**Fig. 13.123**). The indifferent earth achieved by the concrete or rammed earth around the bottom of the structure

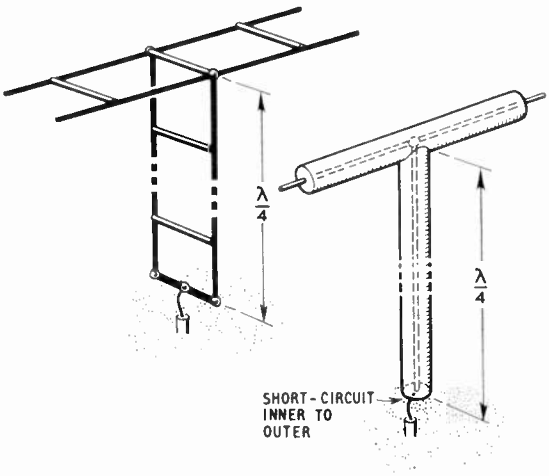


Fig. 13.122. Quarter-wave lightning protection. The balanced line is approximately a quarter-wave in free space: the coaxial line is λ/4 V where V is the velocity factor of the cable employed to make the stub.

is not sufficient. For wooden structures a heavy wire should be run up the side from a good earth, to terminate in a short vertical metal spike at the top, projecting 1 to 2 ft. above the actual structure.

CONSTRUCTING AERIALS

All the aerials described in this Chapter can be constructed at home without much difficulty provided that suitable materials are selected. They fall approximately into two groups, those which are formed from wires under tension between anchoring points (e.g. dipoles, Zepp, Windom, etc), and those which employ rigid or semi-rigid conductors either self-supporting or mounted on light rigid frames (Yagis, quads, verticals, etc). All rotatable beam aerials come in this group.

Wire Aerials

The cardinal points of design of a fixed wire type aerial are:

(a) Choice of conductor type and gauge,
(b) insulators, and
(c) end anchorages for tensioning.

The size of conductor to be employed depends upon the working tensions involved. The longer the span and the heavier the insulators, the greater will be the tension for a given amount of sag. Single copper conductors should be avoided particularly in the larger wire gauges (> 20 s.w.g.) as they are difficult to handle in long lengths, tend to be naturally springy, and are prone to kinking and breakage if nicked accidentally. Ideally wire aerials for the h.f. bands should be made up from 7/0·029 in. stranded hard drawn copper wire, although finer stranded sizes, fine braided wires or even single 30 s.w.g. enamelled conductor can be used if lightweight insulators and light tensions are employed. The latter are of course much less conspicuous in areas where aesthetic requirements call for a near-invisible aerial.

Insulators can be obtained as proprietary items in glazed china or glass, or can be manufactured at home from polythene or Tufnol sheet or rod. In the case of the latter, an all-weather quality is essential which should be varnished, particularly over cut edges, to preclude de-lamination. It is also possible to cast individual tension and centre feed point insulators in *Araldite* resin, which has an acceptable power factor and low dielectric loss up to 30 Mc/s.

The correct method of attachment of halyards and conductor wires to various forms of tension and insulators is shown in **Fig. 13.124(a).** To form the P.O. splice in the stranded conductor, one strand of the free end of wire is untwisted back to the insulator, and then wrapped tightly around both wires to bind them together. A second strand is then untwisted back to the end of the first wrapping, and wrapped to continue the binding action. The process is repeated until all the strands of the free end have been used. This neat splice may also be used on galvanized stay wires for masts, and is very strong. It is aesthetically better and inherently safer as a permanent splice, than the familiar Bulldog wire rope grips sometimes employed.

The preferred method of securing the feed-point insulator for coaxial and for open-wire line is shown in **Fig. 13.124(b).** It is important to ensure that the weight of the line is supported firmly by the insulator and not by the connections made off to the aerial conductor. The latter should be wrapped to form a sound mechanical joint before soldering.

Aerial halyards can readily be made up from the plastic covered clothes line available in hardware shops, or from the finer cords used in yacht rigging. They should always be reeved as an endless loop to avoid a climb to the top of the pole if an aerial or halyard connection fails **Fig. 13.125(a).**

Wooden or metal poles, towers, or the walls or eaves of buildings are all suitable tension anchorages. In the latter case a long free end to the halyard is preferable to keep the aerial away from any immediate guttering. The branches or trunks of trees are not recommended as halyard anchors as they move excessively in the wind and cause large variations of sag in aerial wires, and possibly failure under extreme tension. If they must be used, then some form of counterbalance on the lower end of the halyard is essential **Fig. 13.126(b).** The weight should be free to move up and down, and is equal to the tension in the aerial wire.

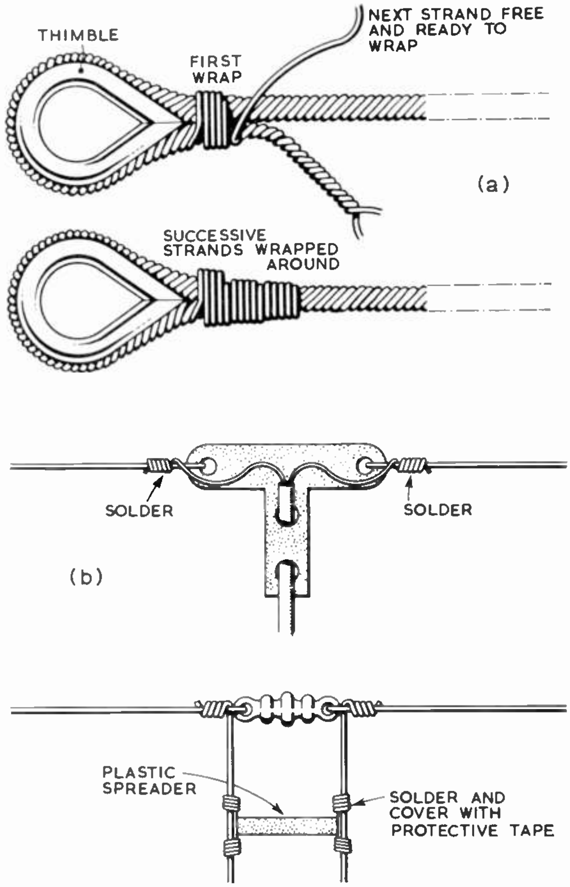


Fig. 13.124. (a) The PO splice for securing stranded wire around a thimble or insulator. (b) (c) Methods of supporting feeders from dipole centre insulators. The exposed end of the coaxial cable in (b) must be taped to exclude moisture. The "tee" insulator is described in the text.

Beam Aerials

Various forms of beam aerials can be constructed either from wires tensioned on a framework or from self-supporting tubes. Wires may generally be of finer gauges than for ordinary wire aerials since the spans and tensions are generally less. A suitable size of wire for a quad would be 20 s.w.g. enamelled. Insulators can also be smaller and more compact. The framework can be made up from suitable timber which can be screwed together, or from bamboo poles clamped to end fixing plates. When timber is used, particularly for rotary Yagi type aerials, a hardwood is to be preferred. Softwoods should always be primed and painted for normal outdoor protection, or given a thorough creosote or polyurethane treatment.

The most elegant form of Yagi aerial construction is the "plumbers delight," so called because all the elements and the supporting boom are made from tube sections. Since all elements in a Yagi are at zero r.f. potential at their centres, they may be joined to and supported by a metallic boom. This form of construction is used exclusively in commercially available aerials, and the prices of these are sufficiently low to make it uneconomic for the amateur to try to build his own beam to the same necessary engineering standards. Particular problems for the home constructor are the cross-tube clamps, and the centre insulator for the driven element, both of which can be commercially mass produced far more cheaply.

Vertical Aerials

Vertical aerials may be either in the form of a wire tensioned between top and bottom insulators along a pole, or more usually self supporting. In this case the great problem for the home constructor lies in the provision of a suitable base insulator which can also take the weight of the aerial and hold it vertical. Insulated stay wires are a solution to the second aspect, but are to be avoided if possible because of their inevitable interference with the radiation pattern of the aerial, even if well broken up with insulators. Again, the commercially available trap verticals for multiband operation are so cheap as to render it uneconomical for the amateur to construct his own. For those who wish to erect a single band vertical, a useful base insulator can be made from a soft drink or wine bottle, clamped around the lower part of the outside, and with the bottom end of the vertical tubing either inside or over the neck of the bottle. A hock bottle is probably the most suitable shape for this purpose. In all such cases it is essential to prevent the bottle filling up with rainwater.

Poles

Although traditionally aerials have been supported by stayed or self-supporting wooden poles, usually creosoted for weather protection, there is an increasing tendency to use steel or aluminium tubing with suitable coupling and stay attachment fittings. A most convenient mast kit can be built using nominal 2 in. diameter aluminium *scaffold* poles as the basis, and selecting fittings from a range manufactured by Gascoignes Ltd., Reading, Berks, under the trade name "Kee Klamps." A typical arrangement for a rotating lightweight tubular mast suitable for carrying light h.f. or v.h.f. beams is shown in **Fig. 13.126.** A simpler form using the same principle can be employed for fixed poles.

An alternative source of tubes and clamp fittings can be obtained from firms specialising in TV aerial installations particularly in fringe areas of reception.

Planning Consent

In the UK all aerial supports erected at a height exceeding 10 ft. a.g.l. are subject to the provisions of the Town and Country Planning Acts. Before embarking upon any major aerial erection, it is advisable to consult with the Surveyor to the local Council and if necessary make formal application for Planning Consent. Failure to comply with the Town and Country Planning Acts could possibly lead to an injunction to dismantle the entire erection at a later date.

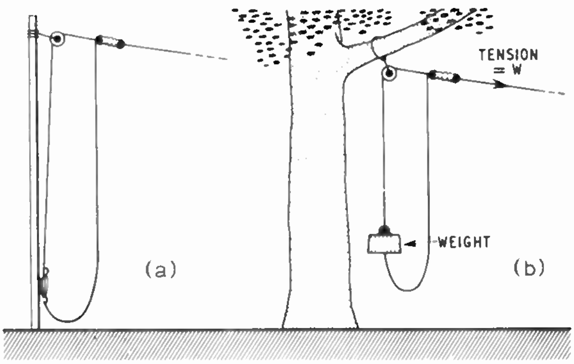


Fig. 13.125. Halyard connections to poles and trees. The weight W is equal to the tension T required in the aerial wire. In both cases an endless loop is employed.

In other countries various zoning laws also apply in built-up areas, and a similar procedure is recommended. When purchasing or renting property, it is important to ascertain whether there are any covenants in the title deeds prohibiting the erection of outdoor aerials. This is particularly important in municipally owned estates and in areas where piped television is in use.

AERIAL MEASUREMENTS

Measurement of the various aspects of aerial performance is a specialized art. It differs from the normal techniques of laboratory work in as much as the aerial under test can be greatly influenced by its surroundings and unless these are selected with sufficient care, the measured results will be meaningless in terms of the real performance of the aerial when erected to its normal operating position. It is not even sufficient to make measurements on a known test aerial in the same position, and to use these as a standard since the extent to which the surroundings influence any aerial depends upon the aerial itself.

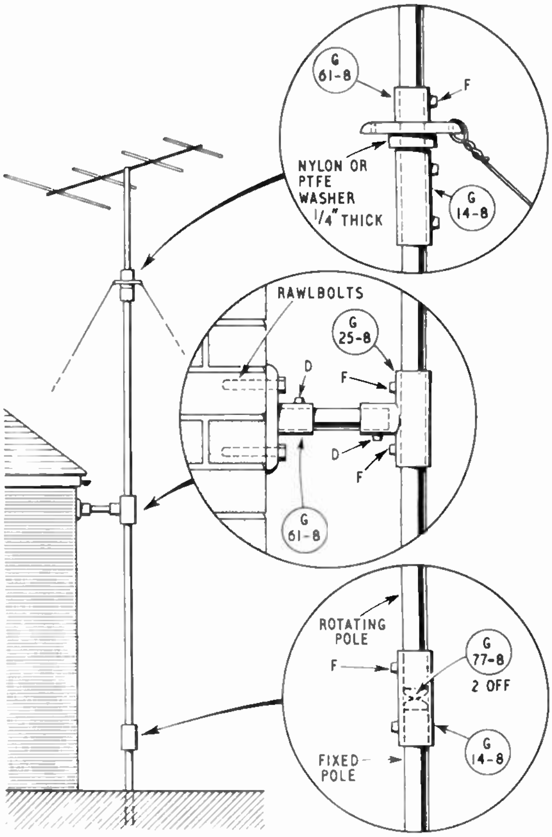


Fig. 13.126. A light rotating mast using Kee-Klamp scaffold fittings. The clamping screws at F should be left slack and grease forced into the fitting at those positions. For extra security the stub pole can be drilled ¼in clearance at points D and a ¼in BSW bolt substituted for the Allen screw, to penetrate the wall of the tube and form a lock against possible pull out. A simpler form of construction can be used for fixed poles.

The measurement of aerial performance usually involves three particular properties of the aerial:

(i) Impedance,
(ii) Polar diagram (in both planes), and
(iii) Gain relative to a reference standard.

These are listed in increasing order of difficulty of measurement.

Impedance

It is rarely necessary for the amateur to know the precise input impedance of an aerial in absolute terms. The object of most, if not all adjustment of impedance is to match the aerial to a transmission line to give a minimum s.w.r. on that line. Whereas impedance bridges of any accuracy are difficult to construct, it is readily possible to build an instrument which will indicate the s.w.r. on a line fairly accurately in terms of separate indications of the forward and reflected waves. Such an instrument is known as a *reflectometer*, and when used in a test set-up as shown in **Fig. 13.127**, can be used to indicate correct adjustment of the aerial for minimum reflected power in the line. When carrying out such adjustments it is important to maintain a constant forward power from the test oscillator, and this too can be checked from time to time using the appropriate position on the reflectometer.

Adjustment of aerial impedance should ideally be carried out with the aerial in its operating position, since this is affected by height above ground. Such adjustment is therefore a compromise between height and accessibility for modifications to the aerial.

When a highly directional beam aerial is to be adjusted a useful arrangement is to lay the beam on its " back," facing upwards. The low amount of back radiation will reflect from the earth to give only a small effect upon the aerial impedance, and the height above ground need only be 3-4 ft. This technique is particularly applicable at v.h.f. and u.h.f.

It is permissible, and usual, to leave the reflectometer connected in the transmission line during normal operation (with a reduction of sensitivity appropriate to the transmitter power employed). It is then possible to maintain a continuous check on the s.w.r. and detect aerial faults as they arise.

When aerials using tuned feeders are employed, the adjustment of impedance is transferred to the aerial coupler (see page 13.35). Again a reflectometer may be used in a short piece of line of correct impedance for the transmitter output load, and the aerial coupler adjusted to give minimum s.w.r. on this piece of line. This is shown in Fig. 13.44.

The theory of coupled line reflectometers is covered in Chapter 19 —*Measurements*, and a design is shown which is ideally suited to v.h.f. However, the sensitivity of this type

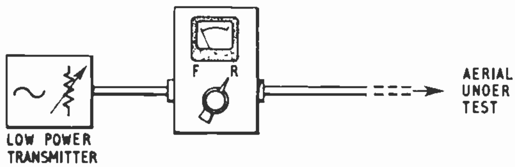


Fig. 13.127. Measurement of aerial impedance using a reflectometer. The output of the transmitter is set to maintain a constant indicated forward power. The aerial is then adjusted for minimum reflected power to *match its impedance to the cable employed*. The forward power should be checked regularly for any changes in transmitter loading.

of reflectometer is rapidly reduced as the frequency is lowered and the coupling lines become very short fractions of a wavelength. To increase the sensitivity by lengthening the instrument would not be practicable, the 144 Mc/s model requiring to be ten times as long for 14 Mc/s. An alternative type of coupling arrangement is to be preferred for the h.f. bands, where the *L* and *C* of the " coupling bridge " are obtained by lumped components.

A type of reflectometer, developed by G3FRV, which has a sensibly constant sensitivity with frequency is shown in **Fig. 13.129.** Commercial instruments of this type have accuracies of ± 5 per cent when used as power measuring devices.

The inner conductor of the coaxial line is passed through the centre of toroidal current transformer T1, and the screen or outer conductor bypasses the transformer. Current flowing in the inner conductor, which acts as the primary of the current transformer, induces a current in the secondary winding of T1 resulting in out-of-phase voltage components *E1* and *E2* appearing across R1 and R2. Two capacitive voltage dividers C1/C2 and C3/C4 develop equal and in-phase voltage samples *E3* and *E4* across C2 and C4. *E1* and *E2* are proportional to the vector sum of the forward and reflected currents, and *E3*/*E4* are voltage samples proportional to the vector sum of the forward and reflected voltages.

When the line is correctly terminated, and reflected voltage and current non-existent, the component values are proportioned to cause the voltages *E2* and *E4* to cancel and *E1* and *E3* to add, giving a maximum forward power indication, and a zero reflected power indication. When the line is incorrectly terminated, it can be shown that indications proportional to forward and reflected power are given on the meter. The values of the series resistor R5 may be adjusted, using a calibrated test load, for a given full scale power, alternatively two switched sensitivities may be provided, say 10 and 100 watts. For operation on s.s.b., C5 and C8 give a long discharge time constant, so that the meter tends to indicate peak envelope power on speech.

Setting up the unit is very simple. Operate the reflectometer with a suitable transmitter and a calibrated load, and adjust C1 for minimum reflected power as indicated. Reverse the input/output connections to the reflectometer, and repeat this operation using C3.

Using the circuit values given, an approximate f.s.d. sensitivity (in forward power position) of 100 watts will be obtained. With R5 at 15 K ohms, an input sensitivity of 10 watts will result.

Polar Diagram

The horizontal polar diagram of an aerial can be measured to a reasonable degree of accuracy without having to raise the aerial to an excessive height above ground. A figure of 0·2-0·3 wavelength may be considered adequate, with a minimum of 8 ft., to ensure adequate general clearance. At

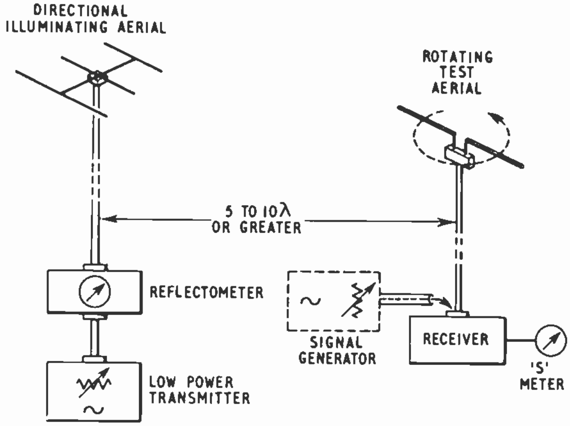


Fig. 13.128. Test site for measurement of aerial radiation patterns. The site must be free from obstructions: a method of proving this is suggested in the text. The reflectometer is used in the forward position to monitor the transmitter output which must remain constant. The S meter should first be calibrated using a signal generator into the receiver input to prove the law of the receiver.

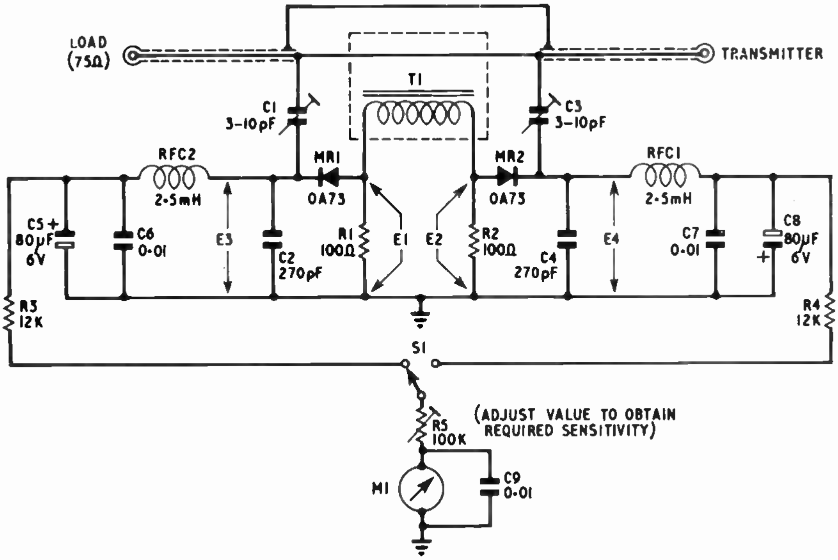


Fig. 13.129. A wideband reflectometer for 1–30 Mc/s.

C1, 3, 3-10PF philips type; C2, 4, 270pF ± 2 per cent silvered mica; C5, 8, 80 μF 6V d.c.; C6, 7, 9, 0·01 μF ceramic disc; M1, for powers up to 20W, 100μA f.s.d., for powers above 20W a less sensitive meter may be suitable with appropriate adjustment of R5; MR1, 2, OA73 or similar; R1, 2, 100 ohms ± 5 per cent carbon, ¼ W carbon; R3, 4, 2·2K ohms ± 5 per cent, ¼ W carbon; R5, 100K ohms (adjust value to obtain required sensitivity); RFC1, 2, 2·5mH r.f. choke (Bulgin type SW68 or similar); T1, *primary* inner conductor of coaxial cable, *secondary* 44 turns 26 s.w.g. enamelled copper wire evenly wound around the core, core GEC toroid type G29S. It is important to keep the unscreened section of the cable very short in order to avoid appreciable impedance discontinuity.

the lower frequencies for which beam aerials are practicable and for which the polar diagram becomes of interest, the minimum height will rise to 12 ft. More important is the distance from the test or *illuminating* aerial, and the nature of the ground and side obstructions. The general test layout is shown in **Fig. 13.128.** The object of planning a test site is to ensure that the aerial under test (which is generally used for receiving) is evenly illuminated by a constant plane field from the other aerial. This can only be achieved by establishing a reasonable distance between them ideally not less than 5–10 wavelengths and by avoiding reflections of the radiated field from nearby obstructions. This can be overcome by using a directional aerial at a sufficient distance that the aerial under test is totally within the main beam (± 1db), and that little radiation falls outside this beam to be reflected from obstructions along the side and rear, e.g. fences, houses, trees. A useful check on the validity of the test site can be obtained by rotating a halfwave dipole at the test aerial end, and comparing the measured pattern with the theoretical one (see page 13.44). For this purpose the dipole *must* be fed with a truly balanced feed to avoid pattern asymmetry.

Gain

This is the most difficult thing of all to measure, even for professionals. The method open to the amateur is called the *substitution* method and is shown in **Fig. 13.130.** A standard field is laid down by a distant aerial, and the level of the signal received on a reference aerial, usually a halfwave dipole, is compared with that received on the aerial under test, the voltage ratio expressed in db being the relative gain. Because the gain of an aerial depends upon its radiation pattern in both planes, it is essential that a uniform plane wave front is laid down at the aerial under test, as explained in the preceding section. Any spurious reflected signals will be picked up in differing amounts on the test and reference aerials due to the difference in their radiation patterns, and will give rise to erroneous answers. For this reason any measurements of gain attempted on any but a totally unobstructed site will be almost meaningless.

Carefully obtained measurements *under correct conditions* will give a fair approximation to the free-space gain, particularly at v.h.f. when the heights of the aerials above ground can readily be made greater than several wavelengths.

However the vagaries of ionospheric propagation and the effect of vertical polar diagram differences upon the radiated field at the desired wave angles mean that the actual gain achieved in practice for any particular aerial at h.f. will depend very much upon its operating conditions, and caution

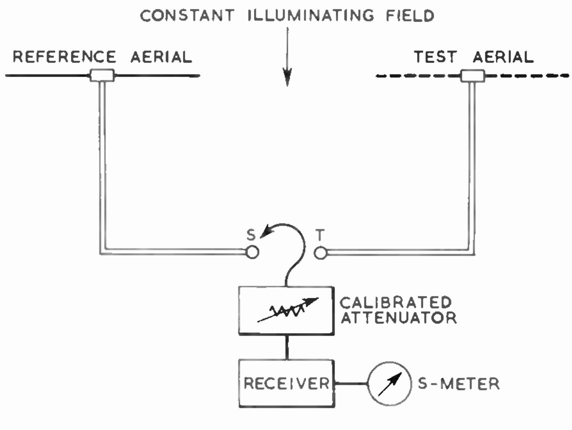


Fig. 13.130. Measurement of gain by substitution. The reference aerial is connected to the receiver, the gain of which is adjusted to give a suitable deflection on the meter with a small amount of attenuation. The test aerial is then substituted and the attenuator setting altered to give the same meter reading. The difference in the attenuator setting is a measurement of the relative gain of the aerials.

should be exercised when interpreting the meaning of gain figures both measured and claimed.

CHOOSING AERIALS

This chapter has attempted to explain aerials and the principles underlying their design: a large number of aerials have been illustrated together with many accessory devices. The beginner may well be confused by the range of information, and therefore need advice on the best way to start.

The beginner is no doubt ambitious to do the things other amateurs do, but must not expect to succeed immediately. It is advisable therefore not to start off with the latest exotic " wonder-worker "; it will be found very puzzling, even if it can be made to work. Most pleasure will be obtained and more will be learnt by starting with a simple aerial which is bound to work. For instance, start with a simple one-band dipole or, if there is no space, a ground-plane on the roof, and then graduate to one of the simple multiband aerials (e.g. Figs. 13.83 or 13.116). If the dipole can be erected 40 ft. high it will be found that on 7 Mc/s the radiation angle is rather high, but it will be possible to work up to 1,000 miles or more with it, and on higher frequencies it may be possible to span the oceans.

Such aerials will be excellent during peak propagation conditions: to do better either more space or a compact beam is needed. The latter can be purchased, but much more will be learnt by building a two or three element rotating array or a " Minibeam " at home.

Those fortunate enough to be able to choose a site, should pick one with plenty of clear space, preferably on high ground. With such a site, success comes more easily. Aerials designed for low angle projection, will increase the hours per day during which stations in, say, U.S.A. or Australia can be worked. The long wire aerials are easiest to erect, though there may be short periods when a simple aerial will be better. Ground-plane aerials give outstanding results when propagation is normal but though it will always pay to instal them solely for receiving, they may more easily cause broadcast interference than horizontal aerials. With more than one aerial, or even for multiband work, preset aerial couplers will be found a great advantage.

RECEIVING AERIALS

At the beginning of this chapter it was stated that the receiving problem is not quite the same as the transmitting one. The reciprocity theorem as applied to aerials is well founded but does not allow for interference coupled to the equivalent transmitter-receiver system. The rule concerning the use of the transmitting aerial for reception is a good one; tuned arrays, especially the unidirectional ones, can be a great help for reducing interference from directions not in the beam of the aerial. Long wire types are not so good in this respect, and often bring in more cosmic or ionospheric noise than other types: vertical aerials normally produce a much cleaner background than any others, and help to discriminate against short-range interference—say from 500 to 1000 miles, because of their selective vertical patterns. A properly constructed aerial, with a good transmission line always helps to minimize man-made interference radiated by house wiring, etc.

It may be considered worthwhile to erect special aerials for the receiver and **Fig. 13.131** is an example of what can be done. A pair of ground planes are erected at a spacing of $\lambda/2$ to λ, the feeders brought in separately, and connected to the ends of a half-wave cable delay line. The delay line (which can be 24 ft. of coaxial line for 14 Mc/s) is coiled and its inner conductor brought out to switch studs at many points. When the receiver input is switched along this delay line, the horizontal patterns shown in Fig. 13.64 can effectively be rotated to steer the nulls against interfering stations.

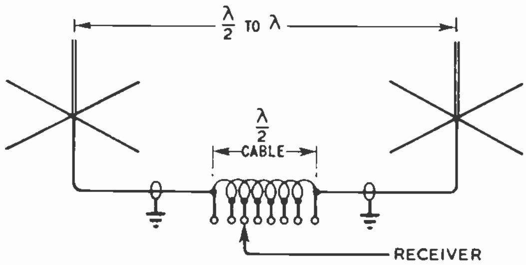


Fig. 13.131. Special arrangement of two ground-plane aerials providing electrical steering of the patterns given in Fig. 13.64. This is a great help for interference reduction in a receiver on noisy bands like 7 or 14 Mc/s, though it is not any great advantage in transmission because the patterns are very broad.

INTERFERENCE

Even the best communication receiver will sometimes show signs of overloading when supplied with the full output of a very large array, such as a *V* or a rhombic. The effect is a large increase in background noise of a type which is clearly the interaction between high power stations in or near the working frequency; often a pair of stations will beat together to give an i.f. signal which does not respond to the r.f. tuning of the receiver. This is most likely to occur on the 14 and 21 Mc/s bands, and is recognized because it disappears more rapidly than the tunable signals when the gain of the receiver is gradually reduced. It really calls for preselection in front of the receiver. Since the normal outside noise level from cosmic and man-made interference overrides the internal noise of a good receiver, there is no need to give it the full output of a large aerial, and often a smaller aerial can be used.

In order to prevent interference with broadcasting, special steps are taken in the transmitter to prevent overmodulation, key clicks and harmonic output. Loose connections in and around the aerial system can undo this work and rectify the transmitter output, causing serious interference of a harmonic type. Corroded or oxidized metal contacts make rectifiers, and can act as if a diode were connected in the aerial. Offending items include poor or broken solder joints, lead-through terminals, dirty r.f. connectors in the cable system, corroded earth connections, and loose metal joints such as those in gutters or rusty wire fences. Joints in the aerial should be carefully cleaned, made tight before soldering, and arranged so that the strain on the wire does not pull on a sharp bend. The solder joint may fracture in time through vibration, and should not therefore take the strain: it should also be protected with tape, so that it does not become too dirty to re-solder. Regular inspection of the aerial system and its surroundings is necessary.

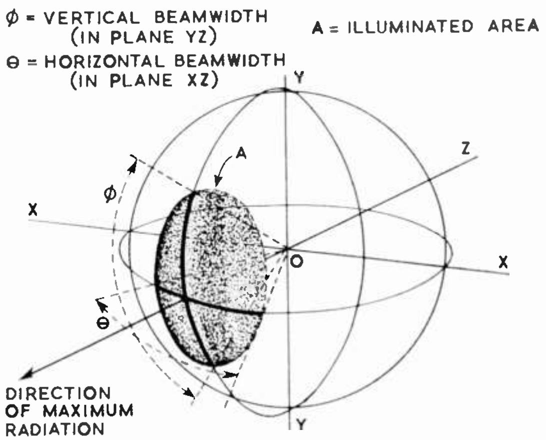
V.H.F./U.H.F. AERIALS

THE range of frequencies considered in this chapter extends from 30 Mc/s to 3000 Mc/s. In accordance with the accepted terminology a distinction ought strictly to be made between the range 30–300 Mc/s which is described as *very high frequency* (*v.h.f.*) and the range 300–3000 Mc/s which is described as *ultra high frequency* (*u.h.f.*). Over these two ranges, however, the problems of aerial design are often quite similar in character and to avoid unnecessary repetition the term *v.h.f.* is used here to cover both of them.

Above 30 Mc/s the wavelength of the radiation becomes short enough to allow efficient aerials of comparatively small dimensions to be constructed, since the effectiveness of an aerial system generally improves when its size becomes comparable with the wavelength that is being used. A special feature of v.h.f. aerial design is the possibility of focusing the radiated energy into a beam, thereby obtaining a large effective gain in comparison with a half-wave dipole which is the standard by which most v.h.f. aerials are judged.

Power Gain and Beamwidth

A simple way to appreciate the meaning of aerial gain is to imagine the radiator to be totally enclosed in a hollow sphere, as shown in Fig. 14.1. If the radiation is distributed uniformly over the interior surface of this sphere the radiator is said to be *isotropic*. An aerial which causes the radiation to be concentrated into any particular area of the inside surface of the sphere, and which thereby produces a greater intensity than that produced by an isotropic radiator fed with equal power, is said to exhibit gain relative to an isotropic radiator. This gain is inversely proportional to the fraction of the total interior surface area which received the concentrated radiation.



$$\text{POWER GAIN} = \frac{\text{SURFACE AREA OF SPHERE}}{\text{SURFACE AREA OF "ILLUMINATED" REGION}}$$

Fig. 14.1. Radiation from an aerial. An isotropic radiator at the point O will give uniform " illumination " over the inner surface of the sphere. A directional radiator will concentrate the energy into a beam which will illuminate only a portion of the sphere as shown shaded.

The term *gain* of any particular aerial system always applies in the direction of maximum radiation.

The gain of an aerial is usually expressed as a power ratio, either as a multiple of so many " times " or in decibel units. For example, a power gain of 20 times could be represented as 13db (i.e. 10 $\log_{10}$ 20).

The truly isotropic radiator is a purely theoretical concept, and in practice the gain of beam aerials is usually compared with the radiation from a single half-wave dipole fed with an equal amount of power. The radiation pattern of even a single half-wave dipole is markedly non-uniform, and in consequence the power gain of such an aerial compared with the hypothetical isotropic radiator is about 64 per cent (i.e. 5/3 times or 2·15db), but since the half-wave dipole is the simplest practical form of radiator it is generally acceptable as a basis of comparison.

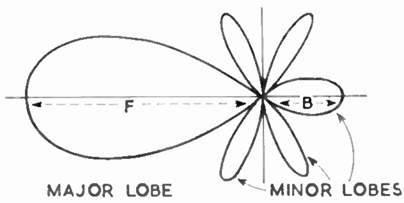
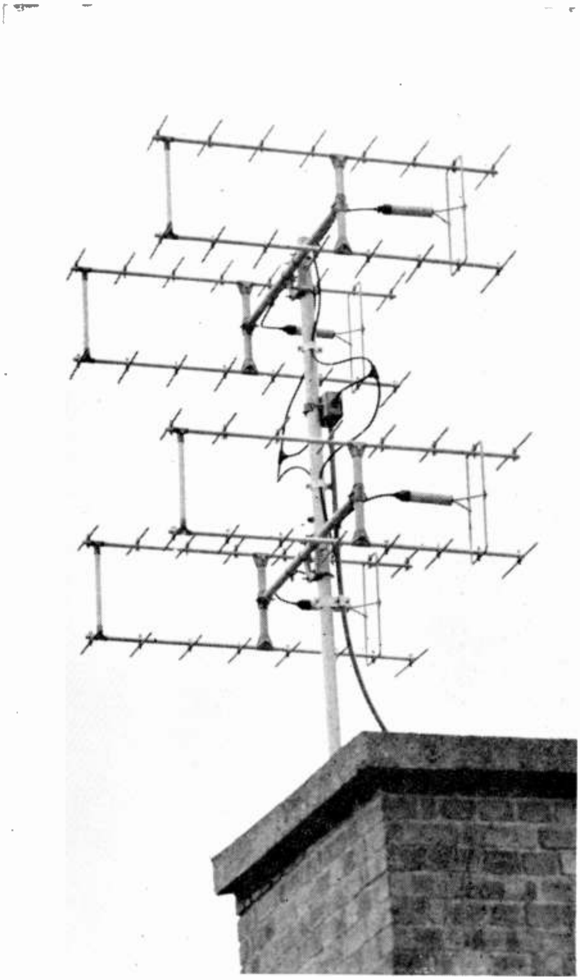


Fig. 14.2. Polar diagram of a v.h.f. beam aerial. The front-to-back ratio is represented by F/B.

The area of " illumination " is not sharply defined as shown by the shaded region in Fig. 14.1 but falls away gradually from the centre of the area. The boundaries of the illuminated area are determined by joining together all points where the radiation intensity has fallen by half (i.e. 3db): these are known as the *half-power points*. The gain of the aerial can then be determined by dividing the total surface area of the sphere by the illuminated area: e.g. if the total surface area were 100 sq. cm. and the illuminated area bounded by the half-power points were 20 sq. cm., the gain of the aerial would be five times or 7db. The radiation in any particular plane can be plotted graphically, usually in polar co-ordinates; such a plot is called a *polar diagram*. A typical polar diagram is shown in Fig. 14.2. The region of maximum radiation is called the *major lobe*. Unwanted radiation is occurring in other directions, and these regions, when small compared with the major lobe, are called *minor lobes*. All practical aerials exhibit such lobes and the aerial designer frequently has to compromise to obtain the optimum performance for any particular application. For example, an



The J-Beam aerial array located above the factory of M-O Valve Co. Ltd., Hammersmith, for the beacon station GB3GEC.
(Photo by courtesy of M-O Valve Co. Ltd.)

aerial may be designed for maximum *front-to-back ratio*; i.e. for minimum radiation in the direction opposite to the major lobe. To achieve such a condition it may be necessary to sacrifice some gain in the major lobe (or forward radiation) with a possible increase in other minor lobes (or side lobes) and thus the designer will need to consider all the implications before finalizing any particular design.

In practice the radiation from an aerial is measured in a horizontal and a vertical plane. The *beamwidth* is the angle between the two half-power points in the plane under consideration. The vertical polar diagram is greatly influenced by the height of the aerial above ground; the higher the aerial the lower will be the angle of maximum radiation, and at the same time the effects of neighbouring objects such as houses and trees will be minimized. The important requirement is to place the aerial well clear of such objects and this frequently means as high as can be safely achieved. Any aerial which has the property of concentrating radiation into any particular direction is said to possess *directivity*.

Bandwidth

The performance of an aerial array generally depends upon the resonant properties of tuned radiators such as dipoles or other elements, and therefore any statement regarding its power gain or beamwidth will be valid only over a restricted frequency band. Beyond the limits of this band the properties of an aerial system may be entirely different. Hence it is useful to define *bandwidth* as that range of frequencies over which the power gain of the aerial array does not fall by more than a certain percentage as compared with the frequency at which maximum gain is obtained (e.g. a bandwidth of 15 Mc/s for a 50 per cent reduction in power gain). Alternatively the bandwidth may be defined as the frequency band over which the standing wave ratio of the aerial feeder does not exceed a prescribed limit (e.g. bandwidth of 10 Mc/s for a standing wave ratio not exceeding 2 : 1). The latter convention is the one generally used. It should be noted that these examples are not related to one another.

Capture Area or Aperture

Besides examining the action of a transmitting aerial array in concentrating the radiated power into a beam it is also helpful to examine the way in which the same aerial structure will effect the reception of an incoming signal. In this study it is convenient to introduce the concept of *capture area* or *aperture* of the aerial. This concept is frequently misunderstood, probably because it may appear to relate to the cross-sectional area of the radiated beam (as represented by A in Fig. 14.1): it is in fact related to the *inverse* of the cross-sectional area of the beam inasmuch as an aerial which has a high gain usually has a sharply focused beam (i.e. one of small cross-sectional area) but at the same time the capture area of the aerial is large. The larger the capture area, the more effective is the aerial.

The actual size of the aerial system does not always give a reliable indication of the capture area. A high-gain array may have a capture area considerably greater than its frontal area determined by its physical dimensions.

The fundamental relationship between the capture area and the power gain of an aerial system is—

$$A = \frac{G_I.\lambda^2}{4\pi}$$

where A is the capture area and λ is the wavelength (measured in the same units as A) and G_I is the power gain (arithmetically not db) relative to an isotropic radiator.

A half-wave dipole has a gain of $\frac{5}{3}$ relative to an isotropic radiator, and therefore this formula can be modified to give the capture area in terms of the gain of a half-wave dipole G_D instead of G_I simply by introducing the factor 3/2, thus—

$$A = \frac{5}{3} \times \frac{G_D.\lambda^2}{4\pi} = \frac{5G_D\lambda^2}{12}$$

This formula shows that if the wavelength is kept constant the capture area of an aerial is proportional to its gain, and therefore if an increase in gain results in a narrower beamwidth it must follow that a narrower beamwidth corresponds to a greater capture area (the term beamwidth being used here to signify both horizontal and vertical dimensions, i.e. in effect the cross-sectional area).

The formula also shows that for any given power gain the capture area is proportional to the square of the wavelength. For example, an aerial having a power gain of, say, 10 times relative to a dipole at 600 Mc/s (0·5m) would have a capture area one-sixteenth of that of an aerial having a similar power gain at 150 Mc/s (2m); to achieve equal capture area the gain of the 600 Mc/s aerial would thus have to be 16 times greater than that of the 150 Mc/s signal, i.e. 160 times relative to a dipole. This is unfortunate because it is the capture area of the aerial that determines its effectiveness in absorbing the incoming radiation: it means that as the wavelength is reduced it becomes increasingly important to design the aerial to have a higher gain in order to produce the same voltage at the receiver terminals and thus the same basic signal-to-noise ratio.

A very useful rule of thumb method of calculating the gain of a multiple array aerial system is that each time the aperture is doubled the gain is increased by approximately 3db, although in practice, the increase is usually a little less. For example if two 5 element Yagi arrays are stacked, provided they are spaced so that their apertures just touch, then the overall gain will be increased from about 8 to 11db. It is useful to remember this point when considering *stacking* later in the chapter.

These observations apply only to signals being received or transmitted in the direction of maximum gain. For directions other than the optimum the relationships become more complex.

Multi-radiator Arrays

High-gain aerial arrays can be built up from a number of individual radiators such as half-wave dipoles. To achieve the maximum gain, the spacing of these radiators should be such that their respective capture areas just touch. Where the individual radiators are themselves high-gain systems, such as Yagi arrays, the centre-to-centre spacing of each radiating system needs to be larger, since the individual capture areas are greater.

Reciprocity Theorem

The theorem of reciprocity states that any particular aerial gives the same performance either as a transmitting or as a receiving system. Practical aerial designs are therefore worked out in terms of transmission because the characteristics are more easily determined in this way, and the resulting aerials are assumed to have similar reception properties.

Angle of Radiation

The characteristics of propagation in the v.h.f. ranges are determined principally by the influence of the troposphere, i.e. the part of the atmosphere extending from ground level up to a few thousand feet. There is rarely ionospheric propagation on frequencies higher than about 100 Mc/s and thus any energy which is radiated at more than a few degrees above the horizontal is wasted. Similarly to transmit to a particular point it is unnecessary to radiate a broad beam in the horizontal plane. Generally, therefore, the aerial designer tries firstly to reduce the vertical beamwidth to avoid wastage of power into space and secondly to reduce the horizontal beamwidth according to the required ground coverage. A narrow horizontal beamwidth can however be a disadvantage because stations situated off the beam may be missed, when searching. Under conditions of high activity, on the other hand, a narrow bandwidth can be helpful in rejecting interference from unwanted stations in other directions. A compromise between gain and beamwidth must be made.

Height of Aerial above Ground

Several considerations apply when deciding the height of the aerial above the ground for optimum performance:

(a) the aerial should be well above local screening from buildings and other obstacles;

(b) even if there is no screening, the earth will act as a plane reflector and will tend to direct the radiated energy into space rather than along the horizontal as is required at v.h.f. To minimise this effect, the aerial should be as high as possible (and the height should be judged in terms of wavelengths rather than feet);

(c) the higher the aerial the greater the feeder length and the more difficult and costly is its erection, especially in confined spaces. Thus, again, a compromise has to be made.

It is well to remember that even in an ideal location from the v.h.f. point of view, such as a hilltop with no local screening, it is still necessary to erect the aerial as high as can be achieved within the resources available. This consideration is also very important when aerials are stacked as the effective height of the aerial is the height of the physical centre or electrical centre of gravity above ground. In general, the improvement to be obtained from increasing aerial height outweighs the increase in feeder loss. At an average amateur location the height/gain relationship is often better than 6db for doubling of height. Only in the case of well-sited stations on high ground is such a rate of height gain not achieved.

Polarization

Radio waves are constituted from electric and magnetic fields mutually at right angles and also at right angles to the direction of propagation. The ratio of the electric component E to the magnetic component H in free space ($E/H = Z$) is known as the *impedance of free space* and has a value of about 377 ohms. When the electric component is horizontal, the wave is said to be *horizontally polarized*. Such a wave is radiated from a horizontal dipole. If the electric component is vertical, as in a vertical dipole, the wave is said to be *vertically polarized*.

Sometimes the polarization is not exclusively horizontal or vertical and the radiation is then said to be *elliptically polarized* or, in the special case where the horizontal and vertical components are equal, *circularly polarized*. The effect of the addition of two components of the same kind (i.e. electric *or* magnetic) at right angles is to create a rotating field, the direction of rotation of which depends on the relative phase of the two components. Thus the polarization of the wave will appear to have either clockwise or counter-clockwise rotation, a feature which is important in the use of helical aerials. A dipole will receive an equal pick-up from a circularly polarized wave irrespective of whether it is mounted horizontally, vertically or in an intermediate position, but the strength will be 3db less than if an aerial designed for circular polarisation is used. This means that the full *gain* of a helix will only be realised when received by a helix. Horizontally or vertically polarized waves (or any other waves from linear sources) are known as *plane polarized* waves.

It has been found by some experimenters that in the v.h.f. range horizontally polarized waves suffer less attenuation over long distances than vertically polarized waves, and this system is therefore often preferred. It has in fact been universally adopted for amateur communication in the British Isles and many other countries. Vertically polarized waves may be more suitable for special purposes such as short-distance or mobile communication and a simple ground-plane or similar aerial can then be used, as described in Chapter 16 (*Mobile Equipment*).

Tests have been conducted at frequencies of the order of 70 Mc/s between fixed stations employing horizontally polarized beam aerials and mobile stations employing vertically polarized ground plane aerials which show that the loss in signal strength due to cross polarization is much less than might be expected due to the multiple reflections that occur from the objects that surround mobile radiators. The vertically polarized wave is transformed into one of random polarization and hence there is no difference in performance if either a horizontally or vertically polarized aerial is used at the fixed station. Further work needs to be done in this field in the u.h.f. band to see whether the same results occur although tests by the BBC on television Bands IV and V (470–582 Mc/s and 606–960 Mc/s) show less multi-path propagation from horizontally polarized aerials and hence less fading.

AERIAL FEEDERS

Before discussing v.h.f. aerial design it will be helpful to review the methods of conveying power from the transmitter to the aerial. The feeder length should always be considered in terms of wavelengths rather than the actual length of the conductors. If the feeder length is short compared with the wavelength, the loss caused by its ohmic resistance and by the dielectric conductance is unimportant as also is the effect of incorrect impedance matching. For v.h.f. operation, however, the aerial feeder is usually many wavelengths long, and therefore both the loss introduced and the matching of the load to the feeder are of the utmost importance.

Two types of feeder, or transmission line, are in common use, the *unbalanced* or *coaxial feeder* and the *balanced pair:* the latter may be either of open construction or enclosed in polythene ribbon or tubular moulding. Each type has its own particular advantages and disadvantages.

In a coaxial cable, the radio frequency fields are contained entirely within the outer conductor and hence there should be no r.f. currents on the outside. This enables the cable to be carried in close proximity to other cables and metal objects without interaction or serious change of its cable properties which might cause reflections and thereby introduce appreciable loss. There is no loss by external radiation.

The open-wire or balanced feeder has a radiation-loss which is dependent upon the ratio of the spacing of the wires to the wavelength and becomes more serious as the frequency is raised. However, if this spacing is less than 0·01λ the radiation loss is negligible. The properties can also be severely changed by the close proximity of metal objects and the accumulation of ice or water on the separating insulation, and therefore much greater care must be taken in the routing of the feeder. Difficulties are often experienced when attempting to use this type of feeder with rotatable aerial

TABLE 14.1

Characteristics of Typical Radio Frequency Feeder Cables

Type of Cable	Nominal Impedance Z_0 (ohms)	Dimensions (in.) Centre Conductor	Over Outer Sheath	Over Twin Cores	Velocity Factor	Approximate Attenuation (db per 100 ft.) 70 Mc/s	145 Mc/s	433 Mc/s	1296 Mc/s	Power Handling (see footnote)	Remarks
Standard-TV feeder	75	7/·0076	0·202	—	0·67	3·5	5·1	9·3	17	—	—
Low-loss TV feeder (semi-air-spaced) Super Aeraxial	75	0·048	0·290	—	0·86 approx.	2·0	3·0	5·5	10	1	Semi-air-spaced or cellular.
Flat Twin	150	7/·012	—	0·18 × 0·09	0·71	2·1	3·1	5·7 *	11 *	—	*Theoretical figures, likely to be considerably worsened by radiation.
Flat twin	300	7/·012	—	0·405 × 0·09	0·85	1·2	1·8	3·4 *	6·6 *	—	
Tubular twin . .	300	7/·012	—	0·446	0·85	1·2	1·8	3·4 *	6·6 *	—	
Uniradio 1 . .	71	0·056	0·45	—	0·67	1·4	2·2	4·6	8·6	2	Replaced by UR57.
Uniradio 4 . .	46	7/·032	0·405	—	0·67	1·9	2·9	5·6	10·6	2	Replaced by UR67.
Uniradio 6 . .	100	0·036	0·365	—	0·86	1·4	2·2	4·6	8·6	1	Semi-air-spaced.
Uniradio 32 . .	71	0·022	0·202	—	0·67	3·5	5	10	18	—	—
Uniradio 57 . .	75	0·044	0·405	—	0·67	1·5	2·3	4·5	10	2	—
Uniradio 67 . .	50	7/·029	0·405	—	0·67	1·7	2·4	4·6	10	2	RG8A/U similar.
Uniradio 74 . .	51	0·188	0·870	—	0·67	0·8	1·2	2·4	5	2	—
Uniradio 77 . .	75	0·104	0·870	—	0·67	0·8	1·2	2·4	5	2	—
RG-11A/U. . . .	75	7/·016	0·405	—	0·67	1·9	2·8	5·5	10·5	2	—

This table contains information supplied by Aerialite Ltd., and British Insulated Callender's Cables Ltd. The power handling capacity of any cable falls with rising frequency. Cables suitable for handling 100 watts at 433 Mc/s or 1296 Mc/s have the figures 1 or 2 allocated in the Power Handling column respectively (at low S.W.R.).

arrays. It is, for instance, quite unsatisfactory to bind a ribbon feeder directly against a metal mast. When the feeder is kept well clear of metal objects, however, the loss tends to be less than that of coaxial cable unless the frequency is so high that the radiation-loss is serious (i.e. 450 Mc/s and above).

Generally speaking, the use of open-wire feeders is restricted to bands of 144 Mc/s and below; coaxial cable is used for all frequencies up to about 3000 Mc/s. Above about 1000 Mc/s the loss in conventional types of flexible coaxial cable becomes prohibitive and rigid semi-air-spaced types of cable are then used. Above 3000 Mc/s waveguides are usually necessary for feeder runs of more than a few inches.

A list of typical aerial feeder cables obtainable in Great Britain is shown in **Table 14.1**. Further information can be obtained from the various manufacturers.

In the semi-air-spaced type of coaxial cable, the centre conductor is supported either on beads or on a helical thread: in some forms a continuous filling of cellular polythene is used as the insulator. This type of cable has a lower capacity per unit length than the solid type and hence the *velocity factor* (i.e. the ratio of the wave velocity in the cable compared with that in free space) is higher, being about 0·88–0·98 for helical and bead types and about 0·8 for cellular polythene types compared with 0·66 for solid cable. The use of semi-air-spaced centre conductors allows cables to be designed with less attenuation for a given size, or with the same attenuation for a smaller size, but unfortunately the bead and helical types suffer from the disadvantage that moisture can easily enter the cable; special precautions must therefore be taken to ensure a good watertight seal at the aerial end if such cable is used. Suitable material for this purpose is Telecompound* or Bostik sealing strip.

Cellular polythene cables do, however, suffer from another disadvantage: the effective dielectric constant of the cellular polythene varies according to the number and volume of the air cells per unit length which unfortunately cannot be controlled with precision during manufacture. Thus the characteristic impedance may tend to vary along a length of feeder and although this may not matter at the lower end of the v.h.f. spectrum such cable is not recommended for frequencies above about 500 Mc/s.

It is also important to exclude water even from the outer braiding of any coaxial cable. If water has once entered it is very difficult to dry it out, and the loss in the feeder becomes progressively higher as the copper braiding corrodes.

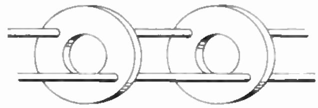
Open-wire Feeders

To obtain very low losses open-wire feeder line can conveniently be made from hard-drawn 16 s.w.g. copper wire with separating insulators placed at intervals. The insulators should be made from polythene and be shaped in the form of a disc with the centre removed, as shown in **Fig. 14.3**. This ensures that in the places where the maximum electric stress occurs, i.e. between the conductors, the dielectric is air and not the solid insulating material, thus minimizing losses. To prevent excessive radiation-loss the characteristic impedance should not exceed about 300 ohms. Care must be taken to avoid sharp bends and also the close proximity of surrounding objects as stated earlier. The characteristic impedance of an open-twin line is given by—

$$Z_0 = 276 \log_{10} (D/d)$$

where D is the centre-to-centre spacing and d is the diameter of the wire (measured in the same units). A chart of characteristic impedance in relation to the conductor size and spacing is given in Chapter 13 (H.F. Aerials).

Fig. 14.3. Open-wire line with disc insulators. The centre portions of the insulators are removed to minimize dielectric losses.



Metal Insulators

A quarter-wave short-circuited transmission line presents a very high impedance at the open end and hence may be connected across an open-wire line without affecting the power flow in any way; such a device is called a *quarter-wave stub*. The stub so formed may conveniently be used as a support or termination for an open-wire line as shown in **Fig. 14.4**. Since the stub must be resonant in order to behave as an insulator it can function only over a narrow range of frequency. Note that the characteristic impedance of the stub does not have to be the same as that of the line.

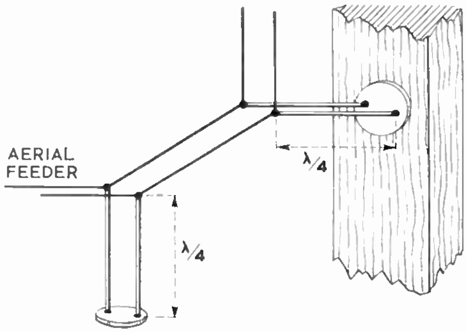


Fig. 14.4. Quarter-wave closed stubs are often used for supporting open-wire feeders. When used in this way they are known as " metal insulators." The inductance of the short-circuiting conductors is minimized by making them in the form of large metal plates.

Surface-wave Transmission Lines

A type of feeder which becomes useful above about 400 Mc/s is the surface-wave transmission line. The wave is directed on to a single conductor by means of a horn: see **Fig. 14.5**. The dimensions of the horn are not critical but the angle should be correct, and for the best performance the side should be several wavelengths long. The single conductor wire should be covered with a thin dielectric, preferably polythene, although enamel or even an oxidized covering is usually sufficient. This covering layer minimizes radiation-loss by reducing the effective diameter of the field surrounding the wire. Typical losses measured at a frequency of 3300 Mc/s (9cm) using horns 21 in. long and 13 in. in diameter and a No. 14 s.w.g. wire feeder (bare but oxidized) are about 1·35db per hundred feet plus about 0·4db per horn, making a total of just over 2db per hundred feet.

At 430 Mc/s the attenuation of the field 8 in. away from the feeder is about 20db. The feeder can therefore be run

*A softened polythene compound manufactured by *Telcon Plastics Division of the BICC Group.*

reasonably near to walls and other objects for short distances. If the feeder is a fairly long one, however, it would be desirable to maintain a spacing of several feet over most of its length. The radius of any bends should be kept greater than about one wavelength in order to avoid losses due to the tendency of the energy to be radiated into space whenever there is a sudden change in the direction of the feeder.

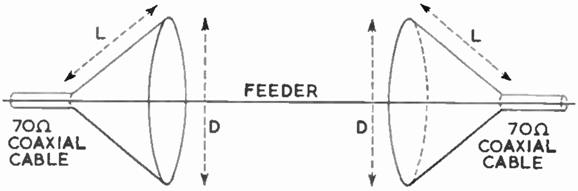


Fig. 14.5. Surface-wave transmission line. The outer sheathing of each length of coaxial cable is terminated by an open cone. The feeder itself should be of 10-16 s.w.g. copper wire, preferably enamelled; its length may be several hundred feet. The cone diameter D must be 0·6 L, but L itself may be of any length greater than 3λ.

Balance-to-Unbalance (Balun) Transformers

In most cases the aerial requires a balanced feed with respect to ground, and it is therefore necessary to use a

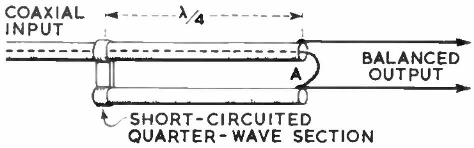


Fig. 14.6. Quarter-wave open balun or Pawsey stub.

device which converts the unbalanced output of a coaxial cable to the balanced output required by the aerial. This device also prevents the wave which has been contained within the cable from tending to " spill over " the extreme end and travel back over the surface of the cable. Whenever this occurs there are two important undesired effects: first the re-radiated wave modifies the polar diagram of the attached aerial, and secondly the outer surface of the cable is bound to have a radio frequency voltage on it.

To prevent this, a balance-to-unbalance transformer (abbreviated to *balun*) is connected between the feeder cable

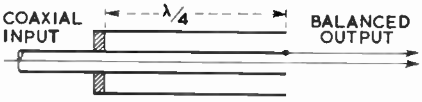


Fig. 14.7 (a). Coaxial-sleeve balun.

and the aerial. The most simple balun consists of a short-circuited quarter-wave section of transmission line attached to the outer-braiding of the cable, as shown in **Fig. 14.6**. This is often known as a *Pawsey stub*. At the point *A*, the quarter-wave section presents a very high impedance which prevents the wave from travelling over the surface. Note that the $\frac{\lambda}{4}$ dimension given is in air and not in the cable.

Several modifications to the simple balun are possible: for example, the single quarter-wave element may be replaced by a quarter-wave coaxial sleeve, thus reducing radiation-loss: see **Fig. 14.7** (a). To prevent the ingress of water and to improve the mechanical arrangement, the centre conductor may itself be connected to a short-circuited quarter-wave

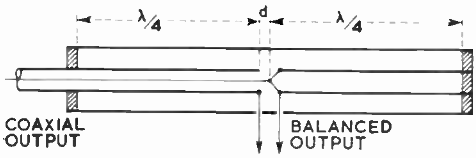


Fig. 14.7 (b). Totally enclosed-coaxial balun. The right-hand section acts as a " metal insulator."

line acting as a "metallic insulator" as shown in **Fig. 14.7 (b)**. The distance *d* should be kept small, and yet the capacity between the sections should also be kept small otherwise the quarter-wave section will not be resonant at the desired frequency. A satisfactory compromise is to taper the end of the quarter-wave line, although this is by no means essential. In practice, at a frequency of 435 Mc/s about ⅛ in. is a suitable spacing. The whole balun is totally enclosed, the output being taken through two insulators mounted in the wall.

A useful variation is that shown in **Fig. 14.8**, which gives a 4 : 1 step-up of impedance. The half-wave loop is usually made from flexible coaxial cable, and allowance must therefore be made for the velocity factor of the cable when calculating a half-wavelength.

At frequencies above about 2000 Mc/s it may be inconvenient to mount the coaxial-sleeve balun close to a dipole radiator. In this case the sleeve can be mounted a short distance back from the end of the line. The characteristic impedance of the balun element is not critical.

A Pawsey stub may be constructed by attaching a piece of

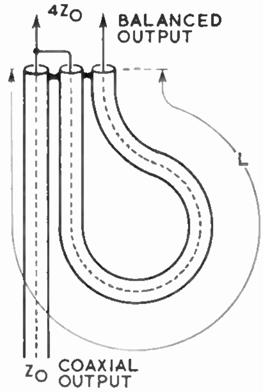


Fig. 14.8. Coaxial balun giving a 4 : 1 impedance step-up. The length *L* should be λ/2, allowing for the velocity factor of the cable. The outer braiding may be joined at the points indicated.

coaxial cable one physical quarter-wave long (the centre conductor being unused) to the braiding of the feeder cable. The two sections should be spaced sufficiently to ensure an air dielectric between them. If the two pieces lie closely alongside one another the resonant length must be reduced and an inferior dielectric introduced. It is important to note that since it is the electrical characteristics of the outer surface that are being used, there is no need to allow for the velocity factor of the cable. Coaxial-sleeve baluns should have an outer-to-inner diameter ratio of between 2 : 1 and 4 : 1.

The performance of these baluns will be frequency dependant as they rely for their operation upon the properties of resonant transmission lines. In all cases, however, these lines are effectively terminated in a low impedance and the bandwidth is comparatively large. All are quite suitable for use in the v.h.f./u.h.f. amateur bands.

IMPEDANCE MATCHING

For an aerial feeder to deliver power to the aerial with minimum loss, it is necessary for the load to behave as a pure resistance equal in value to the characteristic impedance of the line. Under these conditions no energy is reflected from the point where the feeder is joined to the aerial, and in consequence no standing waves appear on the line.

When the correct terminating resistance is connected to any feeder, the voltage and current distribution along the line will be uniform. This may be checked by using a device to explore either the magnetic field (H) or the electric field (E) along the line. One such device, suitable for use with a coaxial feeder, is a section of coaxial line with a longitudinal slot cut in the wall parallel to the line. A movable probe connected to a crystal voltmeter is inserted through the slotted wall. This samples the electric field at any point, and the standing wave ratio may be determined by moving the probe along the line and noting the maximum and minimum readings. The distance between adjacent maxima or between adjacent minima is one half-wavelength.

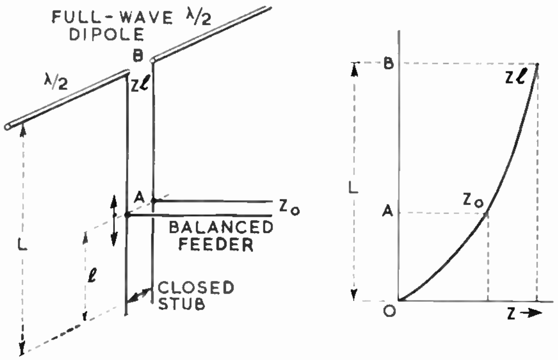


Fig. 14.9. Stub matching applied to a full-wave dipole.

The fields surrounding an open line may be explored by means of an r.f. voltmeter, but it is much more difficult to obtain precise readings than with a coaxial line because of hand-proximity effects and similar disturbances.

Another device which measures forward and reflected waves is the *reflectometer* (see page 14.21).

The term *matching* is used to describe the procedure of suitably modifying the effective load impedance to make it behave as a resistance and to ensure that this resistance has a value equal to the characteristic impedance of the feeder used. To make a *complex* load (i.e. a load possessing both resistance and reactance) behave as a resistance, it is necessary to introduce across the load a reactance of equal value and opposite sign to that of the load, i.e. the reactance is "tuned out." A convenient device which can theoretically give reactance values from minus infinity to plus infinity, (i.e. large capacitance to large inductance) is a section of transmission line either of length variable between zero and one half-wavelength having an open-circuited end or alternatively of length a little greater than one half-wavelength having a

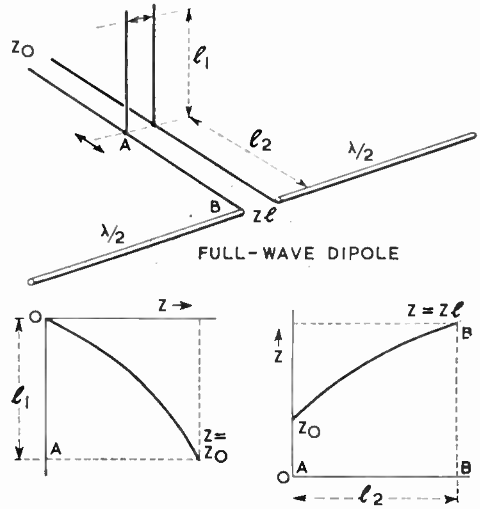


Fig. 14.10. Stub matching with a movable short-circuited stub.

movable short-circuit capable of being adjusted over a full half-wavelength.

Although there is no need to make the characteristic impedance of a stub equal to that of the transmission line, it may be desirable to do so for practical reasons.

In addition to tuning out the reactance, a match still has to be made to the transmission line. The impedance at any point along the length of a quarter-wave resonant stub varies from zero at the short-circuit to a very high impedance at the open end. If a load is connected to the open end and the power is fed into the stub at some point along its length the stub may be used as an auto-transformer to give various values of impedance, according to the position of the feed point. This is shown in **Fig. 14.9.** The distance L is adjusted to tune the aerial to resonance and will be one quarter-wave long if the aerial is already resonant. The distance l is adjusted to obtain a match to the line. However, it is usually more convenient to have a stub with an adjustable short-circuit which can slide along the transmission line: see **Fig. 14.10.**

In practice matching can be achieved entirely by the "cut and try" method of adjusting the stub length and position until no standing waves can be detected. The feeder line is then said to be *flat*. However, the frequency range over which any single-stub matching device is effective is quite small, and where wideband matching is required some other matching system must be used.

Stub-matching

When it is possible to measure the s.w.r. with good accuracy and to locate the exact position on the feeder at which the minimum current occurs, the length of the stub required to match the line and the correct point at which to connect it may be predicted and much of the time required by cut-and-try methods saved. The predictions may be read off from **Fig. 14.11.** The application of the method is best suited to balanced feeders; adjustments are somewhat easier with open-wire lines.

The measurement of s.w.r. may be made when transmitting on low power:

(a) by inserting in series with the feeder an s.w.r. indicating or measuring device, near the point at which the stub is to be connected. For wide bandwidth this stub should be very near the aerial.

(b) by running along the feeder a calibrated probe which has arrangements to keep its coupling to the feeder constant while its position is moved along the line. It is helpful to have the feeder well tensioned. Either voltage- or current-probes can be used having a thermocouple or crystal diode connected to a sensitive milliammeter. The probe must be *loosely* coupled to the line.

The position of a minimum is sharply defined, a maximum is flat. Even so, a minimum position is best found by "bracketing" it. The positions at which equal readings of current or voltage are obtained on either side of the minimum are found: the minimum position is taken to be midway between them. A voltage minimum can be assumed to be a quarter wave distant from a current minimum. With the bracketing method even a torch bulb coupled to the line can be used to locate the I_{min} position with fair accuracy. Its coupling to the line must be the same in each position. The two positions at which the filament just begins to glow are noted.

It is very important that harmonic currents are not flowing in the line during measurements, and as filters do not work well unmatched, the probe should be sharply tuned to the operating frequency. A "field strength meter" with suitable constant-coupling arrangements can be used as a probe. It may be calibrated by connecting a short circuit on the line to set up an infinite s.w.r. The theoretical sinusoidal current pattern along the line is used to set the probe in positions of known current (i.e. current of known ratio relative to the maximum current).

e.g. (a) Set the probe at the I_{max} position and reduce the probe coupling and adjust transmitter power, or transmitter loading to get a full scale meter deflection.

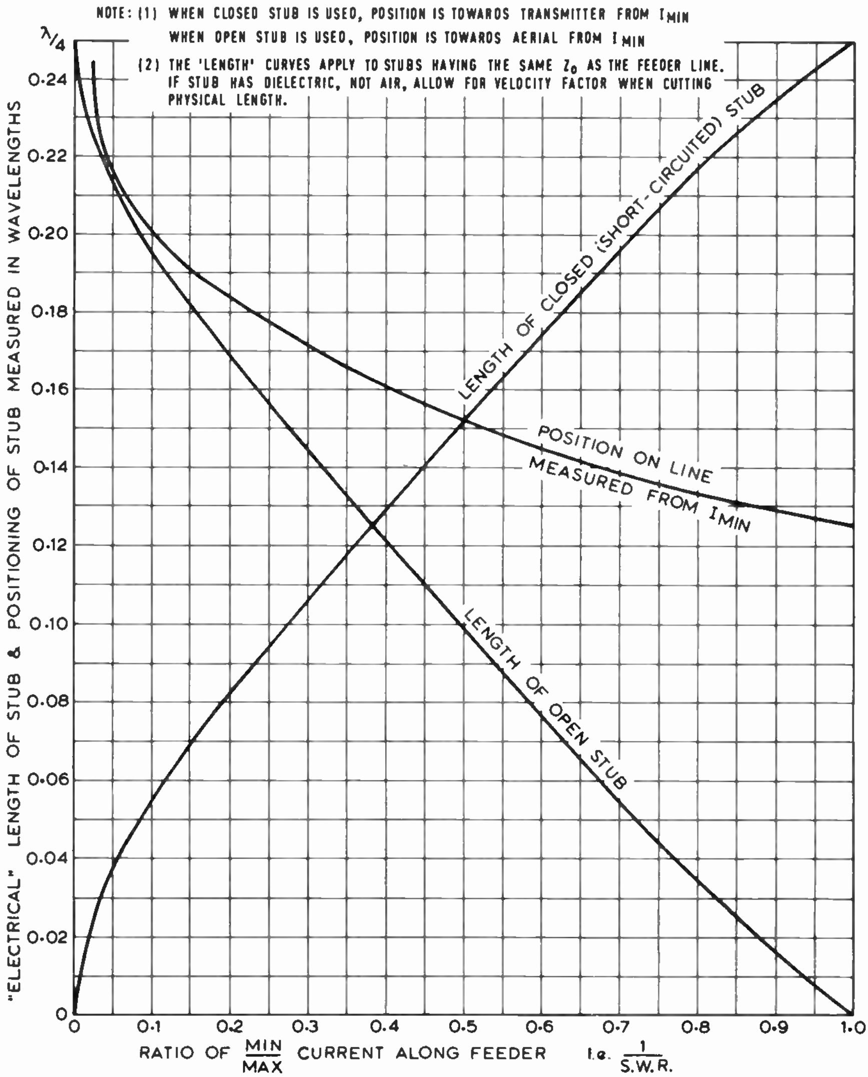


Fig. 14.11. Chart giving details of stub lengths.

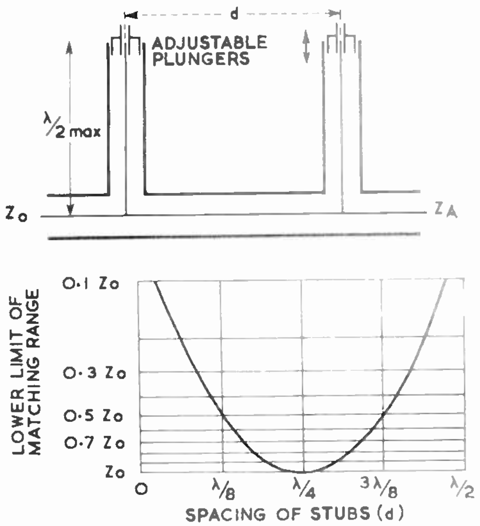


Fig. 14.12. Two-stub coaxial tuner. The graph shows the lower limit of the matching range: the upper limit is determined by the Q of the stubs (i.e. it is dependent on the losses in the stubs). Z_0 is the characteristic impedance of the feeder.

(b) Locate the I_{min} position accurately by bracketing.

(c) Set the probe $\lambda/12$ from *Imin*. Note the meter reading. The actual current here is 0·5 I_{max} ($\lambda/12 = 30°$ phase. Sin $30° = 0·5$).

(d) Set the probe $\lambda/8$ from I_{min}. Here the actual current is 0·707 I_{max}.

(e) Set the probe at other positions and calibrate similarly.

Any error due to a slightly incorrect position of I_{min} may be eliminated by repeating the readings on the opposite side of I_{min} and taking an average meter reading. Any appreciable asymmetry may indicate excessive probe coupling, variation in probe coupling, or harmonic currents.

If the predicted stub does not produce a perfect match immediately, note whether the point of attachment of the stub is now a position of I_{min} or I_{max} of the new standing wave on the "matched" line. If neither, then the stub is not tuning the aerial system to resonance. The first adjustment is therefore made to the length of the stub so that the point of attachment is resistive. Correct adjustment is when the position of I_{min} or I_{max} is coincident with the point of attachment. If this is now a point of I_{max}, the resistance is too low for a perfect match. If a closed stub is being used, lengthen it and simultaneously move its point of attachment nearer the aerial by the same amount. (These adjustments are equal only when the stub Z_0 is the same as that of the feeder.) If the correct position is passed, the "matched" line will now have an I_{min} at the point of attachment, and the adjustment is to shorten the closed stub and simultaneously move it further from the aerial by the same amount. The total distance from the short circuit of the stub to the aerial remains the same to keep the system resonant. For an open-circuit stub the adjustments are in the opposite sense.

Stub Tuners

On a coaxial line it is impracticable to construct a stub having an adjustable position. However, two fixed stubs spaced by a certain fraction of a wavelength can be used for matching purposes: see **Fig. 14.12.** The spacing usually employed is $\lambda/8$ or odd multiples thereof. With this spacing independent adjustment of the short-circuit plungers gives a matching range from 0·5 times the characteristic impedance of the transmission line (Z_0) upwards. As the spacing is increased towards $\lambda/2$ or decreased towards zero, the matching range increases, but the adjustments then become

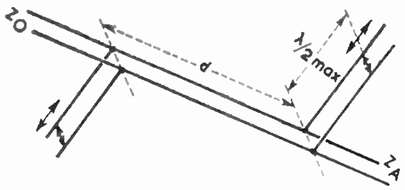


Fig. 14.13. Two-stub open-wire tuner. With an open-wire line stubs should be mounted on opposite sides of the line as shown so as to avoid mutual coupling. The matching range can be seen from the graph in Fig. 14.12.

extremely critical and the bandwidth very narrow. The theoretical limit of matching range cannot be achieved owing to the resistance of the conductors and the dielectric loss; i.e. the Q is limited. To obtain the highest Q the ratio of outer-to-inner conductor diameters should be in the range 2 : 1 to 4 : 1 (as for coaxial baluns). An important mechanical detail is the provision of reliable short-circuiting plungers which will have negligible inductance and also ensure low-resistance contact. They can be constructed of short lengths of thin-walled brass tubing, their diameters being chosen so that when they are slotted and sprung they make a smooth sliding contact with both inner and outer conductors.

The two-stub tuner may be applied to open transmission lines if it is inconvenient to have a movable stub. In this case the stubs must be mounted laterally opposite to each other to prevent mutual coupling: see **Fig. 14.13.**

This type of tuner may, of course, be used for other purposes than to feed an aerial. For example, it will serve to match an aerial feeder into a receiver, or a dummy load to a transmitter. A greater matching range can be obtained by using a three-stub tuner, the stubs being spaced at intervals of one quarter-wavelength apart, as shown in **Fig. 14.14.** The first and third stubs are usually ganged together to avoid

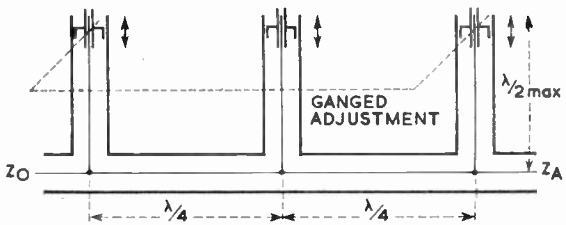


Fig. 14.14. Three-stub tuner. This provides a greater matching range than a two-stub tuner. Z_0 is the characteristic impedance of the feeder.

the long and tedious matching operation which becomes necessary when adjustments are made to three infinitely variable stubs.

Quarter-wave Lines

An impedance transformation can be effected by using a certain length of transmission line of a different characteristic impedance from the feeder. This may be used to match a load to a transmission line. A special condition occurs when the length of the section of line is an odd number of quarter-wavelengths and the following formula then applies:

$$Z_t = \sqrt{Z_0 \, . \, Z_l}$$

where Z_t is the characteristic impedance of the section of quarter-wave line and Z_0 and Z_l are the feeder and load impedance respectively. For example, if Z_0 is 80 ohms and Z_l is 600 ohms:

$$Z_t = \sqrt{80 \times 600} = 251 \text{ ohms}$$

This matching section is useful for transforming impedance and is called a *quarter-wave transformer:* see Fig. 14.15. A wider bandwidth match can be obtained by reducing the transformation ratio and using a number of quarter-wave sections of progressively changing impedance in cascade.

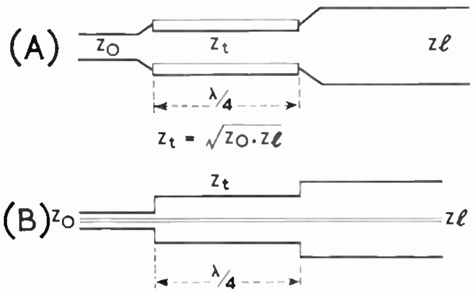


Fig. 14.15. Quarter-wave transformers. (A) shows a construction suitable for open-wire lines, and (B) is the corresponding method for coaxial cables. Where a solid-dielectric section is used, due allowance must be made for the velocity factor.

It is often convenient to incorporate a quarter-wave transformer into a coaxial sleeve balun of the type shown in Fig. 14.7 (a). The length of the quarter-wave section will only be a physical quarter-wavelength if the inner coaxial transformer is air spaced but as a quarter-wave transformer usually has an impedance different from that of standard coaxial cable it has to be fabricated anyway and thus this is no disadvantage. If a transformer is built into the balun shown in Fig. 14.7 (b), the characteristic impedance of the right hand quarter-wave section need not be changed since it is acting as a " metal insulator."

A section of tapered line can also be used to effect an impedance transformation, and an application of the principle is described later in this chapter. Again a quarter-wavelength section is only a special case, and to achieve a match in a particular installation the line lengths and the angle of taper should be varied until a perfect match is achieved. This form of matching device is often called a *delta match.*

Multiple Feed Arrangements

It is often required to feed two or more aerial arrays, such as stacked Yagis, in phase and there are a number of convenient ways of doing this using standard impedance coaxial cables. Details of one possible arrangement is shown in Fig. 14.16. Others can be devised using similar principles.

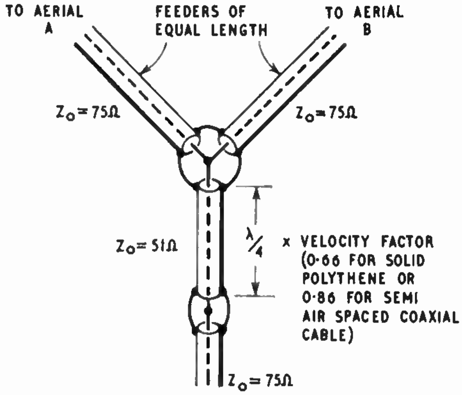


Fig. 14.16. Arrangements for feeding two identical aerial arrays in phase using standard coaxial cables. Links between the braiding of the cables should be as short and of as low an impedance as possible. Braid can be used for this purpose.

AERIAL ARRAYS

The basic elements of v.h.f. aerial arrays are usually half-wave or full-wave dipoles, depending upon the circumstances. The characteristics of such dipoles are described in Chapter 13 (*H.F. Aerials*).

When elements are arranged together to form a beam aerial, the radiation pattern most interesting to amateurs is one having a large forward gain. However, by the study of beam-aerial polar diagrams it becomes apparent that in the course of building up a major lobe, other minor or subsidiary lobes may be created. No purpose would be served here by discussing these minor lobes further but it is well to remember that they exist and do not necessarily imply an inferior aerial array although they do represent a loss of power from the main lobe.

Parasitic Arrays: The Yagi Array

By placing a *reflector*, usually a resonant element one half-wavelength long behind a half-wave dipole, the radiation can be concentrated within a narrower angle. By adding further elements somewhat shorter than one half-wavelength, called *directors*, at certain spacings in the forward direction, a further gain can be achieved. Any aerial array which employs elements not directly connected to the feed line, i.e. parasitic elements, is known as a *parasitic array*. If the

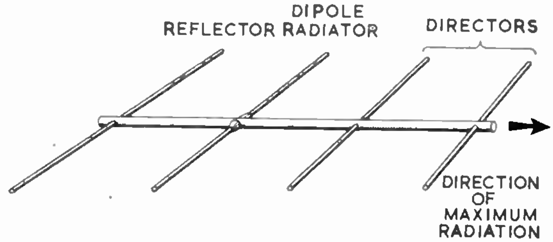


Fig. 14.17. Yagi array (4 elements). See Table 14.3 for typical dimensions.

arrangement consists of a dipole with a reflector and two or more directors it is known as a *Yagi array:* see Fig. 14.17.

When compared with other aerial systems of similar size the Yagi array is found to have the highest forward gain, and it can be constructed in a very robust form. The effect of adding the reflector and director(s) is to cause the feed impedance of the dipole to fall considerably, often to a value of about 10 ohms, and the matching is then critical and difficult to obtain. This, however, may be overcome by the use of a folded-dipole radiator. If the folded dipole has two elements of equal diameter, a 4 : 1 impedance step-up is obtained. By varying the ratio of the diameters, different impedance step-up ratios become available.

The length of the dipole or folded dipole required for resonance depends not only on the frequency but also to a lesser extent on ratio of the diameter of the element to the wavelength and the distance between the arms of the fold, the length required for resonance diminishing as the wavelength/diameter ratio is decreased: see Table 14.2. The

TABLE 14.2

Resonant Lengths of Half-wave Dipoles

$\left(\frac{\text{Wavelength}}{\text{Diameter}}\right)$	Value of $\left(\frac{\text{Dipole length}}{\text{Wavelength}}\right)$ for resonance	Feed Impedance
50	0·458	60·5 ohms
100	0·465	61·0 ohms
200	0·471	61·6 ohms
400	0·475	63·6 ohms
1,000	0·479	65·3 ohms
4,000	0·484	67·2 ohms
10,000	0·486	68·1 ohms
100,000	0·489	69·2 ohms

The dimensions used in calculating the ratios must be in similar units (e.g. both in metres or both in centimetres).
From *Aerials for Metre and Decimetre Wavelengths* by R. A. Smith.

overall length of conductor in the folded dipole remains approximately constant: as spacing is increased, the length of the radiating part is shortened accordingly.

The forward gain is not appreciably affected by a variation of reflector spacing over a range of $\lambda/8$ to $\lambda/4$: under these conditions the forward gain is approximately at its maximum value. A considerable change in feed impedance takes place when this spacing is varied and this may be used as a convenient form of adjustment. The reflector is usually $0{\cdot}5\lambda$ long although this should be reduced to about $0{\cdot}475\lambda$ for the closer spacing.

The length of the directors is usually made about $0{\cdot}43\lambda$ and the spacing approximately $0{\cdot}25\lambda$, but experiments have shown that where several directors are used, the bandwidth can be broadened by making them progressively shorter in the direction of radiation. The greater the number of directors, the higher the gain and the narrower the beamwidth. There is no advantage to be derived from using more reflectors spaced behind the first, but the front-to-back ratio may be improved somewhat by the use of additional reflectors as shown in Fig. 14.18. These additional reflector elements should subtend a fairly wide angle at the farthest director to be effective. In practice, a trigonal reflector element/element spacing of about a quarter-wavelength is sufficient.

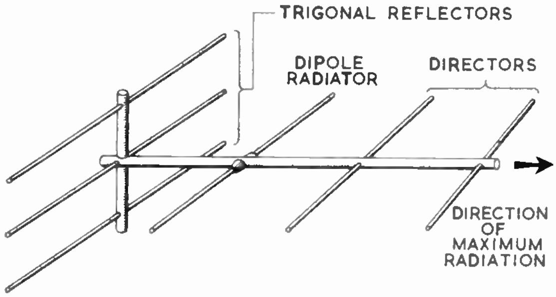


Fig. 14.18. Yagi array with trigonal reflectors.

TABLE 14.3

Typical Dimensions of Yagi Arrays—Parasitic Elements

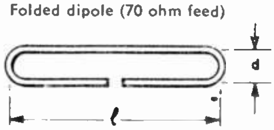
Element	Length of Element		
	70·3 Mc/s	145 Mc/s	433 Mc/s
Reflector	85½ in.	40 in.	13¼ in.
Director D1	74 in.	35½ in.	11¼ in.
Director D2	73 in.	35¼ in.	11⅛ in.
Director D3	72 in.	35 in.	11 in.
Succeeding directors	1 in. less	½ in. less	⅛ in. less
Final director	2 in. less	1 in. less	¼ in. less
One wavelength (for reference)	168 in.	81½ in.	27¼ in.
Diameter range for length given	½–¾ in.	¼–⅜ in.	⅛–¼ in.
	Spacing Between Elements		
Reflector to radiator	22½ in.	17½ in.	5½ in.
Radiator to director 1	29 in.	17½ in.	5½ in.
Director 1 to director 2	29 in.	17½ in.	7 in.
Director 2 to director 3, etc.	29 in.	17½ in.	7 in.

The above figures are based on a number of proved Yagi array designs. If the slot is used two sets of reflectors/directors are required, one mounted above the other thus forming a stacked array—see Fig. 14.22. Match to the feeder can be effected by moving the radiator relative to the first director and the reflector relative to the radiator.

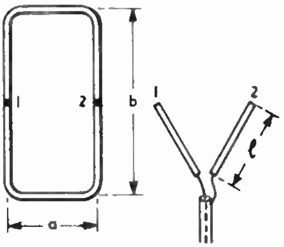
Typical Dimensions of Yagi Arrays—Driven Elements

	Length in Inches		
	70·3 Mc/s	145 Mc/s	433 Mc/s
Dipole (for use with gamma match)	79	38	12¾
Diameter range for length given	½–¾	¼–⅜	⅛–¼

Folded dipole (70 ohm feed)



	Length in Inches		
	70·3 Mc/s	145 Mc/s	433 Mc/s
l length centre/centre	77½	38½	12½
d spacing centre/centre	2½	⅞	½
Diameter	½	¼	⅛



	Length in Inches		
	70·3 Mc/s	145 Mc/s	433 Mc/s
a centre/centre	32	14	5¾
b centre/centre	96	43¾	15⅛
Delta feed sections (length 70 ohm feed)	22½	12	4⅜
Diameter of slot and delta feed	½	⅜	⅜

The gain obtainable from a Yagi array compared with a half-wave dipole is shown approximately by the curve in **Fig. 14.19.**

The bandwidth for a standing-wave ratio less than 2 : 1 is about 2 per cent for close-spaced beams and about 3 per cent for wider spacing. Element lengths, particularly those of the directors, are very critical (i.e. within fractions of an inch), and ideally telescopic rods should be used to enable fine adjustments to be made. Each change of element length necessitates a readjustment of the matching either by moving the reflector or in the matching device itself.

Typical element lengths for spot frequencies in the 4m, 2m and 70cm bands are given in **Table 14.3.** The lengths are based on the assumption that the element diameter lies within the stated limits for the respective bands. Any departures from these diameter ranges will necessitate a change in the lengths of the elements; for a larger diameter the length will need to be decreased, and vice versa. Provision has been made for simple dipole, folded dipole or skeleton slot radiating elements. The latter radiator, which is described later in the chapter, is the feed system in a stacked pair of Yagi arrays, the dimensions of which are the same as for a single Yagi array given in Table 14.3.

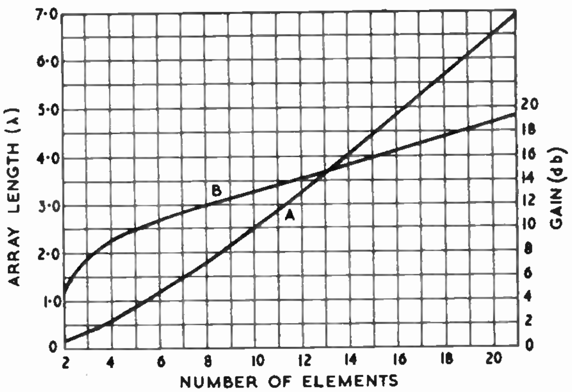


Fig. 14.19. Design information for Yagi aerials. Curve A shows the optimum boom length in wavelengths for any number of elements. Curve B shows the maximum gain that can be expected when the design information of Curve A is used.

With acknowledgements to the A.R.R.L.

Stacked and Bayed Yagi Arrays

To obtain a greater gain several Yagi arrays can be disposed horizontally or vertically or in both directions, the radiating elements being fed in phase. Normally, *stacking* refers to vertical addition and *baying* to horizontal. It is important to remember that with stacking the effective height of the aerial above the ground is half way between the lowest and highest individual arrays and thus the advantage of increased gain may be reduced—see Height of Aerial above Ground on page 14.3. Typical optimum stacking distances are shown in **Table 14.4.** The optimum stacking distance will

TABLE 14.4

Optimum Stacking Distance for Yagi Arrays

Yagi Array	Centre-to-centre Stacking Distance		
	70·3 Mc/s	145 Mc/s	433 Mc/s
3 element	84 in.	—	—
4 "	108 in.	82 in.	—
8 "	—	120 in.	—
10 "	—	132 in.	44 in.
14 "	—	—	66 in.
24 "	—	—	—
4-over-4 slot fed	—	120 in.	40 in.
6-over-6 slot fed	—	147 in.	49 in.
8-over-8 slot fed	—	160 in.	53 in.

With acknowledgments to J-Beam Aerials Ltd.

depend on the gain of the individual Yagi arrays and is the distance at which the apertures of each array just touch. Where the vertical and horizontal beamwidths of an individual array are approximately the same, i.e. with five or more elements, the optimum stacking distance will be approximately the same, both horizontally and vertically; this distance varies from about 0·75λ for an array of 4-element Yagis to about 2·0λ or more for 8-element Yagis. For practical reasons the spacing is usually less than optimum and there is a consequent reduction in the total gain. It was stated earlier that the increase in gain obtainable by stacking two identical arrays in such a way that their capture areas do not overlap is simply two times (i.e. 3db), but in practice this can sometimes be exceeded if there is a suitably favourable degree of coupling between the two arrays: for example, a pair of three element Yagi arrays could be made to yield an increase of 4·2db. However, the increase usually proves to be less than the theoretical 3db and a figure of 2·2–2·5db is all that can be ordinarily expected.

Theoretically the feed impedance of a stacked array is the feed impedance of an individual Yagi array divided by the total number of Yagis employed; in practice the feed impedance of stacked Yagis is slightly less than this because of interaction between each Yagi array although this reduction is not so marked at the greater spacings.

Disadvantages of Conventional Yagi Arrays

Perhaps the most important disadvantage is that the variation of the element lengths and spacings causes interrelated changes in the feed impedance of a Yagi array. To obtain the maximum possible forward gain experimentally is extremely difficult because for each change of element length it is necessary to readjust the matching either by moving the reflector or by resetting a matching device. A method has been devised, however, for overcoming these practical disadvantages by the use of a radiating element in the form of a *skeleton slot*, this being far less susceptible to the changes in impedance caused by changes in the parasitic-element lengths. This development is due to B. Sykes, G2HCG.

A true slot would be a slot cut in an infinite sheet of metal, and such a slot when approximately one half-wavelength long would behave in a similar way to a dipole radiator. In contrast with a dipole, however, the polarization produced by a vertical slot is horizontal (i.e. the electric field is horizontal).

The skeleton slot was developed in the course of experiments to determine to what extent the infinite sheet of metal could be reduced before the slot aerial lost its radiating property. The limit of the reduction for satisfactory performance was found to occur when there remained approximately one half-wavelength of metal beyond the slot edges. Further experiments showed that a thin rod bent to form

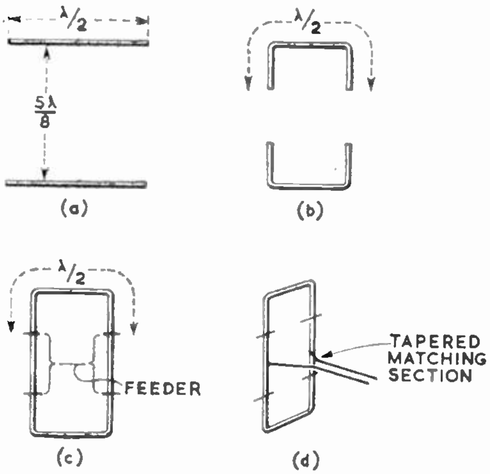


Fig. 14.20. Development of a skeleton-slot radiator from two half-wave dipoles spaced five-eighths of a wavelength apart.

a "skeleton slot" of dimensions approximately $5\lambda/8 \times 5\lambda/24$ exhibited similar properties to those of a true slot. The manner in which a skeleton slot functions can be understood by referring to the diagrams in **Fig. 14.20.** Consider two half-wave dipoles spaced vertically by $5\lambda/8$. Since the greater part of the radiation from each dipole takes place at the current anti-node, i.e. the centre, the ends of the dipoles may be bent without serious effect. These ends may now be joined together with a high-impedance feeder, so that end-feeding can be applied to the bent dipoles. To radiate in phase, the power should be fed midway between the two dipoles. The high impedance at this point may be transformed down to one suitable for the type of feeder in use by means of a tapered matching-section transmission line (i.e. a delta match). Practical dimensions of a skeleton-slot radiator are given in **Fig. 14.21.**

It is important to note that two sets of parasitic elements are required with a skeleton-slot radiator and not one set as required with a true slot. One further property of the skeleton slot is that its bandwidth is somewhat greater than a pair of stacked dipoles. A disadvantage of the basic slot-fed beam is that the spacing between upper and lower sets of parasitic elements is not optimum and thus the gain increase

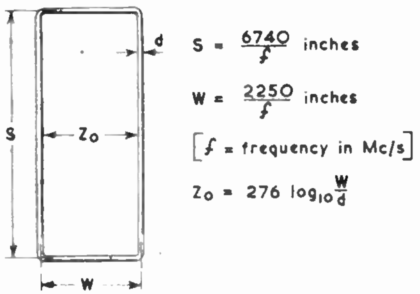


Fig. 14.21. Dimensional relationships of a skeleton-slot radiator. Both *S* and *W* may be varied experimentally from the values indicated by these formulae. For small variations the radiation characteristics of the slot will not change greatly, but the feed impedance will undergo appreciable change and therefore the length of the delta matching section should always be adjusted to give a perfect match to the transmission line.

is less than 3db compared with the equivalent single Yagi. This is increasingly so with slot-fed beams arrays having a greater number of parasitic elements.

Stacked Skeleton Slot Yagi Arrays

Skeleton slot Yagi arrays may be stacked to increase the gain but the considerations of optimum stacking distance previously discussed apply; in this case the centre-to-centre spacing of a pair of skeleton slot Yagi arrays should vary between 1λ and 3λ or more according to the number of elements in each Yagi array.

Each skeleton slot Yagi array may be fed by 72 ohm coaxial cable, using equal lengths of feeder to some common feed point for the stacked array, and it would of course be desirable to use a balun at the point where the cable is attached to each array. A coaxial quarter-wave transformer can be used to transform the impedance to that of the main feeder. For example, if a pair of skeleton slot Yagi arrays, each of 72 ohms feed impedance, is stacked, the combined impedance will be one-half of 72 ohms, i.e. 36 ohms; this may be transformed to 72 ohms by the use of a quarter-wave section of 52 ohm coaxial cable, allowance being made for its velocity factor. Larger assemblies of skeleton slot Yagi arrays can be fed in a similar manner by joining pairs and introducing quarter-wave transformers until only one feed is needed for the whole array.

The Long Yagi

The gain of the Yagi depends upon the number of directors and thus its length as already shown in **Fig. 14.19.** The term long Yagi has been applied to arrays with, say, eight or more directors. Such arrays have become popular in recent years, largely due to the work of W2NLY and W6QKI. The construction of a typical Long Yagi is described on page 14.23.

Stacked Dipole Arrays

Both horizontal and vertical beam widths can be reduced and gain increased by building up arrays of driven dipoles. This arrangement is usually referred to merely as a *stack* or sometimes as a *bill-board* or *broadside array.* Since this type of array is constituted from a number of radiating dipoles, the feed impedance would be extremely low if the dipoles were centre-fed. However, the impedance to earth of a dipole at its end is high, the precise value depending upon the ratio of its length to diameter, and it will therefore be more convenient to use a balanced high-impedance feeder to end-feed a pair of collinear half-wave dipoles, a system called a *full-wave dipole.* The length for resonance and the feed impedance in terms of wavelength/diameter ratio is shown in **Table 14.5.**

The full-wave dipoles are usually mounted with a centre-to-centre spacing, horizontally and vertically, of one half-wavelength and are fed in phase. Typical arrangements for stacks of full-wave dipoles are shown in **Fig. 14.22.** Note that the feed wires between dipoles are one half-wavelength long and are crossed so that all dipoles in each bay are fed in phase. The impedance of these phasing sections is unimportant provided that the separators, if used, are made of low-loss dielectric material and that there is sufficiently wide separation at the cross-over points to prevent unintentional contact.

To obtain the radiation pattern expected, all dipoles should be fed with equal amounts of power (as indeed would be desirable in any multi-radiator array), but this may not be achieved in practice because those which are at the edges of the array have least mutual coupling to other dipoles and therefore have different radiation resistances. However, by locating the main feed point as nearly symmetrically as possible these effects are minimized. Hence it would be preferable for the aerial shown at (A) in **Fig. 14.22** to be fed in the centre of each bay of dipoles; the feeder to each bay must be connected as shown to ensure that the two bays are fed in phase. If they were fed 180° out-of-phase the resultant beam pattern would have two major side lobes and there would be very little power radiated in the desired direction. The diagram (B) in Fig. 14.22 shows two vertically stacked bays of full-wave dipoles fed symmetrically and in phase.

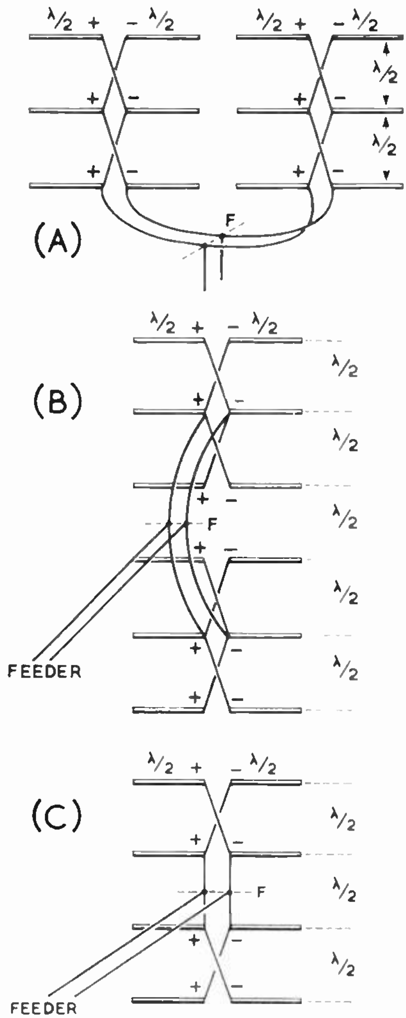


Fig. 14.22. Typical stacks of full-wave dipoles. Note that the feed-point F is equi-distant from each bay of dipoles. Examples of lengths are given in Table 14.5.

Diagram (C) in Fig. 14.22 shows one vertically stacked bay of full-wave dipoles fed symmetrically half way between the centre pair of dipoles.

The spacing at the centre of each full-wave dipole should be sufficient to prevent a reduction of the resonant frequency by the capacity between the ends. In practice this spacing is usually about 1 in. for the 1454 and 433 Mc/s bands.

Matching can be carried out by the use of movable short-circuited open-wire stubs on the feed lines.

As with the Yagi array, the gain can be increased by placing half-wave reflectors behind the radiating elements at a spacing of 0·1–0·25λ, a figure of 0·125λ being frequently chosen. A perfect plane reflector will yield a gain of 6db compared with the 3db obtainable from a rod reflector. For the 433 Mc/s band and for higher frequencies, a plane reflector made up of 1 in. mesh wire netting stretched on a frame can be used in place of the resonant reflector at a similar spacing. The mesh of the wire should be so orientated that the interlocking twists are parallel to the dipole. The wire netting should extend at least one half-wavelength beyond the extremities of the dipoles in order to ensure a high front-to-back ratio.

TABLE 14.5

Resonant Lengths of Full-wave Dipoles

(Wavelength / Diameter)	Value of (Dipole length / Wavelength) at resonance	Feed Impedance
50	0·85	500 ohms
100	0·87	900 ohms
150	0·88	1100 ohms
200	0·896	1300 ohms
300	0·906	1500 ohms
400	0·916	1700 ohms
700	0·926	2000 ohms
1,000	0·937	2400 ohms
2,000	0·945	3000 ohms
4,000	0·951	3600 ohms
10,000	0·958	4600 ohms

The dimensions used in calculating the ratios must be in similar units (e.g. both in metres or both in centimetres).
From *Aerials for Metre and Decimetre Wavelengths* by R. A. Smith.

The half-wave sections of the full-wave dipole should be supported at the current anti-nodes, i.e. at their centres, either on small insulators or in suitably drilled wooden vertical members. Supports should not be mounted parallel to the elements because of possible influence on the properties of the aerial.

The bandwidth of this type of aerial is exceptionally large and its adjustments generally are far less critical than those of Yagi arrays.

For a stack having a wire-net reflector extending λ/2 beyond the extremities of the dipoles, the horizontal beam-width θ, vertical beamwidth ϕ and power gain G (compared with an isotropic radiator) can be calculated approximately from the following formulae:

$$\theta = \frac{51\lambda}{a} \qquad \phi = \frac{51\lambda}{b} \qquad G = \frac{4\pi ab}{\lambda^2}$$

where a and b are the horizontal and vertical dimensions of the reflector respectively, both being expressed in the same units as the wavelength.

These formulae are true only for an array which is large compared with the wavelength, but are suitable as a criterion for judging aerials of any type provided the equivalent aperture or capture area is known.

Skeleton Slots in Stack

Skeleton slots can be used to replace vertically disposed pairs of half-wave dipoles. As the optimum vertical dimension for a horizontally polarized skeleton slot is approximately $5\lambda/8$, it is no longer possible to use the vertical spacings shown for full-wave stacks. The slots are mounted vertically at a centre-to-centre spacing of one wavelength and fed through a tapered matching section, as for the skeleton-slot Yagi array, and are then connected to the phasing lines. Since the spacing between feed points is one wavelength there is no phase difference and it is unnecessary to transpose the phasing wires. The tapered matching sections should be adjusted to present an impedance of N times the desired feeder-cable impedance where N is the number of skeleton slots employed. The impedance resulting from the connection of all the feed points together will then equal the cable impedance.

A broadside array of skeleton slots may be built up by adding further slots horizontally at a centre-to-centre spacing of one half-wavelength.

Disadvantage of Multi-element Arrays

As the frequency becomes higher and the wavelength becomes shorter, it is possible to construct arrays of much higher gain although, as already described, the advantage is offset by the reduction in capture area when used for receiving. However, if the practice already described, that of using many driven or parasitic elements, either in line or in stack, is adopted, the complications of feeding become increasingly greater. Also as the frequency increases, the radiation-loss from open-wire lines and from phasing and matching sections likewise increases, and it is then difficult to ensure an equal power-feed to a number of radiators. Preferably, therefore, the aerial should have a minimum number of radiating or other critical elements, such as resonant reflectors or directors. There are many aerials in this category but only those having immediate amateur application are described here.

The Cubical Quad

The cubical quad aerial has proved highly successful for v.h.f. use especially indoors where its properties are almost unaffected by the close proximity of building structures in contrast to the Yagi array, the high Q elements of which are detuned with the result that its performance is markedly changed.

The basic cubical quad (Fig. 14.23) can be constructed using copper wire of $\frac{1}{16}$ in. or $\frac{1}{8}$ in. diameter. The reflector square can either be made larger than the radiator by about 5 per cent in total length or alternatively can have its length extended electrically by inserting a small inductive stub at a current anti-node, i.e. short circuited and considerably less than a quarter-wave in length. A variation of the quad, developed by G2PU, known as the bi-square, has two symetrically disposed inductive stubs located at current anti-nodes in the reflector which ensures current symmetry in the reflector and hence no beam tilt (see Fig. 14.24). V.H.F. bi-square aerials are usually constructed from $\frac{3}{8}$ in. diameter tube, although wire can be used.

The gain of a quarter-wave square cubical quad is 5·5 to 6db but this can be increased by using a director square in front of the radiator. This square can either be some 5 per cent shorter in total length than the radiator or have a short capacitive stub inserted, i.e. open circuited and considerably less than a quarter-wave. This increases the gain by about 1·5db. It is unusual to see more than one director although there is no reason why the Yagi principle of multiple directors cannot be applied.

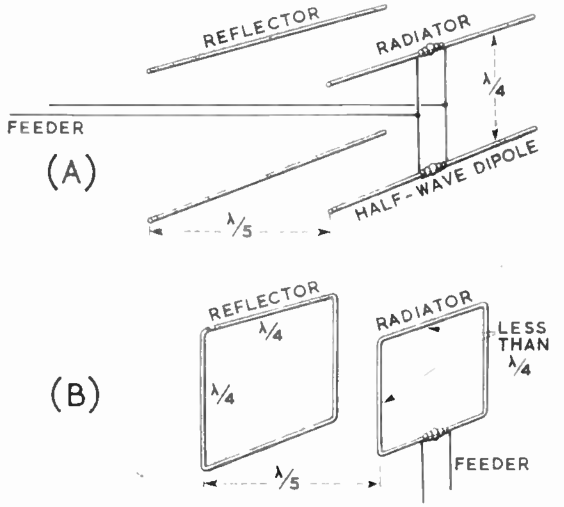


Fig. 14.23. Development of a cubical quad aerial from a pair of dipoles and reflectors. In (A) the reflectors are $\lambda/2$ long and the dipole radiator lengths for resonance are approximately 0·48λ. In (B) the elements have been bent to form two squares, the combined reflector having sides 0·25λ long and the combined radiator having sides 0·24λ long. The two squares can be made equal in size by the method shown in Fig. 14.24.

Two band arrays can conveniently be placed one inside the other as no significant interaction occurs. In the case of a quad without a director, the wires can be attached to bamboo spreaders originating from one central multiple X having eight arms. In the case of the quad with director, three separate bamboo X spreaders can be located on a single boom. In all, the array is non-critical, has a wide bandwidth and is ideal for home construction with limited facilities. It will pay to experiment both with stub lengths and reflector/radiator/director spacing to achieve the maximum gain with minimum feeder s.w.r. in any particular environment.

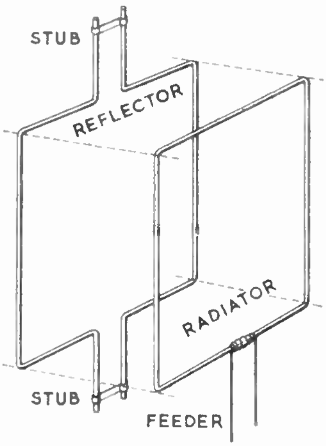
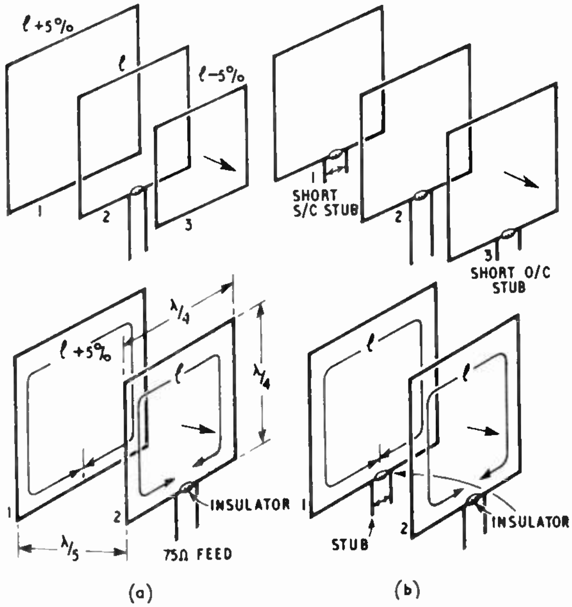


Fig. 14.24. The bi-square aerial may be fitted with small stubs in the reflector element so that all the sides of both reflector and radiator can be made equal in length (approximately 0·24λ).

TABLE 14.6



Band	Reflector 1 total length *	Radiator 2 * total length *1	Director (if used)	Approx. length of stubs if used: reflec-tor s/c	direc-tor s/c	Spacing
70 Mc/s (a)	173 in.	165 in.	157 in.	—	—	34 in.
70 Mc/s (b)	165 in.	165 in.	165 in.	8 in.	8 in.	34 in.
144 Mc/s (a)	84 in.	80 in.	76 in.	—	—	16 in.
144 Mc/s (b)	80 in.	80 in.	80 in.	4 in.	4 in.	16 in.

The total length of the radiator is just less than one wavelength in total (about 0·98 × λ). The stub in (b) should be adjusted in length for maximum forward gain. The gain is about 5½ to 6db. Elements are made from ⅛ in. diameter copper wire although the diameter is not critical. Insulators can be made from polystyrene or polythene about 1½ in. wide. Stub lengths should be adjusted for optimum performance.
*each side is one quarter this length.

Typical dimensions for both quarter- and half-wave types of quads are shown in Tables 14.6 and 14.7.

In Table 14.6 dimensions are given for typical quarter wave types with and without stubs. The half-wave quad has an additional insulator half way along the top conductor of all elements of the squares shown in Table 14.6 i.e. symmetrically opposite the feed or stub points. The feed impedance is high (between 2 and 4 thousand ohms) and this may be transformed down to 75 ohms by means of a quarter-wave transformer.

TABLE 14.7

Half-wave cubical quad.
Gain without director 10db; with director 12·5db

Band	Reflector Total length	Radiator Total length	Director (if used)	Approximate length of stubs if used: Reflector S/C	Director O/C	Spacing
144 Mc/s	174 in.	164 in.	154 in.	—	—	16 in.
144 Mc/s	164 in.	164 in.	164 in.	10 in.	10 in.	16 in.

Corner Reflector

The use of a metallic plane reflector spaced behind a radiating dipole has already been discussed. If this reflector is bent to form a *V*, as shown in Fig. 14.25, a considerably higher gain is achieved. The critical factors in the design of

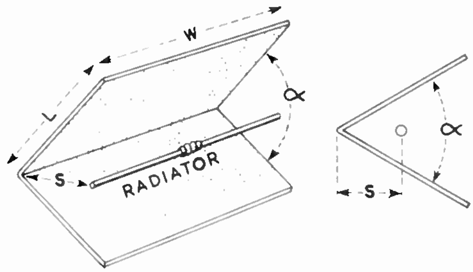


Fig. 14.25. Corner reflector. The half-wave dipole radiator is spaced parallel with the vertex of the reflector at a distance *S*. Its characteristics are shown in Figs. 14.26 and 14.27. For dimensions, see Table 14.9.

such an aerial array are the corner angle α and the dipole/vertex spacing *S*. The curves in Fig. 14.26 show that as α is reduced, the gain theoretically obtainable becomes progressively greater. However, at the same time the feed impedance of the dipole radiator (Fig. 14.26) falls to a very low value, as can be seen from Fig. 14.27: this makes match-

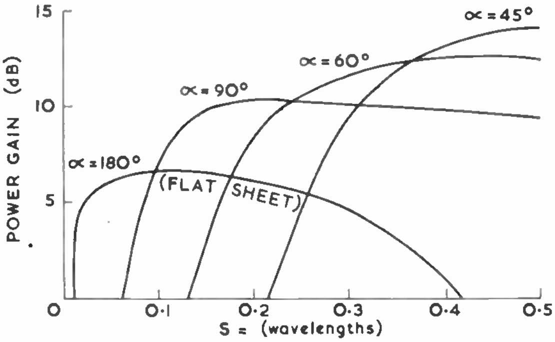


Fig. 14.26. Theoretical power gain obtained by using a corner reflector with a half-wave dipole radiator: see Fig. 14.21.

ing difficult and hence a compromise has to be reached. In practice the angle α is usually made 90° or 60°; adjustments in a 60° corner are a little more critical although the maximum obtainable gain is higher. The final matching of the radiator to the line may be carried out by adjusting the distance *S*, which as seen from Fig. 14.26 does not greatly affect the gain over a useful range of variation but causes a considerable change in feed impedance (see Fig. 14.27). A two-stub tuner may also prove helpful in making final adjustments.

The length of the sides *L* of the reflector should exceed two wavelengths to secure the characteristics indicated by Figs. 14.26 and 14.27, and the reflector width *W* should be greater than one wavelength for a half-wave dipole radiator. The reflecting sheet may be constructed of wire-netting as described previously or alternatively may be fabricated from metal spines arranged in a V-formation, all of them being

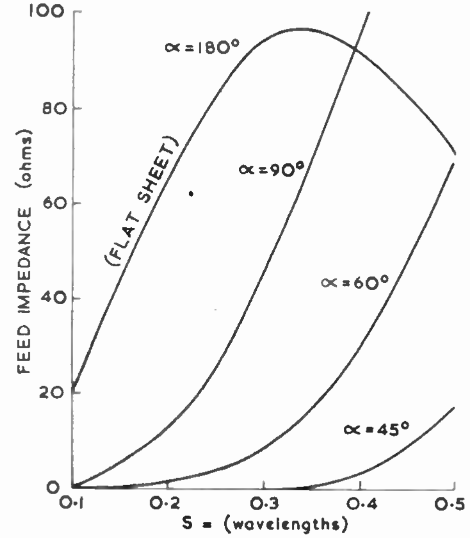


Fig. 14.27. Feed impedance of a half-wave dipole provided with a corner reflector: see Fig. 14.21.

parallel to the radiator: see **Fig. 14.28.** The spacing between adjacent rods should not exceed 0·1λ and the length of the rods is 0·6λ for a half-wave driven element.

A useful approximation for the power gain G referred to a half-wave dipole is $G = 300/\alpha$, where α is the angle between the sides measured in degrees.

The maximum dipole/vertex spacing S included in the curves shown in Figs. 14.26 and 14.27 is one half-wavelength. Spacings greater than this would require rather cumbersome constructions at lower frequencies, but at the higher frequencies larger spacings become practicable, and higher gains can then be obtained. This indicates that the corner reflector can become a specially attractive proposition for the 1296 Mc/s band, but the width across the opening should be in excess of 4λ to achieve the results shown. Dimensions for 60°, 13db gain corner reflectors for 145, 433 and 1296 Mc/s are given in **Table 14.8.**

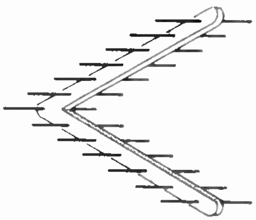


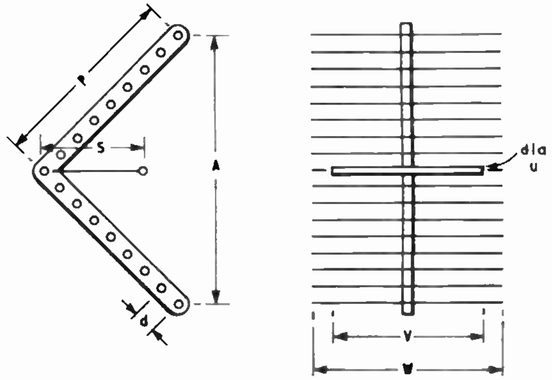
Fig. 14.28. The corner reflector shown in Fig. 14.25 can be modified by using a set of metal spines arranged in V-formation to replace the sheet-metal or wire-netting reflector.

Trough Reflector

To reduce the overall dimensions of a large corner reflector the vertex can be cut off and replaced with a plane reflector, such an arrangement is known as a *trough reflector:* see **Fig. 14.29.** Similar performance to that of the large corner reflector can thereby be achieved provided that the dimensions of the trough do not exceed the limits indicated in **Table 14.9.**

TABLE 14.8

Dimensions for a 60° corner reflector aerial system giving a gain of about 13db. The feed impedance of the dipole radiator is 75 ohms. The apex may be hinged for portable work.



Band	Dimensions in Inches							
	p	s	d	v	w	A	u	
145	100	40	6	38	50	100	⅜	168
430	35	13¼	1½	12½	20	35	¼	27¼
1296	12	4½	½	4	8	12	⅛	9⅛

The resulting aerial has a performance very little different from the corner-reflector type and presents fewer mechanical problems since the plane centre portion is relatively easy to mount on the mast and the sides are considerably shorter.

The gain of both corner reflectors and trough reflectors may be increased still further by stacking two or more and

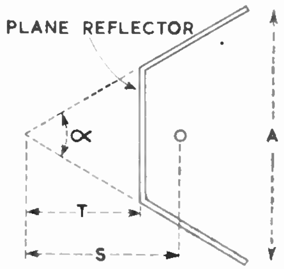


Fig. 14.29. Trough reflector. This is a useful modification of the corner reflector shown in Fig. 14.25. The vertex has been cut off and replaced by a simple plane section.

arranging them to radiate in phase, or alternatively by adding further collinear dipoles within a wider reflector similarly fed in phase. Not more than two or three radiating units should be used since the great virtue of the simple feeder arrangement would then be lost.

TABLE 14.9

Corner/Trough Reflector

Angle α	Value of S for maximum gain	Gain	Value of T
90°	1·5λ	13db	1λ—1·25λ
60°	1·25λ	15db	1·0λ
45°	2·0λ	17db	1·9λ

This table shows the gain obtainable for greater values of S than those covered by Fig. 14.26, assuming that the reflector is of adequate size.

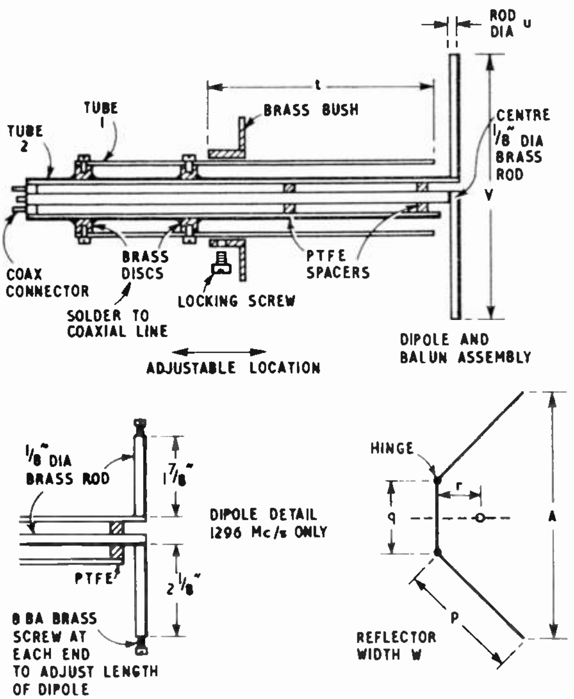
Trough Reflectors for 433 and 1296 Mc/s

Dimensions are given in Table 14.10 for 433 and 1296 Mc/s trough reflectors. The gain to be expected is 15db and 17db respectively. A very convenient arrangement, especially for portable work, is to use a metal hinge at each angle of the reflector. This permits the reflector to be folded flat for transit. It also permits experiments to be carried out with different apex angles.

A housing will be required at the dipole centre to prevent the ingress of moisture and also, in the case of the 433 Mc/s aerial, to support the dipole elements. The dipole may be moved in and out of the reflector to achieve either minimum standing wave ratio or, if this cannot be measured, for maximum gain. If a two stub tuner or other matching device is used, the dipole may be placed to give optimum gain and the matching device adjusted to give optimum match. In the case of the 1296 Mc/s aerial, the dipole length can be adjusted by means of the brass screws at the ends of the elements. Locking nuts are essential.

TABLE 14.10

Trough Reflectors for 433 and 1296 Mc/s



Band	Dimensions in Inches								Tube 1 Brass or copper	Tube 2 Brass or copper	Gain
	A	p	q	r	w	t	u	v			
433	120	100	50	7½	60	6½	¼	12¾	1½ in. i.d. 1/16 in. wall	⅝ in. o.d. 1/16 in. wall	17db
1296	45	36	9½	2½	20	2⅛	⅛	4¼	¾ in. i.d.	⅜ in. o.d. 20 s.w.g.	15db

The reflector should be made of sheet aluminium for 1296 Mc/s but can be constructed of wire mesh (with twists parallel to the dipole) for 433 Mc/s. To increase the gain by 3db a pair can be stacked so that the reflectors are just not touching (to avoid a slot radiator being formed by the edges). The radiating dipoles must then be fed in phase and suitable feeding and matching must be arranged. A two stub tuner can usefully be used for matching either for a single or double reflector system.

The Reflex Aerial

The reflex aerial, shown in Fig. 14.30, comprises a radiating dipole located in front of a large reflecting sheet with a grating type of structure mounted in front of and parallel to the dipole. The total gain attainable with this aerial system is about 12db.

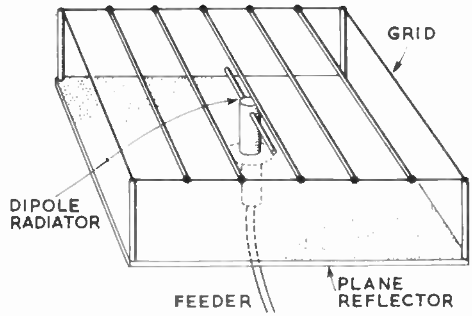


Fig. 14.30. Reflex aerial. Note the grid-like structure in front of the dipole radiator. The plane reflector behind the radiator may be made of sheet metal or wire netting (1 in. mesh for 433 Mc/s; ½ in. mesh for 1296 Mc/s).

The dipole has a feed impedance of about 120 ohms and can be fed from a coaxial cable with a suitable matching device and balun. An arrangement that has been used successfully at 1296 Mc/s is to feed the dipole from open lines some 3½ in. long between the coaxial cable and the dipole (see Fig. 14.31). An open circuited stub is then adjusted in position and length to give maximum forward radiation as detected by a dipole with crystal detector and microammeter at about 20 ft. distant. A typical final setting resulted in the stub about 1½ in. long at about 1 in. from the coaxial feeder. No balun was used. The perform-

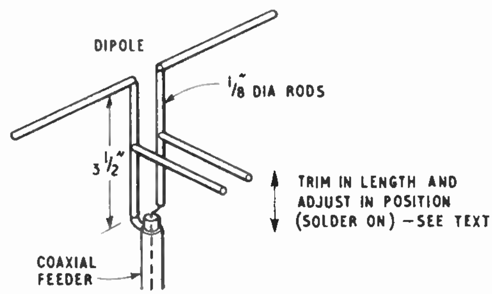


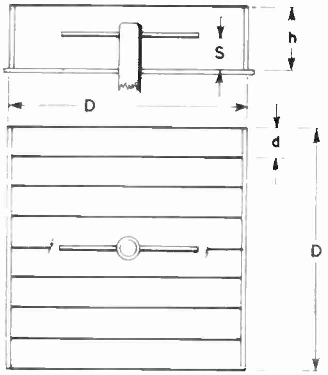
Fig. 14.31. A possible feed arrangement for a 1296 Mc/s reflex aerial.

ance of the aerial adjusted in this way proved to be very satisfactory.

Typical dimensions for the 433 Mc/s and 1296 Mc/s bands are shown in **Table 14.11**. There is no advantage in increasing the reflector size beyond about two wavelengths square since above this size the oblique radiation from the dipole is so slight that little further gain can be achieved. However, the gain may be increased by adding further reflex aerials either horizontally or vertically as desired and feeding them so as to radiate in phase.

TABLE 14.11

Typical dimensions of a reflex aerial for 433 Mc/s and 1296 Mc/s. With these dimensions the gain should be 11–12db. If the size of the reflector D is increased to 2-wavelengths square, the gain rises to about 16db.

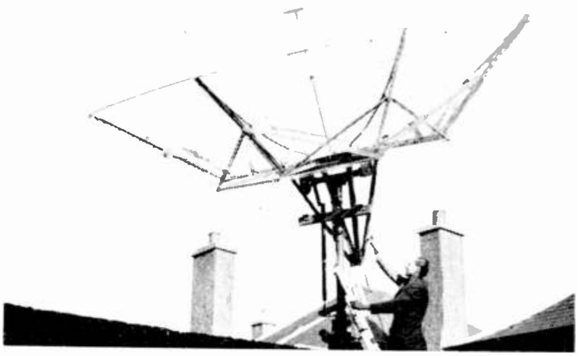


Band	D	S	h	d	rods diam.	Dipole	
						length	diam.
433 Mc/s 1296 Mc/s	30 in. 18 in.	7½ in. 2½ in.	12 in. 4 in.	7½ in. 2½ in.	½ in. 5/32 in.	12½ in. 4¼ in.	¼ in. ⅛ in.

The Parabolic Reflector

The principle of the parabolic reflector as used for light waves is well known. Such a reflector may be used for radio waves, although the properties are somewhat different because the ratio of the diameter of the reflector to the wavelength is not extremely large, as it is in the case of light waves. If a radiator is placed at the focus of a paraboloid and if most of the energy is directed back into the "dish," a narrow beam will be produced. Assuming that the energy is uniformly distributed over the dish, the angular width of the beam θ will depend on the diameter of the reflector A approximately according to the formula $\theta = 58\lambda/A$, where A is expressed in wavelengths. To avoid the obvious difficulties in making a parabolic reflector from sheet metal, a skeleton form of construction of wire netting stretched over wooden ribs may be found successful for frequencies up to 1500 Mc/s but as the wavelength becomes shorter the overall surface contour of the reflector must approach that of the true paraboloid more closely.

A suitable parabolic reflector for the 1296 Mc/s and 2400 Mc/s bands may be 1–3 ft. in diameter. The radiator element is usually a half-wave dipole having a resonant reflector in the form of a disc one half-wavelength in diameter mounted one



Example of an amateur built parabolic aerial.

quarter-wavelength *in front* of the dipole to reflect the radiation back into the parabolic dish. The centre of the dipole should be accurately positioned at the focus of the parabola. A balun transformer of the coaxial-sleeve type may be mounted on the feed stem.

A typical aerial assembly with a suitable feeder arrangement is shown in **Fig. 14.32.** For this system a two-stub tuner will be found particularly convenient. The curve of the dish must not deviate from a true parabola by more than $\lambda/12$ at any point and thus the higher the frequency the more difficult it is to make a satisfactory reflector. Where wire mesh and ribs are used, the approximation to the rue parabola must still be within $\lambda/12$ and the diameter of the holes in the mesh should also not exceed this amount. The radiator assembly can be based on the design given in Table 14.10 for the trough reflector. An alternative feed arrangement that has been used very successfully at 1296 Mc/s is shown in Fig. 14.34.

In the 5000 to 10,000 Mc/s region it is normal to use a wave guide feed either having a resonant dipole or alternatively a horn radiator.

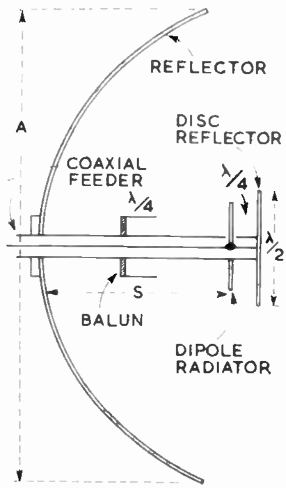


Fig. 14.32. Typical design for a parabolic reflector and feeder arrangement. The dipole should be at the focus of the parabola. To achieve maximum gain the distance S should be made adjustable. The coaxial-sleeve balun should be set back from the dipole.

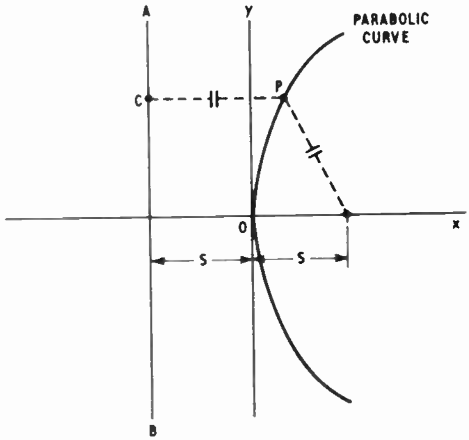


Fig. 14.33. Parabolic curve: equation $y^2 = 4Sx$. The curve is the locus of points equi-distant from a fixed point, the focus F, and a fixed line AB, called the directrix, i.e. FP = PC. The focus F has the co-ordinates (S, O).

A parabolic shape can be computed by plotting the curve $y^2 = 4Sx$ (see **Fig. 14.33**). If a suitable template is constructed from this formula, a complete dish can be fabricated using the template for reference purposes.

The gain of a parabolic reflector aerial system depends upon the diameter of the dish and the illumination of the dish from the feedpoint. Ideally, there should be a minimum radiation of energy from the feedpoint in directions other than into the dish. On transmission a spill-over from the feedpoint represents a loss of possible gain and, in reception, may result in space communication in the interception of unwanted hot body radiation from the earth, i.e. noise. A complete discussion of feed systems for parabolic reflectors is, however, outside the scope of this handbook.

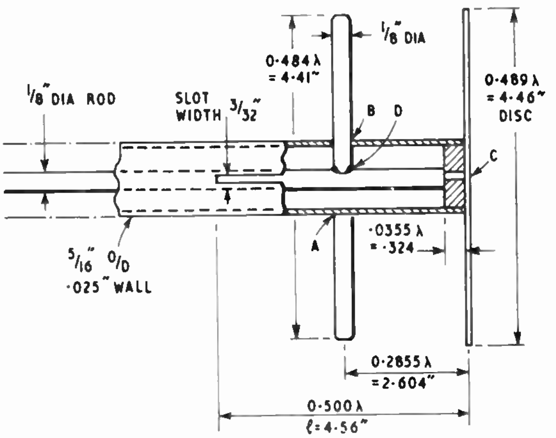


Fig. 14.34. Dipole/reflector symmetric feed arrangement for a 1296 Mc/s parabolic reflector. One half of the dipole is soldered at A. The other half is threaded 6 B.A. and is screwed into the centre conductor at D. It is finally soldered to the outer tube at B.

TABLE 14.12

Parabolic Reflector
Gain for various dish diameters

Band	Gain				
	10 db	15 db	20 db	25 db	30 db
433 Mc/s	3ft.	5ft.	10 ft.	15 ft.	30 ft.
1296 Mc/s	1 ft.	2ft.	3·5 ft.	6 ft.	12 ft.
2400 Mc/s		1 ft.	2 ft.	3 ft.	4 ft.
10,000 Mc/s					1·5 ft.

These are approximate figures. The actual gain achievable will depend upon the feed arrangements and upon the accuracy of the reflector relative to a true parabola.

The Helical Aerial

Another simple beam aerial possessing high gain and wide-band frequency characteristics simultaneously is the *helical aerial:* see **Fig. 14.35** and **Table 14.13.** When the circumference of the helix is of the order of one wavelength axial radiation occurs; i.e. the maximum field strength is found to lie along the axis of the helix. This radiation is circularly polarized, the sense of the polarization depending on whether the helix has a right- or left-hand thread viewed from the driven end

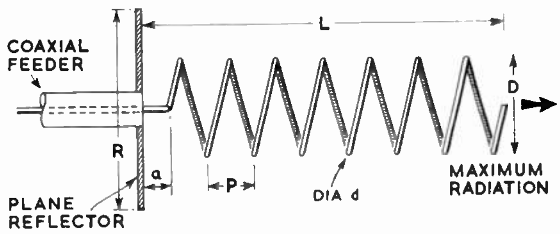


Fig. 14.35. The helical aerial. The plane reflector may take the form of a dartboard type of wire grid. The dimensions given in Table 14.13 are based on a pitch angle of 12 degrees. The helix, which may be wound of copper tube or wire the actual diameter of which is not critical, must be supported by low loss insulators.

If a pick-up dipole is used to explore the field in the direction of maximum radiation, the signal received by this dipole will show no change of amplitude as it is rotated through 360°, thus indicating true circular polarization. At any point to the side of the helix the wave will be elliptically polarized, i.e. the horizontal and vertical components will be of unequal strength.

A helix may be used to receive the circularly polarized waves radiated from a transmitting helix, but care must be taken to ensure that the receiving helix has a thread of the same sense as the radiator; if a thread of the wrong sense is used, the received signal will be very considerably weaker.

The properties of the helical aerial are determined by the diameter of the spiral D and the pitch P and depend upon the resultant effect of the radiation taking place all along the helical conductor. The gain of the aerial depends on the number of turns in the helix. The diameter of the reflector R should be at least one half-wavelength. The diameter of the helix D should be about $\lambda/3$ and the pitch P about $\lambda/4$.

It will be seen from Table 14.13 that the feed impedance is then about 140 ohms unbalanced; this may be transformed to the feeder impedance by means of a quarter wave transformer or a two stub tuner.

It is important to note that the gain figures quoted will only be achieved if a circularly polarized aerial of the same sense, such as a helix, is used for reception. If a plane polarized aerial, such as a dipole, is used there will be a loss of 3db.

TABLE 14.13

The Helical Aerial—Dimensions and Performance

Band	Dimensions				
	D	R	P	a	d
General	0·32λ	0·8λ	0·22λ	0·12λ	
433 Mc/s	8¾ in.	22 in.	6 in.	3 in.	¼ in.
1296 Mc/s	3 in.	7 in.	2 in.	1⅛ in.	¼ in. to ⅛ in.

Turns	6	8	10	12	20
Gain	12db	14db	15db	16db	17db
Beamwidth	47°	41°	36°	31°	24°

The gain and beamwidth of the helical aerial is dependent upon the total number of turns as shown above.

$$\text{Bandwidth} = 0{\cdot}75 \text{ to } 1{\cdot}3\lambda$$

$$\text{Feed impedance} = 140 \times \frac{\text{circumference}}{\lambda} \text{ ohms}$$

$$\text{Beamwidth} = \sqrt{\frac{12{,}300}{\text{No turns}}} \text{ degrees}$$

The Long Helix

As with the Yagi, the gain of a helix aerial can be increased by extending the total effective length of the aerial. In the case of the helix, this is achieved by increasing the number of turns as shown in the second part of Table 14.13. The long helix has been used for frequencies as high as 11 Gc/s with a gain of over 25db and a beamwidth of approximately 5° between half power points. The helix had more than 500 turns. For amateur bands above 1296 Mc/s long helices provide an interesting alternative to the parabolic reflector. For example a helix having 69 turns has a gain of about 20db at 10 Gc/s.

CONSTRUCTION OF AERIALS

Various constructional points have been mentioned in the text of this chapter but certain points are summarized and some are added here.

Great care must be taken to prevent the ingress of water to the coaxial cable at the feed point, Bostik sealing compound or Telecompound is recommended for this. The use of dissimilar metals in contact without weather protection must be avoided otherwise rapid corrosion due to electrolytic action will result. This is especially true of brass and aluminium. Even where protected it is still advisable to avoid dissimilar metals in contact if possible. Elements may be secured to the mounting booms by the use of proprietary cast alloy saddles supplied for television aerial use which are commonly available at a reasonable price. Similar clamps are available for attaching booms to masts. If the aerial is to be used out of doors, as most are, it is very desirable that it should be painted using a suitable primer; zinc chromate is ideal for aluminium and brass (adequate pre-cleaning with detergent or sugar soap is essential) with one or two coats of aluminium paint for finishing. Aerials protected in this way will last for ten years in a corrosive atmosphere whereas, unprotected, especially if made of some aluminium alloys, e.g. magnesium/aluminium, will only last a year or two or even less.

Certain alloys have, however, an inherently high resistance to corrosion. It is essential to have the specification of the material in detail and to have access to the manufacturer's data on corrosion before deciding on an alloy which will give a satisfactory life in an unpainted condition.

Particular care should be taken to ensure that all mechanical and electrical joints—especially sliding ones—are really good otherwise the aerial array will be unlikely to survive the first high wind.

TUNING ADJUSTMENTS

To tune up any aerial system it is essential to keep it away from large objects, such as buildings, sheds and trees. The array itself should be at least two wavelengths above the ground. It is useless to attempt any tuning indoors since the change in the surroundings will result in completely different performance when the array is taken outside.

Undoubtedly the most effective apparatus for tuning up any aerial system is a standing-wave indicator or reflectometer (see Chapter 19—*Measurements*). If there is zero reflection from the load, the standing-wave ratio on the aerial feeder is unity. Under this condition, known as a *flat line*, the maximum power is being radiated. All aerial matching adjustments should therefore be carried out to aim at a standing-wave ratio of unity.

If suitable apparatus is not available, the next best course of action is to tune the aerial for maximum forward radiation. A convenient device for this is a field-strength meter comprising a crystal voltmeter connected to a half-wave dipole placed at least ten wavelengths from and at the same level as the aerial. When adjustments have resulted in a maximum reading on this voltmeter, the feed line may nevertheless not be " flat " and therefore some power will be wasted. However, if the best has been done with the resources available it is highly likely that good results will be achieved.

TYPICAL AERIALS

To illustrate the principles of construction three aerial arrays which are typical of those in use at the present time are described below.

Five-element Yagi Arrays for the 4m and 2m Bands

The five-element Yagi array shown in **Fig. 14.36** is a typical " flat top " for v.h.f. use. With the dimensions indicated the centre frequencies are 70·3 and 145 Mc/s.

The boom is made from wood and the elements are supported on polythene insulators mounted on wooden cross pieces. This form of construction is suitable where the workshop resources are limited and is quite as satisfactory as an all-metal construction although rather heavier.

The driven element is a folded dipole fed with 72 ohm coaxial cable through a balun of the Pawsey-stub type. This balun is constructed from brass tube, the inside diameter of which is a sliding fit over the outer covering of the coaxial

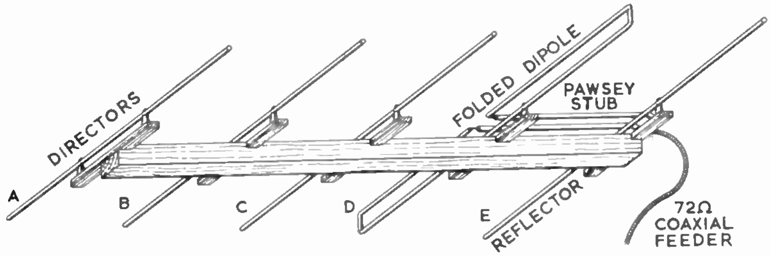


Fig. 14.36. 5-element Yagi arrays for 70·3 and 145 Mc/s with a folded dipole radiator. The gain obtainable is about 9db.

TABLE 14.14

Element Lengths			Spacings		
	70·3	145		70·3	145
A	72 in.	35 in.	AB	25 in.	12½ in.
B	73 in.	35½ in.	BC	25 in.	12¼ in.
C	74 in.	36 in.	CD	25 in.	12¼ in.
D	79 in.	38½ in.	DE	42 in.	20½ in.
E	82½ in.	40 in.			
			balun length L	41 in.	19¼ in.

Materials

Elements: Aluminium rod or tube ⅜ in. dia. (70·3), ¼ in. dia. (145)
Folded Dipole (D): folded part ⅜ in. dia.
fed part ¼ in. dia.
centre-to-centre spacing 1 in.
Balun: brass tube to be a sliding fit over the coaxial feeder cable.

cable. The cable braiding is soldered at the aerial end of the balun to the outside of the balun tubing.

The gain of the aerial is about 9dB, and the beamwidth is about 50° between half-power points.

A 6-over-6 Skeleton-slot Yagi Array for the 2m Band

This array can be constructed entirely from aluminium tubing, the elements being attached to the booms with suitable clamping devices.

With the dimensions given in **Fig. 14.37**, the centre frequency is 145 Mc/s and the bandwidth is approximately 3 Mc/s for a standing-wave ratio of 1·2 : 1. The beamwidth for the half-power points is 45° and the forward gain approximately 11db. The feed impedance is 72 ohms and good-quality 72 ohm coaxial cable should be used as the feeder. A balun is not essential, but it is recommended that one should be used and may be either the Pawsey-stub or coaxial-sleeve type. For convenience in mounting this balun on the aerial mast, a short length of 72 ohm balanced-twin feeder may be used to connect the output of the balun to the feed point. To prevent the ingress of water at the top end of the coaxial cable, the whole balun assembly should be enclosed in polythene film.

Provided the dimensions given are strictly adhered to, no matching adjustments will be necessary.

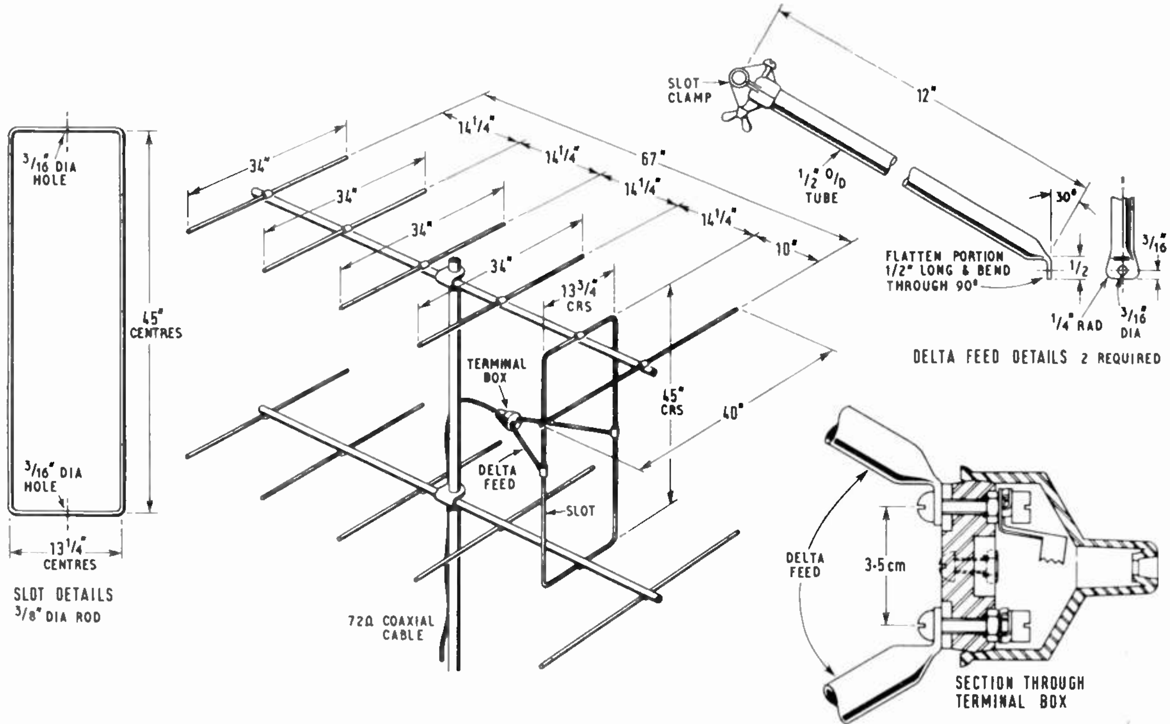


Fig. 14.37. A 6-over-6 skeleton slot Yagi array suitable for 145 Mc/s. Note slot/reflector spacing should be 20 in. not 10 in. as shown.

With acknowledgements to J-Beam Aerials Ltd.

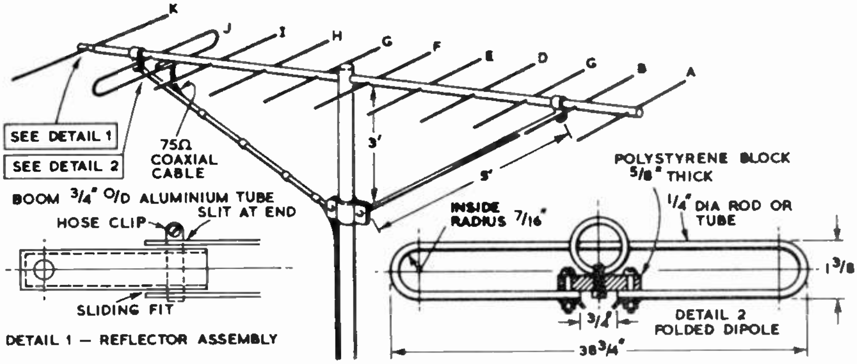


Fig. 14.38. 11-element Long Yagi for 145 Mc/s, gain 13·5db. The gain may be increased by adding additional directors, each ⅛" less in length than its predecessor. The folded dipole is fed using a balun as shown in Fig. 14.39.

Eleven Element Long Yagi for the Two Metre Band

The 11 element long Yagi shown in Fig. 14.38 is light in weight yet extremely effective.

The boom is drilled for the ⅛ in. diameter aluminium rod elements and the ¼ in. diameter folded dipole using a home workshop drill and drill stand located on the floor. The boom must be prevented from turning, this is probably best done by clamping it to a wooden stand or board. It is most important to ensure that the holes are drilled at right angles to the boom so that the elements are parallel to each other when fitted.

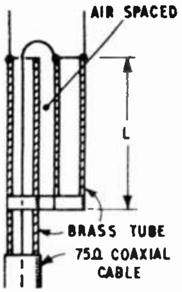


Fig. 14.39. Balun of Pawsey-stub type which can be used for feeding aerials in Fig. 14.36 and Fig. 14.38. Length L —41 in. for 70·3 Mc/s and 19¼ in. for 145 Mc/s respectively.

Elements are either secured to the boom, having been inserted through the holes, by centre punching the aluminium boom around the holes thus forming a compression joint for each element or preferably a clip of the type shown in Fig. 14.40. The folded dipole must have one end bent after insertion in the boom. This can be done by forming the shape around a ⅞ in. diameter broom handle. The boom is secured to the mast by a mast clamp whilst the stay rods are secured by aluminium straps. The centre of the folded dipole is attached to a ⅝ in. thick polystyrene block which is drilled and tapped for the securing screws which are also the feed point. This block is screwed to the boom.

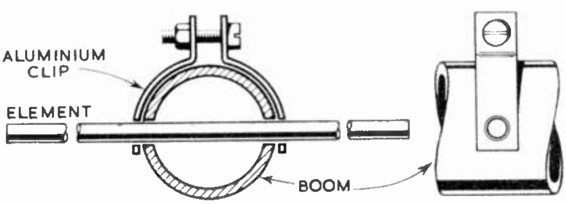


Fig. 14.40. Element securing clip.

The folded dipole/feeder connections are waterproofed with Bostik sealing compound. Matching to the feeder is optimized, preferably using a reflectometer or, if not available, for maximum forward radiation, by sliding the reflector assembly in and out of the boom. The resulting match is better than 1·2 to 1 at 145 Mc/s. If greater gain is required additional directors may be added and the resulting slight change in matching can be effected by movement of the reflector as before. Finally, the reflector assembly should be clamped rigidly inside the boom by means of a hose clip.

TABLE 14.15

Feed details: as for the 5-element Yagi arrays for 70·3 and 145 Mc/s.

element	A	B	C	D	E	F	G	H	I	J	K
element length	35¾	35⅞	36	36⅛	36¼	36⅜	36½	36⅝	36¾	detail 2	41
Parasitic element ⅛" dia aluminium rod	AB	BC	CD	DE	EF	FG	GH	HI	IJ	JK	
element spacing from	16	16	16	16	16	16	16	20	8	19	

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CHAPTER 15

NOISE

IN every receiver there is a continuous background of noise which is present even when no signal is being received. This noise is caused partly by the valves and circuits of the receiver and partly by electrical impulses which reach the aerial from various sources; i.e. it is partly *internal* and partly *external*. In general, the more sensitive the receiver the higher the noise level, and increasing the sensitivity beyond a certain point brings no further increase in range. It is then possible to receive at adequate strength any signal not submerged in the noise, and further amplification will be of no assistance since both noise and signal will increase together. The sensitivity is then described as being *noise limited*, and this is the normal condition in modern receivers intended for weak-signal reception.

The purpose of this chapter is to discuss the characteristics of the various kinds of noise and to outline the broad principles of receiver design which must be followed if a high performance is to be achieved. Detailed treatment of special noise suppressors or noise limiters for reducing interference from ignition systems and other impulse generators will be found in the various chapters dealing with receivers.

An account is given here of the general characteristics of noise, the way it affects reception, the various sources of noise, and the principles involved in reducing its effects to a minimum. Finally the application of this knowledge to the needs of the average amateur is discussed, and it is hoped that this may prove of value to those who have not the time or are not sufficiently interested in the " whys and wherefores " to read the earlier parts of the chapter.

THE NATURE OF NOISE

The noise contributed by the receiver can be minimized by careful design, but it is important to appreciate that internal noise tends to increase with frequency whereas noise from external sources usually decreases. By careful design receiver noise can usually be made negligible compared with external noise below about 100 Mc/s. Internal noise is caused by the unavoidable random movement of electrons

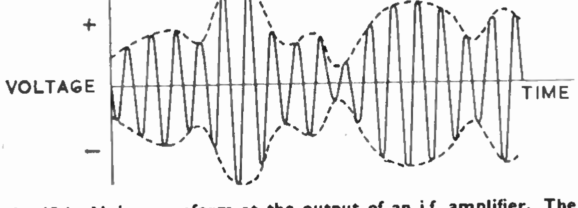


Fig. 15.1. Noise waveform at the output of an i.f. amplifier. The oscillations, which are at the intermediate frequency, are approximately sinusoidal but have randomly varying amplitude.

in valves and circuits, and occasionally by avoidable defects such as dry joints in the wiring, faulty valves, and partial breakdowns of insulation. External noise can be caused by electrical machines and appliances (*man-made noise*), by atmospheric electrical discharges (*static* or *atmospherics*) and by radiations from outer space.

Most forms of electrical noise can be roughly pictured as a mixture of alternating currents of varying amplitudes and of all possible frequencies. Receivers, however, have a limited bandwidth and the observed noise level is a summation of all the components occurring within this band. As might be expected from this picture, the mean noise power is proportional to bandwidth, although at any given instant the noise amplitude can have any value depending on the extent to which the various components happen to be assisting or cancelling each other. **Fig. 15.1** shows a typical noise waveform such as might be produced at the output

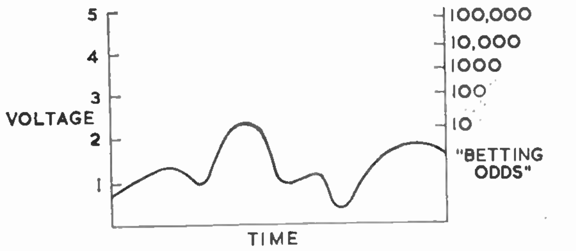


Fig. 15.2. The same noise waveform after linear rectification. Note that it consists of the envelope of the original noise oscillations. The waveform as drawn has an average voltage amplitude of 1 volt and the " betting odds " against any particular instantaneous amplitude being exceeded are indicated by the right-hand scale: this applies to " white noise " only.

of an intermediate-frequency amplifier by random movements of electrons in early stages; the i.f. amplifier can only transmit frequencies within its passband, and the noise applied to the detector has therefore the character of an i.f. signal of fluctuating amplitude. **Fig. 15.2** shows the same noise after detection by a linear rectifier; this consists of a positive (or negative) voltage which fluctuates in accordance with the amplitude of the input noise voltage. The important features of this waveform are its mean level, its statistical properties (summed up by the right-hand scale on the diagram) and its rate of fluctuation which is such that if the bandwidth of the circuits preceding the detector is B cycles per second it requires a time $1/B$ seconds for the amplitude to change significantly in value.

A steady carrier or the output from the beat-frequency oscillator when applied to the detector produces a steady d.c. output voltage V, on which the noise waveform is

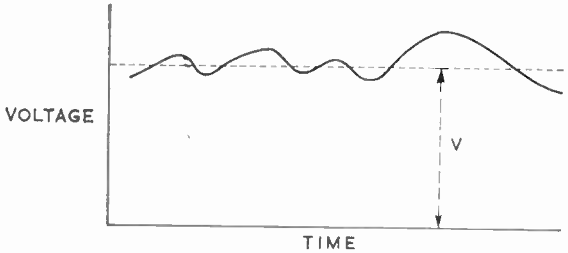


Fig. 15.3. Output of linear detector when the input consists of a carrier (of voltage *V*) plus noise. Note that the average voltage remains equal to *V* as long as the noise peaks are less than *V*, i.e. as long as the noise-modulation of the carrier is less than 100 per cent.

superimposed as shown in **Fig. 15.3**. The noise adds or subtracts from this voltage depending on its relative phase which can have any value and, like the amplitude, tends to change from one value to another in time-intervals of the order of $1/B$ seconds. The noise fluctuations are now symmetrical in the sense that they leave the average voltage V unchanged.

One common effect of the carrier is to affect the apparent pitch of the noise; thus the waveform of Fig. 15.2 can be thought of as being produced by noise frequencies up to a bandwidth of B cycles per second beating together, while that of Fig. 15.3 is due (in the case of normal double side-band reception of A3 signals) to a carrier in the centre of the passband beating with noise frequencies out to a distance of only $B/2$ away from it on either side. This causes a lowering in pitch of the noise at the carrier frequency and a rise in pitch when the carrier or the b.f.o. is detuned as for s.s.b. or single-signal c.w. reception. This behaviour is illustrated in **Fig. 15.4**. With a little practice this rise in pitch can be used as a method of setting up the b.f.o.

Frequently the presence of the carrier or the output from the b.f.o. increases the noise level, and this is commonly mistaken for noise on the carrier. The correct explanation is that when the signal level is low all detectors operate in a square-law manner, which is relatively inefficient; frequently the noise level is too low to cause the detector to operate on the linear part of its characteristic, so that the increase in noise level when the carrier is applied indicates more efficient detection and not a noisy carrier.

Noise originating externally to the receiver is often much more impulsive in character than the internal noise depicted in Fig. 15.1. It can have a wide variety of characteristics, which makes it difficult to discuss in general terms, but the representation of Fig. 15.1 can be applied to it, bearing in mind that (i) the dynamic range may be much greater and the " betting odds " of Fig. 15.2 are not applicable, (ii) the envelope may show much slower variations, and (iii) there may be gaps in the noise waveform which can be exploited for reception particularly when, as is frequently the case in amateur working, low-grade communication is better than none.

THE EFFECT OF NOISE ON RECEPTION

The various random voltages originating in the receiver, together with certain kinds of externally-generated voltages, give rise to what is popularly known as *white noise*. It is

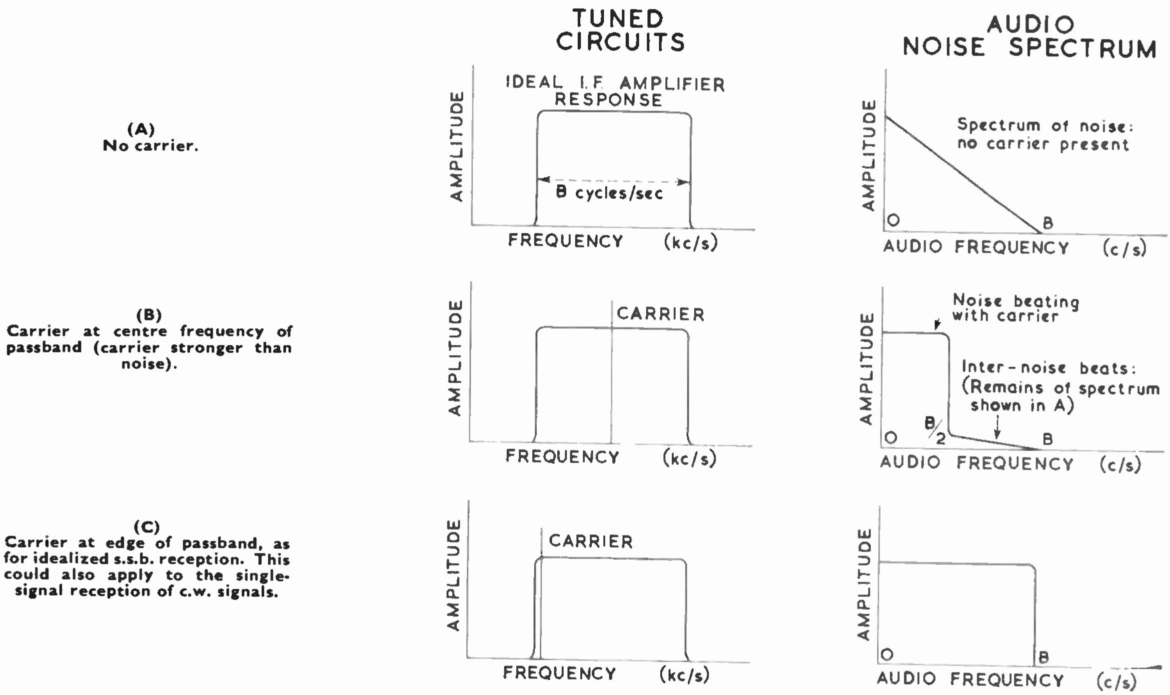


Fig. 15.4. Frequency characteristics of noise in various receiving conditions. When the frequency of the carrier (or the b.f.o.) is at the centre of the passband the noise-pitch is at its lowest, and when it is tuned to the edge of the passband the noise-pitch is at its highest. The band-width of the a.f. amplifier is assumed to be at least as great as the i.f. bandwidth (B cycles/second).

described in this way because, like white light, its energy is uniformly distributed over the relevant frequency-band. The manner in which such noise interferes with the reception of a signal is governed by certain elementary principles. Some discussion of these principles may prove interesting and helpful, although as will be seen their application to practical problems is not always straightforward.

It is useful to start by considering the nature of the signal. Any signal waveform can be considered as being built up from a succession of impulses, and it is obvious that the more rapidly the voltage constituting the signal changes, i.e. the higher the frequencies in its modulation envelope, the shorter the impulses which must be used to obtain a reasonably accurate representation of it. A c.w. signal, for example, could be built up from impulses equal in length to a Morse dot, say 0·1 second, whereas a telephony signal having frequencies up to 3 kc/s would need to be regarded as a succession of impulses not more than 1/6000 second in length. If t is the duration of an impulse, an i.f. bandwidth of $1/t$ is needed for reproducing it, i.e. 6000 cycles per second in the case of a telephony signal. The use of the term *i.f. bandwidth* avoids any possible confusion due to factors arising in the detection process, such as the ratio of two between the overall and i.f. noise bandwidths which is illustrated in Fig. 15.4 (B), but it will be shown later that such distinctions are not important during the actual reception of normal amateur signals when these are interfered with by noise only. The r.f. bandwidth is usually much greater than the i.f. bandwidth and can therefore be ignored.

The optimum bandwidth must be in the region of $1/t$), since this value allows the wanted impulse to build up almost to full amplitude and a greater bandwidth can only increase the effective amount of noise and not the signal. On the other hand, a smaller bandwidth reduces the amplitude of the signal more rapidly than that of the noise.

The noise in a bandwidth of $1/t$ also consists of impulses of duration t, but these vary in amplitude and phase in a random manner. Each impulse, before detection, normally contains a considerable number of cycles of r.f. or i.f. oscillation, like those shown in Fig. 15.1. When a waveform is represented by impulses in this way, the phase and amplitude are regarded as being constant for the duration of any one impulse but there need not be any similarity between consecutive impulses. It will be appreciated that this is merely a convenient approximation, since in practice there is usually a smooth transition from one value to the next.

The size of any one noise impulse is a matter of chance, somes values being more likely to occur than others, and Fig. 15.2 shows the "betting odds" against any given impulse exceeding the indicated values. To see what this means in practice, consider the reception of a single Morse dot, the bandwidth being just wide enough and no more; Fig. 15.2 shows that there is a reasonable chance of the noise amplitude reaching, say, three or four times the r.m.s. value, and unless the signal is even larger in amplitude it is likely to be mistaken for noise or vice versa. It is impossible to detect with certainty the presence of the Morse dot in these circumstances either aurally or mechanically or by any means whatsoever unless it is stronger than any likely value of the noise. In aural reception, however, with the stipulated bandwidth (about 10 c/s), the noise has a rough musical pitch similar to the wanted beat-note, resulting in an unpleasant overall effect, and the discriminatory powers of the ear are not used to best advantage. Under normal conditions it is found that the bandwidth can be extended indefinitely, and although the noise level rises the ability to copy a given weak signal through it is unaffected. This is because the ear itself, from the point of view of discrimination between signals and noise, acts as a filter having a passband of about 50 c/s and thus tends to determine the overall receiving bandwidth. The noise which is effective in masking a wanted 1000 c/s beat note is mainly that part of the noise spectrum between 975 and 1025 c/s, and adding noise outside these limits has no adverse effect at normal audio levels. The usefulness of aural selectivity is not restricted to the separation of signals from noise, and most readers will be familiar with the fact that a weak Morse signal can be read in the presence of a much stronger unwanted signal when the beat notes are well separated in pitch, whereas two signals producing nearly the same heterodyne frequency are much more difficult to copy.

It has been explained above that the problem of separating weak signals from a noise background is one of deciding at each instant whether a true signal or only noise is present at the output of the receiver. This is a useful way of visualizing the basic problem, but in aural or visual reception the process is not carried out consciously and deliberately in the manner stated, and a variety of subjective or intellectual influences are usually at work. These are dependent partly on "integration" and partly on the fact that large portions of most messages are redundant so that missing words and phrases can often be filled in simply because nothing else would make sense. A missing dot in a Morse transmission may be serious if it occurs in a call-sign, but in a group of figures, known to contain no letters (e.g. a contest number or RST report), the figures can be correctly deduced by suitable reasoning even if several of the dots are missing.

Integration is the process of adding together several signal impulses to give one larger one. This is mainly the preserve of radio astronomers and radar engineers, but it also forms part of the subjective processes which are called into play while an attempt is being made to decipher a weak signal aginst a noise background.

Because of these and other complexities (e.g. fading) it is difficult to give any useful figures of signal-to-noise ratios required for amateur communication: more-or-less agreed figures exist for various types of commercial service, but amateur requirements tend to be both more modest and more elusive. The ability to read signals through white noise is not possessed in the same degree by all operators, but in general the human ear is a very efficient device for this purpose. When, as in amateur communication, messages are in plain language and tend to have standardized content such as RST reports, names and greetings, the aural selectivity can be supplemented by discrimination of an intellectual kind and no improvement is obtainable by the use of additional filters or other devices. It must be emphasized, however, that this applies only to white noise. It further assumes the absence of overloading in the receiver, adequate b.f.o. injection voltage at the detector and a reasonable amount of audio gain. At very high volume levels the ear has a non-linear characteristic, reception is degraded and, if the bandwidth is increased at constant gain, the noise level will be further raised and the situation will be aggravated. On the other hand, if the volume is too low, signals comparable with the noise level may not be loud enough for good intelligibility. In the case of impulsive noise (e.g. atmospherics and ignition noise) the ear can often be given considerable assistance by devices such as limiters which prevent overloading and enable a signal to be copied through gaps in the noise (see Chapters 4 and 16).

An understanding of the mechanism of detection (including the effects shown in Fig. 15.4 and the relatively complex phenomenon of modulation suppression) is important in the solution of many noise problems and it is also helpful, for example, in preventing the drawing of false conclusions from the way in which the amplitude and the pitch of the noise vary during the tuning-in of a signal. However, in regard to the actual copying of signals of the kinds met with in amateur radio, the problem can be very much simplified; this is because during such reception the detector is always operating *as a mixer*, i.e. there is always present a continuous signal or carrier which is strong compared with the noise, and the detector output is the beat frequency between the wanted intelligence and this strong oscillation. So long as this holds true, the detector, like any other frequency changer, is merely a device for shifting the frequency spectrum; it has no other effect on the signal or the noise, and no distinction need be made between post-detector and pre-detector selectivity, except where a factor of 2 must be introduced to allow for the superimposing of two similar sidebands in the process of detection, as indicated in Fig. 15.4 (B). This argument breaks down if, in a.m. reception, the carrier is not strong compared with the mean noise level, but normal speech modulation will then be unreadable in any case.

INTERNAL NOISE

Noise originating in the receiver can usually be distinguished from external noise by the fact that it persists after disconnecting the aerial, but this test can be misleading since the change in damping or tuning of the first circuit may alter the internal noise level, and it is better to make the experiment by changing over from the aerial to an equivalent dummy aerial. In the h.f. band, where external noise levels are high, this gives a simple and accurate check of the noise performance of the receiver; thus a change of 6db (i.e. a ratio of 4 : 1) in the noise level would mean that the receiver was responsible for a quarter of the total noise power, and the overall performance would be about 1db worse than with a perfect receiver. These figures, of course, assume linear detection and no overloading in the receiver.

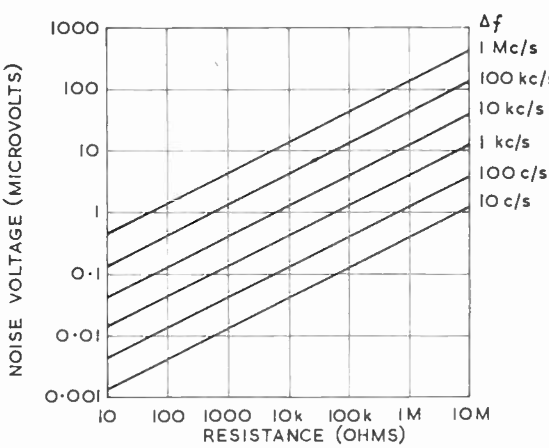


Fig. 15.5. Chart for determining the noise voltage developed in a resistance for a specified bandwidth Δf at room temperature (20° C). Note that the scales are logarithmic.

Provided that no part of the internal noise is due to faulty wiring or components, it can be roughly divided into

(a) thermal noise,
(b) shot noise in valves, and
(c) induced grid noise in valves.

Whichever variety of noise is being considered, it is of practical importance only when generated in the early stages of the receiver, since noise generated later in the receiving chain is not amplified to the same extent.

Thermal Noise

The electrons in a conductor are in continuous random motion at a rate which increases with the absolute temperature (obtained in degrees Kelvin by adding 273 to the temperature in degrees Centigrade). The voltages produced by the electrons add together like other a.c. voltages, i.e. as the square root of the sum of the squares, and therefore if two similar conductors having equal resistances are joined in series the resultant noise voltage produced is $\sqrt{2}$ times that developed across either of the conductors considered singly. For any given conductor the noise voltage corresponding to a bandwidth of B cycles per second is equal to $\sqrt{(4KTBR)}$ where K is Boltzmann's constant ($1{\cdot}37 \times 10^{-23}$ joules per deg. K), T is the absolute temperature (in deg. K) and R is the resistance of the conductor (in ohms). This relationship is plotted graphically in **Fig. 15.5**, but for quick reference a convenient figure to remember is 0·13 microvolts (approximately) for 1000 ohms at 1 kc/s bandwidth and at room temperature. The thermal noise generated in a tuned circuit can be calculated from the same expression if R is taken to represent the dynamic impedance: thus for a parallel resonant circuit R should be substituted by $Q\omega L$: see Chapter 1—*Principles*.

Thermal noise is seldom the most prominent form of noise. In h.f. reception it is overshadowed by noise picked up by the aerial, and in v.h.f. reception by valve noise.

The concept of noise temperature is helpful in assessing and comparing noise from all sources; thus it is now customary to express the external noise level by assigning an *equivalent temperature* to the radiation resistance of the aerial, i.e. the temperature it would be required to have if it were to produce this amount of noise merely as a consequence of thermal agitation. Similarly, valve shot noise is expressed in terms of an *equivalent* (room temperature) *noise resistance*, and induced grid noise in terms of an equivalent temperature attributed to the resistive component of the valve input impedance. The advantage of these concepts becomes apparent in dealing with *noise factor*, discussed later in this chapter.

Valve Shot Noise

This can be pictured by thinking of the anode current as being due to electrons hitting the anode like a stream of shot, or to the impact of hail against a window-pane, rather than as a continuous smooth flow. If the anode current I flows through a resistance R, the noise voltage developed is $\sqrt{(2eIBRF^2)}$ where e is the charge on an electron and F^2 is the space-charge smoothing factor. Comparison of this formula with the thermal-noise formula is the basis of the noise-diode method of measuring receiver noise performance, F^2 being unity when, as in the noise diode, all available electrons are drawn to the anode. For multi-electrode valves operating normally, F^2 is usually about 0·15 and it is convenient for comparison with aerial and thermal noise to represent the valve noise by a fictitious *equivalent noise resistance* (R_{eq}) such that a noise voltage $\sqrt{(4KTBR_{eq})}$ applied between grid and cathode would produce the equivalent fluctuation of anode current. It should be noted that the equivalent noise resistance is only a mathematical fiction and not a true resistance to which Ohm's Law could be applied.

Formulae for the calculation of the values of R_{eq} are given in Chapter 2—*Valves*.

Induced Grid Noise

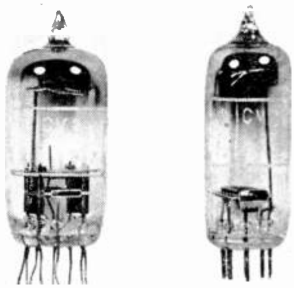
At low frequencies the voltage on the grid of a valve due to an applied signal can be considered as being constant for the complete interval of time taken for one electron to leave the cathode space charge and arrive at the anode. As the frequency is raised this state of affairs ceases to exist and in consequence a current is induced in the grid circuit giving the effect of a resistor, which decreases as the square of the frequency in shunt with the grid circuit. This resistor also appears as a generator of thermal noise at an equivalent temperature comparable with that of the valve cathode, i.e. about five times the absolute room temperature. There is partial correlation in phase between the shot noise and the induced grid noise with the result that one can be partially balanced against the other by detuning the grid circuit and thereby altering the relative phases. In general the reduction in noise is small and the two sources can be treated as independent to a first approximation, although it is often found that the signal-to-noise ratio tends to vary in an unsymmetrical way as the tuning of the input circuit is varied.

AERIAL NOISE

If a receiver is connected to a signal generator or to a dummy aerial a certain thermal noise voltage is fed into the receiver. This so-called *aerial noise* would determine the minimum usable signal level if the receiver itself generated no noise and if the real aerial produced the same noise as the dummy aerial. It has been found convenient to use the thermal noise level of the dummy aerial as a basis for the definition of *noise factor* referred to later in this chapter, because in this way the definition is directly related to the normal conditions of measurement. Unfortunately the noise levels of real and dummy aerials are usually quite different, so that both the definition and the measurements are unrelated to actuality and can be very misleading unless the appropriate allowances are made.

When the impedance of a real aerial is measured, for example, with an impedance bridge, it is found to consist of resistance and probably some reactance as well. The reactance can be ignored for the purpose of the present discussion since it is usually tuned out in one way or another. The resistance consists partly of radiation resistance and partly of loss resistance. The loss resistance is made up of the ohmic resistance of the aerial wire, leakage across insulators, losses in the immediate surroundings of the aerial, and also includes the resistance of the earth connection when a Marconi aerial is used; it is generally important in the case of l.f. reception and when using indoor and frame aerials, and is one of the two factors (the other being the bandwidth) which determine the extent to which it is practicable to reduce the size of an aerial system. The loss resistance is a source of thermal noise corresponding to a temperature which is usually taken as 300°K although it can of course vary by 10 per cent or more from this figure depending on the local climatic conditions. In the higher

Typical noise diodes. Left, the CV2398 suitable for use at frequencies well above 500 Mc/s; right the A.2087 (CV2171). Noise generators employing these valves are described in Chapter 19—Measurements.



frequency bands the loss resistance can normally be made negligible compared with the radiation resistance, which in practical cases can be considered to have any noise temperature from less than 10°K up to thousands of millions of degrees depending on the frequency and other factors.

Radiation Resistance

There is no connection between the noise associated with the radiation resistance of an aerial and the actual physical temperature of the aerial or its surroundings, and to appreciate why this must be so it is necessary to understand the meaning of radiation resistance. When power is fed into a load, it is absorbed in the resistance of the load, and this is just as true for a transmitter feeding power into the radiation resistance of an aerial as it is for any other generator and load, but as is well known power radiated from an aerial may travel immense distances, even to outer space, before being absorbed. The radiation resistance of an aerial therefore tends to be a property of the whole of space of which the local features occupy merely a trivial part. The radiation resistance as such depends only on the fact that power is dissipated, regardless of where this takes place, but the associated thermal noise level depends on the nature of the absorbing objects and not on the actual temperature of the aerial.

EXTERNAL NOISE

The main sources of noise external to the receiver which affect communication may be grouped under three headings:

(a) man-made,
(b) atmospheric and
(c) cosmic.

During periods of intense solar disturbance, radio noise from the sun also tends to interfere with normal reception.

Man-made Noise

In remote rural areas, the level of man-made noise may be pleasantly low or even negligible, whereas in congested towns it may be intolerably high. The various kinds are often identifiable by reason of their time characteristics. Most of them originate in domestic electrical appliances or in industrial equipment. Common sources of this kind of noise are high voltage power lines, light switches, vacuum cleaners, refrigerator motors, advertising signs, electric shavers, electric drills, petrol engines, trolley buses and electric trains.

Intermittent clicks, such as those produced by switches and well-designed thermostats—unless very frequent—do not cause serious trouble since isolated noise impulses can only destroy occasional short elements of the signal and comparatively little of the information content is lost.

Continuous interference, such as that caused by commutators or discharge-tube lighting, is much more damaging to good communications. Such interference spreads over a very wide frequency range, although some fluorescent tubes generate a broadly tunable noise band of unstable frequency.

The type of noise associated with commutators, often known as *hash*, is only partly impulsive in character and cannot be dealt with successfully by any noise-limiting device in the receiver. The true impulse type of noise, such as ignition interference, is much easier to suppress. One of the advantages of the frequency-modulation system is that if the receiver is correctly designed and adjusted, interference from high-amplitude impulse noise is virtually eliminated, though a similar result can be achieved in the reception of a.m. signals if sufficient i.f. bandwidth and appropriate limiters are used. Individual sources of man-made noise usually have a relatively small range, and considerable improvement may be obtained by raising the aerial system as high as possible and by using a screened feeder. In the lower h.f. range, however, man-made noise from the large numbers of distant sources tends to be a limiting factor even in districts remote from centres of population.

Atmospheric Noise

The primary source of atmospheric noise (or *static*) is the ordinary lightning discharge. A typical flash has been estimated as carrying a current of 20,000 amperes. At any one time there are so many storms in progress in various parts of the world and the range of radio propagation is so great that much of the external noise which forms the background in long-distance h.f. communication must be attributed to lightning discharges in remote storms. The noise from local storms is naturally much more intense and is considerably greater on the lower radio frequencies, but of course this is only intermittent and does not require to be taken into account in the design of circuits.

A not infrequent, but usually short-lived, form of interference is due to rain static, i.e. currents induced in the aerial by electrically charged raindrops.

Cosmic (Galactic) Noise

Considerable noise radiation is received from the Milky Way and other galaxies, some of which are so remote that they are not visible even in the most powerful optical telescopes. At the wavelengths mainly used for long-distance communication, cosmic noise penetrates the ionosphere and is received at a high amplitude level. Apart from the variation with frequency, cosmic noise is similar in character to thermal noise.

TOTAL NOISE LEVEL

A general idea of the variation of atmospheric and man-made as well as cosmic noise level with frequency over the h.f. and v.h.f. bands is given in **Fig. 15.6**. These curves have been somewhat freely adapted from previously published data, bearing in mind the need for a simplified presentation. Below 14 Mc/s the diagram should be regarded only as a rough guide, since atmospheric and man-made noise levels are extremely variable, and the noise is generally more impulsive in character. As explained earlier, impulsive noise tends to be relatively less serious the lower the grade of communication, and for amateur purposes the *effective* lower-frequency noise levels may be lower than indicated although some slight "weighting" in this direction has taken place in the course of adapting the data.

The marked difference in atmospheric noise levels between day and night is due to the more efficient propagation at night from distant storm centres located mainly in the tropics. Above 100 Mc/s the external noise level continues to decrease, being much less than room-temperature thermal noise in the 430 Mc/s band, and frequently equivalent to only a few degrees absolute in the microwave bands.

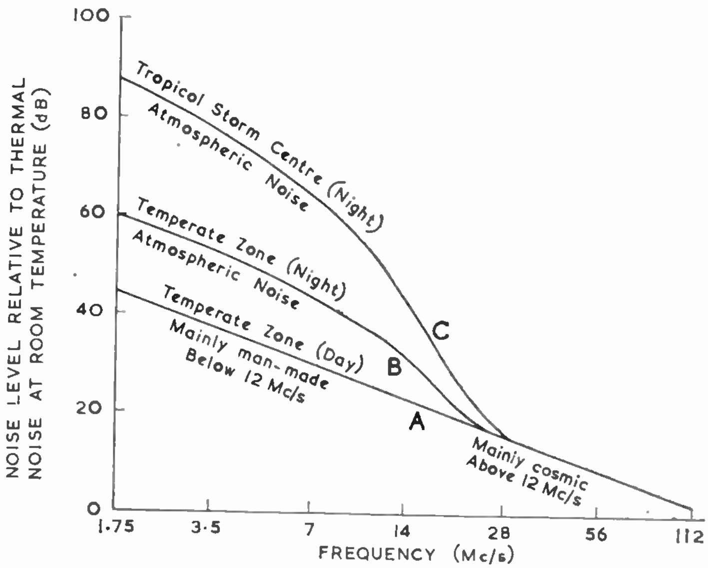


Fig. 15.6. Variation of external noise level with frequency. Curve A shows that during the day in temperate zones the noise is mainly man-made at frequencies above about 12 Mc/s. In these zones, atmospheric noise adds considerably to the total noise level at night (curve B). In tropical zones the atmospheric noise is relatively severe: curve C represents the worst conditions in these zones. The vertical scale indicates the number of decibels by which the noise level in a perfect receiver would increase if it were disconnected from a dummy aerial and fed from an efficient aerial of similar impedance.

ESTIMATION OF RECEIVER NOISE

When it is desired to measure the sensitivity of a receiver, some form of test signal is fed in through an impedance which simulates that of the aerial from which the receiver is designed to work. Such an impedance is known as a *dummy aerial* and is a source of thermal noise. If the receiver were perfect the noise at its output terminals would consist only of amplified noise from the dummy aerial, but in practice the receiver adds a certain amount of noise as previously explained.

If the aerial noise is represented by T_a degrees (i.e. the equivalent temperature) the increase in noise due to imperfections in the receiver can be represented by T_r degrees (i.e. a rise in the equivalent temperature of the aerial). The value of T_r in comparison with T, the room temperature, is readily measurable by the noise-diode technique described in Chapter 19 (*Measurements*). If T_r is expressed in decibels relative to T, it can be directly compared with the noise levels indicated in Fig. 15.6 to find out whether it is significant in relation to the external noise level. Fig. 15.7 illustrates the lowest values of T_r which can be achieved without undue difficulty, as a function of frequency. As explained later, however, it is not always desirable to aim at a low value of T_r, and h.f. receivers almost always have much higher values than v.h.f. receivers.

Noise Factor

The equivalent-temperature concept of receiver noise is not yet universally adopted and the noise performance of receivers is usually specified in terms of *noise factor*. Unfortunately extreme care is required in specifying the conditions of measurement and in interpreting the figures obtained. However, if the advice offered in Chapter 19 on the subject of noise measurement is closely followed the reader is not likely to arrive at false conclusions. The method of measurement described there is based on the definition—

$$\textit{Noise Factor} = (T + T_r)/T$$

where T is the room temperature. This can of course be arranged in the form—

$$T_r = (\textit{Noise Factor} - 1)T$$

to enable the noise factor as normally measured to be expressed directly as an equivalent aerial temperature. The object of good receiver design is to reduce T_r until it is negligible compared with the aerial noise temperature but not necessarily to make T_r as small as possible; any further possibility of improvement can generally be used instead to improve the receiver in other respects as discussed later, or perhaps to reduce its cost.

To reduce T_r to a negligible value, all that is necessary in principle is to step up the aerial impedance by means of a transformer to a value R_a which is large compared with the equivalent noise resistance R_{eq} of the first valve in the receiver: R_a and R_{eq} can then be thought of as two noise generators connected in series and the value of T_r/T will be equal to R_{eq}/R_a. In practice, however, the aerial impedance is shunted by other noise-generating resistances which represent losses in the input circuit and by the transit-time damping resistance associated with induced grid noise. This resistance has an equivalent temperature of $5T$ and decreases as the square of the frequency. These effects cause T_r to rise with frequency, not only because of the added noise but also because of the shunting effect which reduces aerial noise T_a relative to the valve noise which forms a major part of T_r.

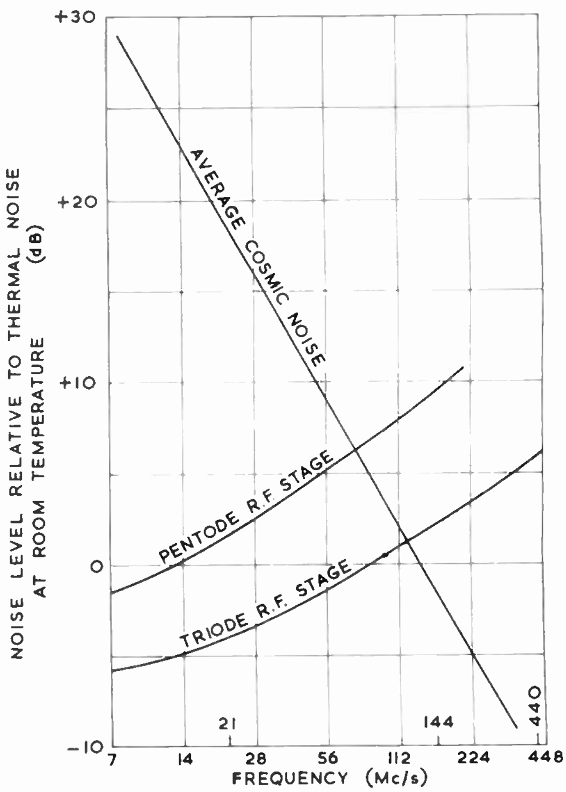


Fig. 15.7. Approximate relationship between aerial noise and receiver noise over the amateur frequency bands. The curves shown for pentode and triode r.f. amplifier stages represent typical good designs.

Since all "temperatures" have to be referred to the aerial where T_a is fixed in accordance with Figs. 15.6 and 15.7, any reduction of T_a relative to T_r is equivalent to an increase of T_r. In order to minimize this effect there is an optimum value of R_a which decreases with frequency and normally corresponds to over-coupling of the aerial into the first circuit of the receiver. This optimum value of R_a occurs because large values are heavily shunted by the transit-time damping so that as R_a is increased from zero the aerial noise voltage at the valve at first rises, as in the no-loss case, and then falls again. For the lowest noise factor, the aerial needs to be more tightly coupled than it would be for maximum gain and there is a consequent loss in r.f. selectivity. It should be remembered that R_{eq} is not an actual resistance but a designation for a noise generator; it therefore does not experience the shunting effects which R_a experiences.

Other precautions to ensure the lowest value of noise factor are to use input circuits with the highest possible inductance and Q-value, and to choose valves with low equivalent noise resistance (i.e. triodes in preference to pentodes) and high transit-time damping resistance. In r.f. or i.f. amplifiers having very great bandwidth where a low noise factor is required it is better to rely on tighter coupling to produce the necessary bandwidth than on the use of shunt resistors.

When inverse feedback is present in the input stage it usually has an equal reducing effect on the signal and on all forms of noise, and therefore does not directly affect the noise factor or the optimum design conditions, although as in the case of the grounded grid triode, it may produce heavy damping of the input circuit. On the other hand, the inductance of the cathode lead in a pentode causes inverse feedback which leaves partition noise unaffected and therefore degrades the noise factor.

PRACTICAL ASPECTS OF NOISE

Every amateur will wish to ensure that the performance of his station is not handicapped by avoidable defects in the receiver. This situation can arise from the following causes:

(a) Gain is not sufficient to make certain that the useful sensitivity is noise-limited.

(b) Receiver contributes significantly and avoidably to the noise background level.

In the case of insufficient gain the trouble and the remedy are both obvious. Those designing their own receivers should have no difficulty in making a rough estimate, with the aid of Figs. 15.5 and 15.6 and the foregoing explanations, of the amount of r.f. and i.f. gain required to give a reasonable working level (say 1 volt) at the detector, and hence of the amount of audio gain needed to bring the rectified noise (about 0·4 volt r.m.s. assuming a diode detector with 1 volt input) up to a reasonable listening level.

Above about 100 Mc/s, low noise factor is all-important and Fig. 15.7 gives a guide to the performance which should be achievable by using a low-noise triode r.f. stage. On the h.f. bands it is the external noise which limits the ability of the receiver to deal with weak signals, provided of course that the aerial is a good one and that the receiver is functioning correctly. In such circumstances there would be no need for any r.f. amplifier stages if a good triode mixer were used and thus the risk of blocking and the production of spurious responses due to the overloading of the mixer by strong signals would be reduced. Another danger is that of an apparent increase in the noise level caused by intermodulation at the mixer between large numbers of moderately strong signals within the passband of the r.f. circuits. All these effects are aggravated by excessive r.f. gain and by some of the other measures which may be taken to imporve the noise factor. On the other hand, if it is necessary to use an inefficient aerial, such as a wire fastened to the picture rail, or a miniature beam having appreciable loss-resistance in the elements or the feeders, the external noise level will in most cases be lower; the receiver noise must therefore be reduced in proportion and an r.f. stage may be necessary. Indoor aerials are sometimes responsible for increased pick-up of man-made noise from local sources, and in such cases this reasoning will obviously not be valid.

In defence of the standard practice of employing r.f. stages in h.f. receivers it must be added that they provide a margin of safety against the consequences of inefficient coupling between the receiver and the aerial. Moreover the disadvantages associated with r.f. stages can often be mitigated by making sure that r.f. gain control is not advanced beyond the point where external noise just swamps the internal noise as judged by a simple check, e.g. by disconnecting the aerial.

Aural Discrimination

When the above conditions have been satisfied, discrimination between signals and noise becomes mainly a job for the ear and brain working together, and except in a few special cases very little can be done in the receiver to assist this process. If the overall bandwidth is greater than necessary, decreasing it reduces the noise level but does not make copying any easier, because it is only the noise which occupies the same acoustic band as the signal that is effective in preventing it from being copied. This argument, however, must not be carried too far; it can, for example, be used to justify the normal practice of using bandwidths of several hundred kilocycles for the sound channel of television receivers, but it is not valid when applied to amateur v.h.f. receivers. The reason for this is that the amateur is able to make use of low signal levels, and it is possible by increasing the pre-dectector bandwidth to reach a point where the noise voltage at the detector becomes comparable with the carrier voltage of, say, a just-readable telephony signal. The situation at the detector is then as depicted in Fig. 15.4 (C) but with the signal over-modulated by the noise, the wanted modulation of the carrier tends to be destroyed. The effect is the same as modulation suppression, a well-known property of linear detectors whereby the presence of a strong carrier destroys the modulation of a weaker one. There is no advantage in using a square-law detector, since this also by its nature discriminates in favour of strong signals.

Effects of Overloading

Another possible cause of deterioration in the ability to copy signals through noise is overloading, particularly of the i.f. stages, by noise peaks which as shown by Fig. 15.2 may reach several times the mean noise level and will increase in amplitude if the bandwidth is increased. If a reduction of bandwidth appears to give improved reception of weak signals, it is possible that some such effect may be taking place and the gain should be reduced.

In the case of impulsive noise like that due to ignition systems or isolated atmospherics, the dynamic range is relatively large and noise limiters are useful in the prevention of overloading, particularly if the bandwidth is large so that the impulses are not lengthened and caused to overlap—as they may be by the ear itself—before being limited. The time intervals between atmospherics are frequently long enough to enable words or syllables, or Morse characters, to be read through the gaps, and it may thus be possible to sustain low-grade communications through a relatively high noise level; in suitable circumstances this may allow some discounting of the lower-frequency noise levels plotted in Fig. 15.6.

S.S.B. Reception

In the reception of Morse and sideband signals, the voltage injected into a conventional diode detector from the beat frequency oscillator must be large compared with the mean noise voltage, but a ratio of 8–10 times in voltage is sufficient in regard to the signal-to-noise ratio, whereas for maximum discrimination against strong signals an even higher injection level is necessary.

It is becoming common practice to employ s.s.b. reception of ordinary amplitude modulated signals in order to remove interference which may be present on one sideband. This unfortunately entails a reduction in the signal-to-noise ratio since the two sidebands of a double-sideband signal add up in phase and therefore the removal of one of them halves the signal voltage; at the same time, the noise power which accompanies the rejected sideband is eliminated. This halves the total noise power so that there is a decrease of 3db in the noise level to offset the decrease of 6db in the signal level, leaving a penalty of 3db. As a corollary to this, it follows that s.s.b., d.s.b. and a.m. signals should give equal signal-to-noise ratios for equal *total* sideband power.

Effects of Aerial Characteristics

The gain of an aerial, in terms of the signal-to-noise ratio, is the same as its transmitting power gain if there are no losses in the aerial and if the noise is non-directional. In relatively rare circumstances the external noise will be stronger from one direction than another, and the effective gain of the aerial will then be reduced for signals in the same direction as the noise and increased for signals in other directions. The removal of high-angle lobes should tend to reduce the external (cosmic) noise level in the 14–28 Mc/s bands. Losses in the aerial and feeder system are subtracted from the transmitting power gain but do not affect the gain in terms of the signal-to-noise ratio as long as the receiver noise can be kept well below the external noise level: thus, referring to Fig. 15.6, the external noise level at 14 Mc/s is 25db so that an aerial loss of 18db would still leave the external noise 7db above thermal noise, i.e. 6db above the receiver-plus-thermal noise for a noise factor of 1db and the signal-to-noise ratio would be degraded only by about 1db compared with that for a loss-free aerial. This permits the use of much more compact aerial systems for reception than for transmission.

A very interesting point arises in the case of rhombic and certain other long-wire aerials, since the terminating resistance absorbs half the transmitter power and, of course, half the received noise power. On transmission the power which is absorbed is that which would otherwise be radiated in the backward direction, and the forward gain is unaffected. The elimination of noise from the backward direction, however, doubles the signal-to-noise ratio provided that the receiver noise level is low enough for advantage to be taken of the lower aerial noise level. Thus, in general, a terminated rhombic operating at frequencies below about 30 Mc/s has an effective gain 3db greater for receiving than for transmitting.

CHAPTER 16

MOBILE EQUIPMENT

THIS chapter is concerned with equipment for mobile and portable operation on the h.f. and v.h.f. bands, and the special problems which arise in relation to interference suppression and aerial systems.

Unlike its fixed station counterpart, a mobile or portable installation will invariably be subjected to physical stress due to vibration or continual handling, or both. For this reason, the physical construction requires as much care as the electrical design.

CONSTRUCTION

Due to space restrictions, miniaturization of equipment designed for mobile operation is essential, but coupled with this must be rugged construction and a high degree of mechanical strength. A chassis braced by sub-chassis partitions, should be at least 20 s.w.g. material or 18 s.w.g when such partitions are only few in number. The front panel, especially when carrying dials associated with variable frequency oscillators, needs to be not less than 18 s.w.g. Each end of the panel should be braced, and joined to the main chassis by means of triangular lipped plates which are fixed to the panel and chassis by bolts or better still rivets as shown in **Fig. 16.1**. It is not satisfactory to rely on the spindle bush and locknut of a control to hold the panel to the chassis.

When bolting a component to the chassis, a shake-proof washer should always be fitted under the nut. When fitting a solder tag to the chassis, shakeproof washers should be used between the head of the bolt and the chassis and between the solder tag and the nut. Nuts should be finally tightened with a suitable box spanner since maximum tightness will not be achieved with a screwdriver. Brass is superior to steel for electrical connections, if steel nuts and bolts are used try to obtain those which are cadmium plated.

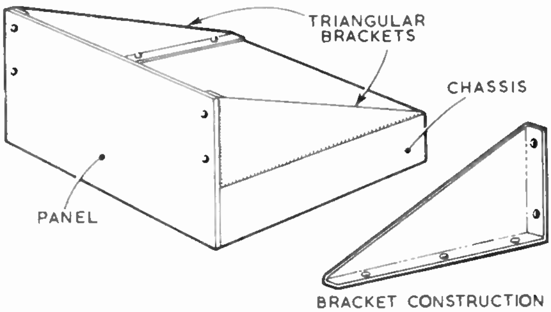


Fig. 16.1 Using triangular brackets to brace both front panel and chassis.

Unlike that of fixed station equipment, the wiring of units intended for mobile working should be undertaken in stranded connecting wire. With its ability to flex, there will be no tendency for the wiring to vibrate under conditions of mechanical shock. Where solid conductors do have to be employed, as in variable frequency oscillators for example, 16 s.w.g. or thicker wire should be used, the lead length being made as short as possible. The connections at each end should be made physically strong before being soldered.

Nearly all equipment for mobile operation has to be tailored to a particular vehicle, or at the very least arranged to occupy the smallest possible space. In addition it has to be operated in a confined space, so the layout and position of the major controls is very important. When planning such an installation thought should be given to how far the driver of the vehicle has to move to reach the various controls. Ideally, this movement should be no more than that needed to operate the gear change lever during normal driving. Two rules need to be applied to controls in general. Firstly, the controls themselves should be restricted to those which are absolutely essential, and secondly, those which have the greatest use should be positioned as near to the driver as possible.

With regard to the actual mounting of the equipment, any form of " soft " mounting is to be avoided. A little resilience can be incorporated, but for all practical purposes, the equipment should be firmly attached to the vehicle body, the mass of which will materially assist in reducing spurious vibrations.

At all times, prime consideration must be given to safety, especially if the installation is to be operated by the driver of the vehicle. It is to the credit of mobile operators that, as far as is known, no accidents have so far been attributed to the operation of mobile radio equipment. Be this as it may, at no time should the slightest risk be taken which is likely to endanger life or limb of either the operator or an innocent bystander.

NOISE SUPPRESSION

A motor vehicle can generate a considerable amount of electrical noise which may cause sufficient interference to prevent satisfactory reception under mobile conditions. Any electrical device fitted to the car in which contacts open and close is a potential source of interference. Even the very small current handled by the impulse contacts of a dashboard clock can produce a ticking noise in a receiver.

Possible sources of electrical noise will now be considered in turn, together with the treatments which should effect a cure. In respect of the remedies, it should be borne in mind that steps taken to suppress noise found on a receiver operating in the h.f. bands may not hold good for a v.h.f. installation. Nearly all the corrective measures involve the use of capacitors which either reduce the arcing at the offending contacts, or which absorb the resultant transient. In most instances fairly large values of capacitance are needed, and, as the frequency rises, so the inductive element of such capacitors becomes more pronounced with the result that, at v.h.f., the capacitor behaves more like an r.f. choke and is thus ineffective as a bypass device.

Unless the current in the circuit being suppressed is small, reducing the value of the bypass capacitor to an amount which is normally effective in the h.f. and v.h.f. range may not have the desired effect since, reducing the value of the capacitor reduces its ability to suppress contact arcing. Ideally, for good suppression at the upper h.f. and in the v.h.f. frequency ranges in high current circuits, coaxial capacitors are used. Examples are the Lucas WS.15 and WS.19 ranges, which are specifically designed for frequencies up to 200 Mc/s. However, in some circumstances a smaller capacitor connected in parallel with the standard suppression unit may be found effective. Where the noise level is found to increase when this course is adopted, it is invariably due to the lower value capacitor resonating with the effective inductance of the larger capacitor, and in such cases experimentation with the value of the lower capacitor may be found helpful. When no reduction in noise can be achieved in this manner, placing a small coaxial feed-through capacitor in the live lead to the large capacitor has been found to achieve a worthwhile reduction. The various effects are illustrated in **Fig. 16.2**.

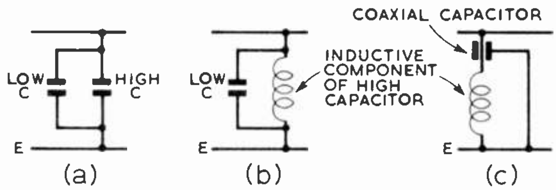


Fig. 16.2 (a) illustrates high and low values of capacitance in parallel in suppression circuits. As the frequency rises the inductive element of the larger capacitor may resonate with the smaller capacitor as in (b) and cause an increase in the noise level. This may be avoided by the use of a coaxial capacitor as shown in (c).

When fitting suppression capacitors, it is of paramount importance that both the earthing and the live leads are extremely short. If the live lead is over 3 in. in length, it should be screened, and the screening earthed at the capacitor end only.

Ignition System

The major source of interference from a petrol engine is the ignition system. The noise from this source rises to its maximum level in the 40–50 Mc/s region, and may there after fall until at 600 Mc/s it is at a low level. As the ignition voltage is of an oscillatory nature, with a peak voltage in the region of 10 kV, severe interference may be created within a ¼ mile radius. In the U.K., statutory requirements are now placed on vehicle manufacturers in respect of noise suppression, and while this may be adequate in respect of TV reception at some distance from the vehicle, and for Medium Wave reception in the vehicle, it is rarely adequate for h.f., v.h.f. and u.h.f. reception under mobile conditions.

Treatment of the ignition system invariably involves the use of resistors in the lead from the ignition coil to the distributor, and in the leads to the individual sparking plugs. Recently, use of resistive high tension cable has been adopted on mass production engines, and while outwardly this appears similar to standard ignition cable, the copper inner conductor is replaced by nylon or a similar material treated with carbon to make it conductive to the desired degree. Such cables are now low in cost, show a marked improvement in suppression over resistor unit elements, and moreover, maintain a high degree of suppression at very high frequencies. Where resistive suppression proves ineffective on a vehicle, consideration should be given to replacing the standard ignition cable with this special type.

With regard to ignition noise in general, it is essential that the sparking plugs are kept clean and the gaps correctly set. As a first step, they should be removed, taken to a local garage and sand-blasted clean, and the gap set to the figure quoted. Generally, plugs cannot be cleaned to a satisfactory degree with a wire brush, and sand-blasting must be employed. The ceramic insulation of all plugs must be spotless, and the use of a proofing and anti-tracking compound such as *Dampstart* is recommended, as is the employment of hooded plug connectors.

The distributor contacts should be unpitted, free from dust or oil, and above all, dry. More than one case has been noted where, despite every possible step in respect of the plugs and their associated leads, very little suppression could be secured but, on replacement of the distributor cap, the interference level fell to an acceptable level.

Once the system itself is up to standard in the respects mentioned, attention can be turned to additional suppression. It is however futile to do so unless the ignition system is in order, for defects in its component parts may give rise to forms of interference which are difficult to reduce.

As a first step, a capacitor of 0·5 μF should be connected from the SW terminal on the ignition coil to a closely adjacent earthing point. This should bring about a noticeable reduction in the interference level. For v.h.f. installations, it is sometimes helpful to bridge this with a disc ceramic capacitor having a value of 0·005 μF. Following this, the cable from the ignition coil to the distributor should be screened, and the screening connected to the same earthing point as the capacitors. Subsequently each plug lead should be screened and the screening connected to a good earthing point. The final step is to fit each sparking plug with an individual suppressor unit integral with the plug connector, but if this course is adopted it is advisable to have the engine tuning checked after fitting.

Generator and Voltage Regulator

The usual generator and voltage regulator arrangement is shown in **Fig. 16.3.** Generator hash may usually be suppressed by connecting a 0·05 μF capacitor from the " D " terminal on the generator itself to the frame of the generator, and connecting an 0·005 μF disc ceramic capacitor in parallel with it. If it is not possible to connect direct to the generator terminals, these capacitors may be connected from the " D " terminal of the voltage regulator unit to earth, but in this case it will be essential to fully screen the " D " lead between the generator and the voltage regulator unit.

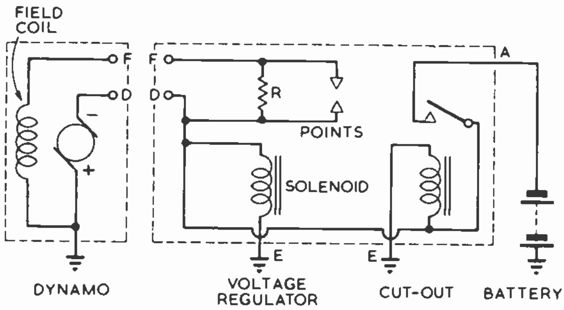


Fig. 16.3 The dynamo and battery charging circuit usually found in a modern car.

Under no circumstances must the " F " terminal on either the generator or the voltage regulator be bypassed in the same manner as the " D " terminal. If this is done, the action of the voltage regulator will be impaired and the unit damaged. If it appears advisable to bypass the " F " terminal in order to remove any interference caused by arcing at the voltage regulator points, a 5 ohm resistor and a 0·001 μF capacitor should be connected in series and wired between the " F " terminal and an adjacent earthing point. Larger capacitors should not be fitted under any circumstances, and neither should the 5 ohm resistor be omitted.

Prior to attempting to bypass the " F " terminal, it is better to determine the effect of a coaxial capacitor with a value of between 0·1 μF and 0·5 μF between the " A " terminal and earth. In many instances this will be found to give a worthwhile reduction in the noise level. For operation in the upper h.f. bands, and on the v.h.f. bands, only a co-axial type capacitor should be used, but on the lower frequencies, a standard capacitor may be satisfactory.

If the generator whine remains pronounced, a fully screened tuned wave trap may be inserted in series with the output from the generator, the trap being located as close to the generator " D " terminal as possible. The connection between the generator and the trap should be fully screened. It must be remembered that the coil of this wavetrap has to pass the full generator output current, and for this reason, the coil must be wound with not less than 12 s.w.g. wire. The large wire size makes such traps impracticable for frequencies below about 28 Mc/s. When such a trap is employed, the bypassing arrangements already described should still be fitted, and these must be positioned before the trap.

For 28 Mc/s a suitable wave trap consists of 12 turns of 12 s.w.g. wire wound as a self-supporting coil with an internal diameter of 1 in. tuned by a 3–30 pF concentric trimmer. For 70 Mc/s, five turns 12 s.w.g., $\frac{3}{4}$ in. diameter tuned by a 3–30 pF trimmer. For 144 Mc/s, three turns 12 s.w.g., $\frac{1}{2}$ in. diameter tuned by a 3–30 pF trimmer. In every case the wavetrap should either be set to the correct frequency by the use of a g.d.o., or by connecting it in series with the aerial lead of a suitable receiver and tuning the capacitor for maximum signal rejection, *before* the unit is fitted to the vehicle.

On most cars of British manufacture the regulator is unscreened and remote from the dynamo. In these cases a 3 μF capacitor should be connected to terminal B+ of the regulator and a 0·5 μF capacitor fitted to terminal D+ of the dynamo. The latter value should not be exceeded. In some cases a reduction in the noise level may be effected by connecting an earthing lead between the dynamo and regulator.

By this time, the general level of interference should have been reduced significantly, but some odd noises, perhaps only intermittent, may remain. Their possible sources are now considered.

Electric Clocks

Interference from this source is characterised by a distinct ticking sound. Suppression can usually be achieved by fitting a 0·1 μF capacitor directly across the connections on the rear of the clock.

Fuel Gauge

The rheostat operated by the float in the fuel tank may set up a spasmodic and wavering sound similar to that caused by drawing a piece of metal across a piece of corrugated metal. With the vehicle stationary and the ignition on but the engine not running, bouncing the rear of the vehicle up and down may persuade the noise to appear. This noise can usually be suppressed by connecting a 0·1 μF capacitor directly across the terminals of the petrol tank unit.

Electric Motors

Windscreen wiper and heater motors may be fitted with a 0·5 μF or a 3 μF capacitor which should be connected to the current input terminal. The capacitor should have a good connection to the vehicle earth and if sufficient suppression is not achieved then a feedthrough capacitor should be inserted in the input lead to the motor. This capacitor should be of 0·05 or 0·1 μF and a current rating of 5 amp. will generally be sufficient.

Brake Static

This is heard as a rushing sound whenever the brakes are applied, and is usually most pronounced in hot dry weather. On damp or wet days it may not occur. Special kits are available for fitting to the wheel mountings and brake drums, and only these should be used.

Tyre Static

This is usually spasmodic, usually being at its peak on hot dry days and when travelling on a well worn Tarmacadam surface. The noise is similar to that of brake static except that it is continuous, and nearly always punctuated by a sharp cracking noise. For tubeless tyres the most likely cure is to inject a little powdered graphite into each wheel, and to earth the brake drums as for brake static.

Slow Speed Noises

Occasionally, pulses of interference not heard during normal running may be noted under one or all of the following conditions, (*a*) when stationary with the engine idling (*b*) when running at slow speeds and (*c*) when the engine labours. Such interference invariably stems from discontinuity between various parts of the bodywork aggravated by the vibration which takes place under some of the above mentioned conditions. The biggest offender, and usually the least suspected, is the exhaust pipe, and if such noises are experienced, a braided copper strap should be connected between the exhaust pipe at the rear of the vehicle and the underchassis metalwork.

A further source of such noises is the copper strap which runs from the terminal of the battery which is connected to the car body, and here it is due to poor contact between the strap and the bodywork. The strap should be removed and thoroughly cleaned together with the bodywork connecting point. The end of the strap should be liberally treated with oil containing colloidal graphite, and then refixed.

Engines are normally gimbal mounted, and while earthing provisions are made, these can become defective. Miscellaneous noises may sometimes be removed by fitting a substantial flexible copper strap between the engine and the adjacent bodywork; similar treatment should be given to the bonnet cover.

Alternator Systems

There is now a trend to replace the usual d.c. charging system and its associated generator by an alternator/rectifier arrangement coupled with an electronic voltage regulator. Aside from the increase in reliability and its other advantages, such systems bring about a substantial reduction in the number of sources of interference to radio equipment.

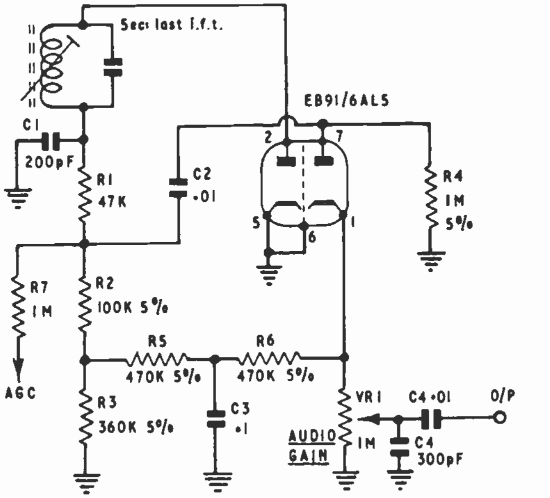


Fig. 16.4. Automatic noise limiter with good performance on ignition pulses. The circuit also functions as a diode detector and provides a.g.c.

NOISE LIMITERS

Although a great deal can be done to reduce noise interference on a vehicle to a very low level, some residual noise will remain unless extensive modifications are fitted such as fully screened units, cabling and screened plug connectors. For amateur purposes, the cost of such changes can rarely be justified. In addition, even assuming a perfect system, interference is almost bound to be picked up from nearby vehicles suppressed only to the statutory level. For the latter reason in particular, it is essential that a receiver intended for mobile communications work is fitted with an effective noise limiter.

Of the simple noise limiter circuits, the series limiter has proved to be somewhat more effective than the shunt limiter insofar as ignition noise is concerned. A self-following series noise limiter is shown in **Fig. 16.4** and in this the values have been selected for optimum operation on noise pulses of ignition duration. The circuit also functions as a detector and provides non-delayed a.g.c.

The shape of the noise pulses is modified by the preceding stages, with the risetime increasing as the bandwidth is decreased. Ideally a noise limiter should operate on the i.f. chain rather than on the audio output, and be positioned as near to the mixer as possible. Noise silencers functioning with the i.f. strip are a practical proposition, but in view of their relative complexity and the increase in the number of valves which they demand, they are not commonly used in mobile equipment.

An improvement in the performance of the circuit of Fig. 16.4 can be achieved by combining both series and shunt limiters in the manner shown in **Fig. 16.5**. In this case however, the double diode is only concerned with noise limiting, and a separate valve has to be provided for detection and a.g.c. Since this limiter circuit is effective on all types of noise pulses, and gives a higher degree of suppression, the use of an extra valve can be justified. The circuits of Fig. 16.4 and Fig. 16.5 are both self-adjusting to the carrier level.

A worthwhile improvement in the performance of a noise limiter can be obtained by following it by an audio filter designed to attenuate the upper frequencies. The very nature of a noise pulse gives it a high harmonic content, and thus if the frequency response of the following stages is restricted, it will have the effect of softening the residual pulses which manage to get through the limiter circuit. **Fig. 16.6** illustrates the circuit of Fig. 16.5 incorporated in an arrangement which includes such a filter, together with detector and a.g.c. circuits. The a.g.c. circuit is interesting in that it provides two levels of a.g.c., one for the i.f. stages and the other for the r.f. stage. In the interests of best signal-to-noise ratio, it is desirable to reduce the gain of the i.f. stages at a faster rate than that of the r.f. stage. This condition is achieved by connecting the r.f. a.g.c. line to a potentiometer across the a.g.c. source. The full a.g.c. voltage is fed to the i.f. stages, but only about one-fifth of the voltage developed is applied to the r.f. stage.

On v.h.f. and u.h.f. where adjacent channel interference is substantially less than that on the h.f. bands, it is practicable to combine a " squelch " circuit with a noise limiter. A squelch circuit is a gated audio circuit arranged in such a manner that the audio output is cut off until a carrier is present. The change in the circuit d.c. conditions which takes place when a carrier is received switches the audio output " on." One of the most satisfactory arrangements is that of the TNS (*Twin Noise Squelch*) the circuit of which is shown in **Fig. 16.7.** In this circuit, provision is made to

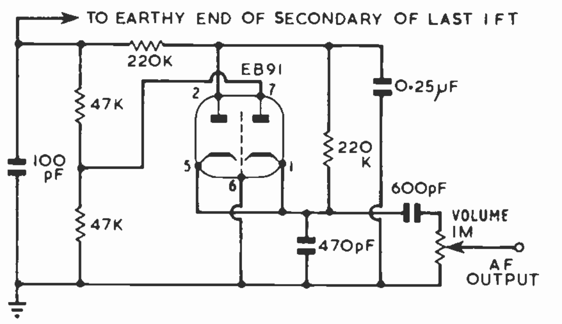


Fig. 16.5 A combined series and shunt limiter self-adjusting to the carrier level. A separate detector and a.g.c. circuit needs to be provided. See also Fig. 16.6 which incorporates this circuit.

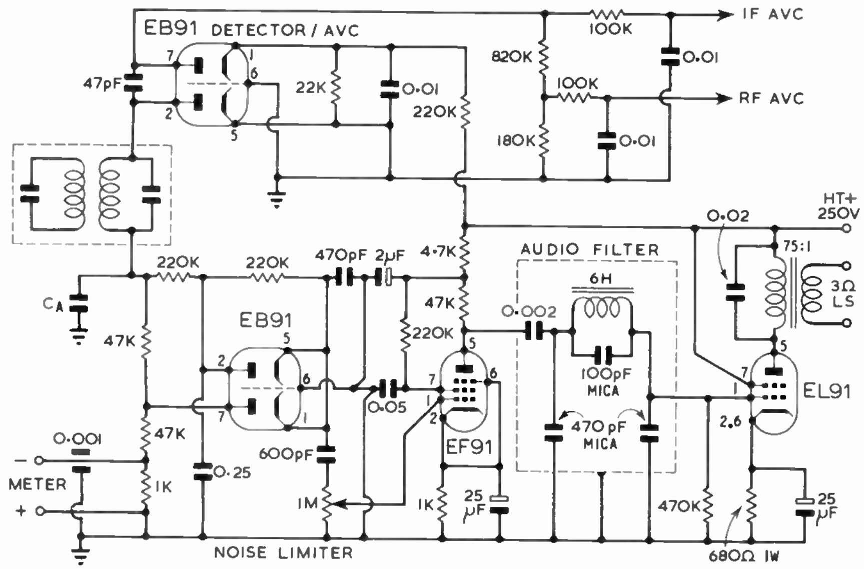


Fig. 16.6 Circuit diagram of noise limiter and a.f. stages for a mobile receiver. For intermediate frequencies between 450 kc/s and 2 Mc/s the capacitor CA should be 100 pF: above 10 Mc/s the capacity should be 47 pF. The two 470 pF capacitors in the audio filter should be ± 2 per cent tolerance type. Variations in these values will alter the cut-off frequency as will different inductance values for the choke. The meter test point is for use when aligning the receiver.

select the carrier level at which the audio output is turned on, this being determined by the position of R3. When R3 is greater than R2, the audio gate is open all the time, and the circuit functions purely as a limiter. As R3 is progressively decreased in value, an increasingly large carrier is required to trip the audio gate into the " on " position.

A setting of R3 will be found where the audio just goes " off." At this point it may be found that heavy interference pulses trip open the audio gate, and for this reason R3 usually has to be set slightly lower. This circuit is not suitable for the reception of signals which are below the inherent noise level of the receiver unless R3 is in the " off " position. However it is particularly useful when waiting for signals in the medium to strong category, say S3 and above, for during the waiting period the receiver will be completely silent. The effect of the audio gate is unusual for those more accustomed to the background noise produced by most v.h.f. receivers.

No matter which circuit is employed, every care must be taken in wiring a noise limiter to ensure that no stray coupling exists which could lead to the noise pulses bypassing the limiter. Input and output circuits should always be run in screened cable, these cables being connected to earth at one end only. Additionally, in the case of high performance limiters such as those shown in Figs. 16.5, 16.6 and 16.7 a low-loss valve base is essential.

Semiconductor Limiters

Generally, semiconductor diodes may be substituted for thermionic diodes in noise limiter circuits provided that certain requirements are met. Since any limiter circuit depends for its satisfactory operation on the diodes having an infinite resistance in one direction, and bearing in mind that the thermionic diode is the most perfect diode in this respect, it will be appreciated that unless the semiconductor diodes are chosen with care, the semiconductor version is likely to show a performance inferior to that of its valve counterpart.

Diodes employed in limiters should exhibit a reverse resistance of tens of megohms, and since germanium diodes cannot approach such values, they are generally unsuitable. Silicon diodes on the other hand can easily achieve reverse resistances of the order of 20 megohms or better, and these will give a performance comparable to a valve. Reverse resistance is not the only concern however. It is equally important that the forward resistance is not excessively high if the insertion loss is not to become unacceptable. Many of the cheaper silicon diodes, while having the attribute of a high reverse resistance, have forward resistances of between 100 ohms and 200 ohms, and this is far too high for noise limiter applications. The forward resistance should not exceed 50 ohms, and should preferably be substantially lower than this.

The circuits of Figs. 16.4, 16.5 and 16.6 may be modified for semiconductor operation without component changes provided that the foregoing points are borne in mind when selecting the silicon diodes.

In **Fig. 16.8** is shown the circuit of a semiconductor Dickert self-following limiter, in which the degree of clipping may be set to any desired level by means of the control VR1. If

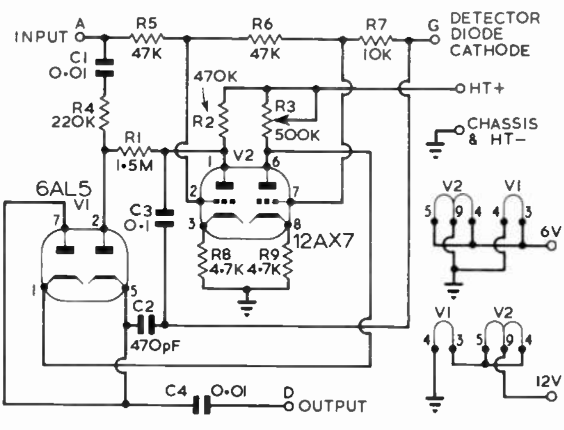


Fig. 16.7 Circuit diagram of the twin noise squelch.

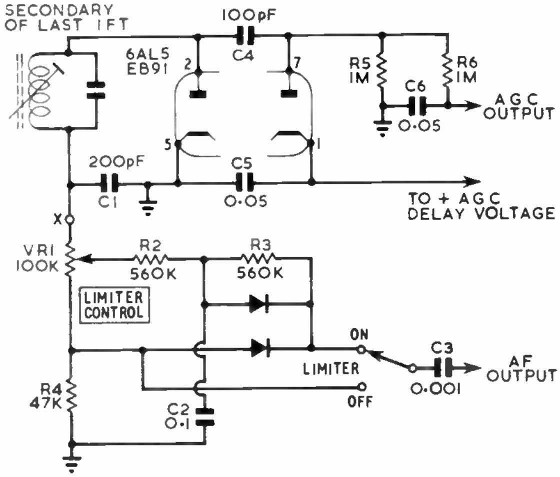


Fig. 16.8 Semiconductor Dickert adjustable noise limiter associated with valve operated detector and a.g.c. rectifier circuits. The a.g.c. delay voltage usually lies between 3V and 8V and may be obtained from the cathode of an audio valve or from a potentiometer chain across the h.t. supply.

the variable limiting feature is not required, VR1 may be replaced by a fixed resistor, and R2 connected directly to the point marked " X." Points requiring special attention are as follows: (*a*) Silicon diodes must be used; (*b*) the ON-OFF switch should be of a type having the lowest possible capacity between the stator terminals; and (*c*) The lead to the switch from the junction of VR1 and R4 must be screened as must be the lead to the switch from the cathodes of the diodes.

HEATER WIRING

In valve operated equipment, it is useful to be able to check the equipment under bench conditions without the need to provide a 12 volt heater supply. Aside from this requirement, if the units can be designed so that they will operate from either a 6·3 volt or a 12 volt supply for the valve heaters, then the equipment may be made to give service under fixed station as well as mobile conditions by employing a standard power pack.

Such a facility can be arranged by wiring the heaters in a balanced series parallel arrangement as shown in **Fig. 16.9**. In this, the requirement is that the current in the arm A/C is equal to the current in the arm C/B and that the valves connected in each arm are designed for 6 volt operation.

The first step is to note all the individual heater currents, and then arrange the valves in a manner similar to that of **Fig. 16.9** so that the discrepancy in the currents between

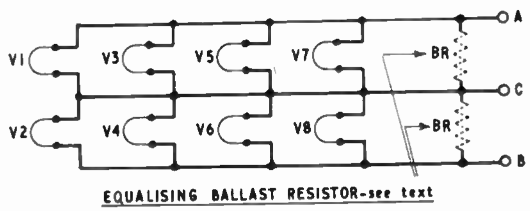


Fig. 16.9. Universal heater wiring allowing the optional use of a 6 volt or 12 volt supply.

A/C and C/B is as small as possible. For mobile operation dial lamps are something of a luxury and are better omitted in most cases. If dial lamps are used they are best wired separately. This is because they tend to be unreliable and on becoming faulty can upset the balance of the heater supply circuits.

The arm of the circuit which is *short of current* will have to be fitted with a ballast resistor of such a value that it consumes the current deficiency and so equalises the currents in the two arms. Since the current in the ballast resistor wastes power, the more similar the currents in each arm can be made by the careful selection of valve types during the design, the less waste there will be in the balancing process.

When 12 volts is applied across A/B then 6 volts will appear across A/C and C/B, and hence the value of the ballast resistor is calculated for an applied voltage of 6 volts according to R (ohms) $= E/I$ where E is 6 volts and I the current deficiency in amperes. The power dissipated in the ballast resistor will be W (watts) $= EI$. To allow adequate margin, the rated wattage of the resistor should be at least twice the calculated working wattage.

For operation on a 12 volt supply, terminals A and B are used. When operation on 6 volts is desired, terminals A and B are linked together, and the supply applied between C and AB. Normally B is the heater earth return through the chassis, and for 12 volt working the supply is applied to A and C is left floating, while on 6 volts, A is earthed, and the supply taken to C.

To avoid the possibility of connecting the heater line to a 12 volt supply when wired for 6 volt operation, it is suggested that the heater wiring is terminated on a socket placed on the rear of the chassis, and to which the various h.t. supplies may also be connected. Separate, and correctly wired plugs, can then be used, and these arranged to pick up and bridge across the ABC connections as required for each mode of operation. The use of a switch is to be avoided.

EQUIPMENT SCHEMES

When contemplating mobile operation, two possible courses are open:

(a) to employ a converter in conjunction with a BC receiver for reception, and use a separate transmitter, or

(b) to construct a transceiver unit to the specific requirements of mobile operation.

There is no doubt that the second course is by far the best. In a transceiver the modulator, or part of the modulator, can function as the audio amplifier during reception. In many respects the design, installation, and servicing of an integrated unit poses fewer problems than those encountered in connecting up and operating a number of individual units.

Added advantages of the self-contained transceiver, especially if designed for operation on either a 6 volt or a 12 volt heater supply, are its potential for use under fixed station conditions, and the ease with which it may be physically moved from one position to another. Even if the construction of a transceiver means duplicating equipment, this may have advantages. There is a degree of unreliability in all equipment, and if the main station receiver or transmitter should break down, then the mobile transceiver could take its place.

While the simplest transceivers are those designed for single band operation, many of the complications of band-

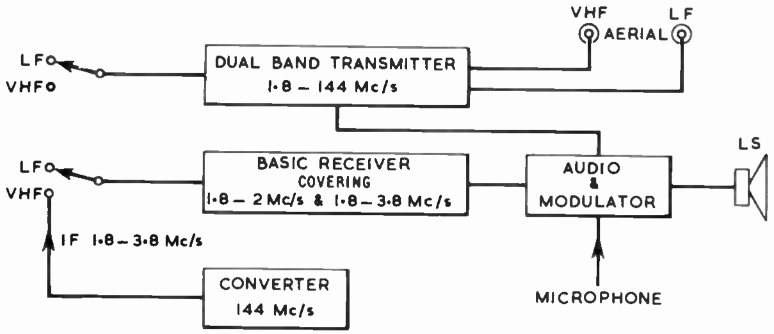


Fig. 16.10. Employing a switched converter with a basic receiver in conjunction with a dual band transmitter. This configuration may be adapted for 1·8 Mc/s and 70 Mc/s operation.

switching can be avoided if the transceiver is designed for operation on widely differing frequencies. The transmitter can be arranged along the lines of the 1·8 Mc/s/144 Mc/s dual frequency unit shown in **Fig. 16.34**, while the receiver can be a basic unit covering the lowest frequency range desired, in front of which is switched a self-contained converter for the higher frequency band. A block diagram of the method is shown in **Fig. 16.10.**

Despite the accent on the use of transceivers, the following circuits are mainly concerned with individual items. There is no reason however why various designs should not be linked together in transceiver form.

RECEIVING EQUIPMENT

With valve operated equipment there is always the problem of providing a suitable h.t. supply, and arranging for the conversion of the car supply up to the required h.t. potential at the highest efficiency. While transistorized receivers working directly from the vehicle battery offer the best solution in respect of reception, for a radio amateur faced by a limited budget and stocks of standard components, the cost of producing a completely transistorized receiver may be too great.

A partial solution is to be found in valves which are designed to operate from an h.t. potential of 12 volts, for by the use of such valves, standard valve practice may be employed in the r.f. mixer and i.f. stages, and these followed by semiconductor detector, a.g.c. and noise limiter circuits feeding transistorized audio stages. While the overall consumption will be substantially higher than that of a fully transistorized receiver, it will also be substantially lower than an arrangement which employs standard valves throughout together with a voltage conversion system. Such a receiver can be operated in conjunction with a transmitter in such a manner that only during the transmit periods does the voltage converter operate for the purpose of supplying h.t. to the transmitter stages. During reception, and by the use of the valves and technique outlined, consumption from the 12 volt source need not exceed 1 ampere for an audio output of 1 watt.

Low-Voltage Valve Converter

A circuit of a low voltage valve converter is shown in **Fig. 16.11.** This employs three readily available Mullard valves, types EF97, ECH83 and EF98, and depending on the coils fitted, will operate up to 30 Mc/s.

The EF97 functions as an r.f. amplifier, the ECH83 as a mixer, and the EF98 is triode connected and used as the local oscillator. It should be noted that although the ECH83 is a triode heptode, and as such could be used as a self-oscillating mixer, the triode section is unused. This course has been found to give higher oscillator stability on the upper h.f. bands. In a complete receiver, the triode section of the ECH83 may be used as the first amplifier of the audio chain.

It will be noted that the i.f. output is specified as 1·5 Mc/s, but this value is only quoted so that the converter may be used in conjunction with a car broadcast receiver functioning as an i.f. and audio unit. Any desired i.f. may be secured according to the frequency difference between received signal and that of the local oscillator.

Due to the lower mutual conductance of these valves, it is essential to employ ferrite cored coils of high Q. Small air cored coils are unsuitable, and their use will result in inferior performance.

One point requires special attention during construction.

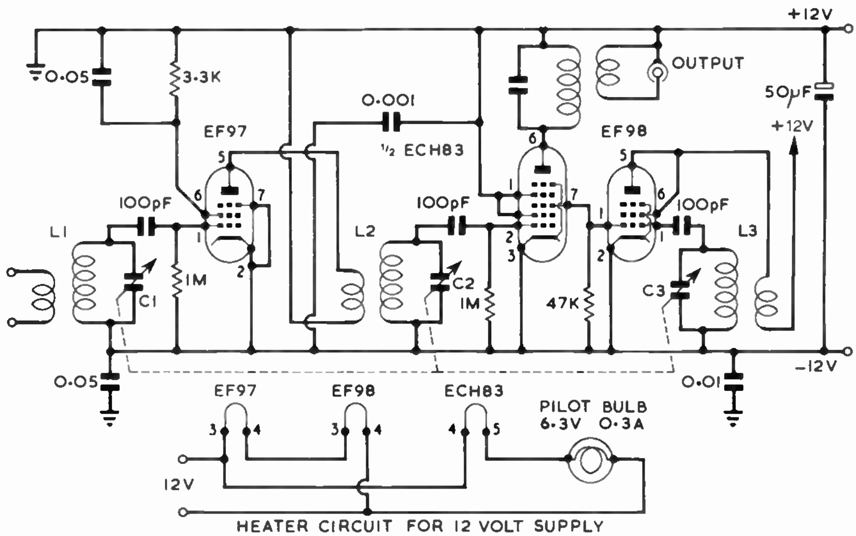


Fig. 16.11 Converter employing valves designed to work on an h.t. supply of 12 volts. Note that to accommodate a positive earth supply the usual earth line takes the form of a bus-bar, to which the frame of the tuning capacitor is also connected. The tuning capacitor must be fully insulated from the chassis. By the use of suitable coils, the converter may be operated on frequencies up to 30 Mc/s.

It will be seen that the "h.t." line is actually connected to earth since, in most vehicles, the positive of the battery is connected to the chassis. Unlike normal wiring therefore, the circuit ends which are usually connected to chassis must be connected to a bus bar which is suitably bypassed for both r.f. and a.f. currents. With regard to the tuning capacitor, while this could be connected directly to the chassis, the circuit then being completed through the earth bus bar decoupling capacitors, it is better to insulate the tuning capacitor from the chassis, and to connect the frame of the capacitor directly to the earth bus bar. If this course is adopted, an insulated coupler must be employed between the shaft of the capacitor and the dial system, or the capacitor shaft insulated in some other manner so that it cannot short circuit to the chassis.

Low Voltage Nuvistor Converter

RCA have introduced a nuvistor valve designed for operation with a 12 volt h.t. supply, in which a high standard of performance is maintained up into the u.h.f. region. This nuvistor has the type number 8056.

The circuit of **Fig. 16.12** illustrates the 8056 nuvistors in a 70 Mc/s converter designed for operation in conjunction with a positive earthed 12 volt supply. L1, L2 are mutually coupled to provide a level bandpass over the range 70·1–70·7 Mc/s. V1 functions as an inductively neutralized r.f. amplifier, L_N being the neutralizing inductance. The output of V1 is fed via a further bandpass circuit L3, L4 to the grid of the mixer V2. Injection from the local oscillator V3, which employs an overtone circuit, is by capacity coupling to the grid of the mixer valve. The purpose of the 15 pF capacitor between anode and cathode of the mixer valve is to form a bypass for the oscillator injection and signal frequencies, and this has the effect of reducing the mixer noise.

It will be noted that values are not given for the i.f. output frequency or the crystal, and neither are winding details of L5 specified. All of these are dependent on the equipment with which the converter is to operate. The converter may be operated with an i.f. down to 1·8 Mc/s, but below this frequency it is difficult to prevent radiation from the local oscillator swamping the signal circuits, and even at this frequency extensive screening is needed between the local oscillator and the aerial circuits L1, L2. At intermediate frequencies of 3·5 Mc/s and higher, the problem becomes somewhat easier, and no difficulties should be encountered.

In view of the proximity of the 70 Mc/s band to Band I TV and Band 2 F.M. transmissions, care should be taken to avoid intermediate frequencies which might give rise to interference.

The oscillator may be operated in either the third or fifth overtone mode, but whichever is selected, crystals designed for overtone operation should be employed. For injection frequencies higher than 60 Mc/s, HC-6/U subminiature surplus overtone crystals are satisfactory in the fifth overtone mode. These crystals are normally marked with their third overtone frequency, and thus their *fundamental* frequency may be determined by dividing the marked frequency by three.

Once the oscillator frequency has been decided, L5 should be proportioned so that it resonates at the injection frequency with the 12 pF capacitor. The coil may be calculated in the usual manner, or its size experimentally determined with the aid of a g.d.o. Initially, the tapping point should be set at one-fifth of the total number of turns counted from the grid end of the coil. Provided that the coil size has been determined with the core substantially out of the coil, winding the core into the coil will compensate for the tap, but in any event, the setting of the core over which oscillation takes place will be found to be broad. Provided that oscillation ceases when the crystal is removed, the circuit may be considered satisfactory. If oscillation continues with the crystal removed, adjust the position of the tap to reduce the number of turns on the grid section of the coil.

Nuvistor Hybrid Converter

The circuit of Fig. 16.12 may be modified to valve/transistor hybrid operation by substituting a transistorized overtone circuit for the oscillator. This arrangement has

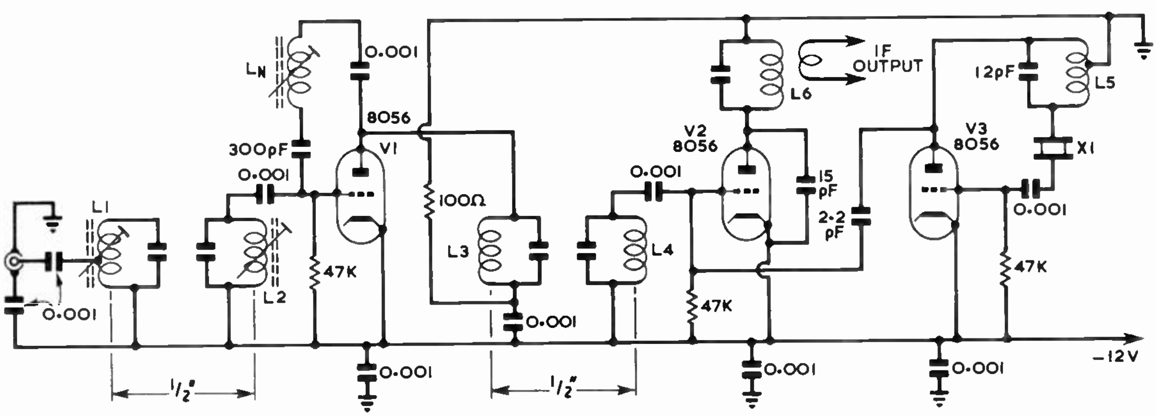


Fig. 16.12 70 Mc/s converter employing nuvistor valves designed for an h.t. supply of 12V. Coils L1, L2, L3 and L4, 5 turns, 22 s.w.g. 5/16 in. long on 1/4 in. dia. formers all fitted with ferrite cores. Aerial tapped on to L1 1½ turns up from earthy end. See text in relation to i.f., L5 and crystal. Ln, 32 s.w.g. enamelled wire close wound on 1/4 in. former for a length of 1/8 in. The design may be modified for other frequencies between 28 Mc/s to 146 Mc/s by appropriate coil changes.

certain advantages over an all-transistor converter and these are: (*a*) by using valves in the r.f. and mixer stages, the inferior high level cross modulation characteristic of bipolar transistors is avoided; (*b*) the reduced possibility of damage to the r.f. stage by stray r.f. when using a valve compared to a transistor. The heaters of V1 and V2 may be connected in series across the 12 volt supply.

The circuit of the transistor injection oscillator is shown in **Fig. 16.13**. Like its valve counterpart, this is also an overtone circuit, the crystal oscillating in its series mode

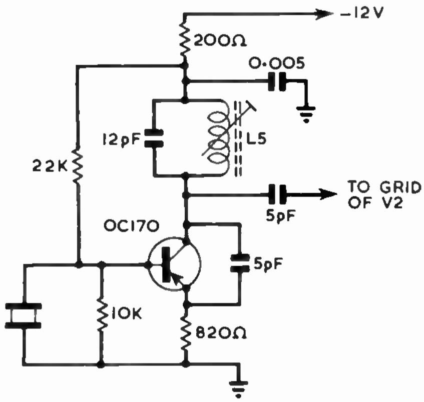


Fig. 16.13 A transistor overtone oscillator which may be substituted for the valve V3 in Fig. 16.12. HC-6/U crystals are suitable for this circuit. See text.

between the base of the transistor and earth. This circuit is not suitable for high capacity crystals, but will function with nearly all crystals in which the electrodes are sputtered or plated on to the crystal surfaces. Third and fifth overtone outputs are adequate, but the amplitude of output on the seventh overtone is insufficient to produce adequate injection into the mixer valve.

The frequency of the crystal, and the winding of the coil, L5, should be determined in the same manner as for the circuit of Fig. 16.12. Since the collector tuned circuit is fairly heavily loaded, L5 should be tuned to the correct frequency with the aid of a grid dip meter *before* the transistor is connected, and without the 5 pF capacitor between collector and emitter in position. Again due to the loading, the setting of the core of L5 over which oscillation takes place will be found to be not critical, and in fact variation of the setting of the core of L5 is a convenient manner in which to adjust the oscillator injection to the mixer.

Low Noise High Stability Converter

Under mobile conditions there is little point in employing equipment designed to give the lowest possible noise level because of the residual interference in a vehicle which has been suppressed, and noise generated by external sources. Under static mobile or portable conditions however, signal-to-noise ratio is as important as in fixed station operation especially in view of the inevitably inferior performance of a mobile aerial compared to that which can be used by a fixed station.

The circuit of **Fig. 16.14** shows a two valve three stage converter for the bands between 1·8 Mc/s and 30 Mc/s in which special attention has been paid to the signal-to-noise ratio, and also to the stability of the local oscillator. It employs two ECF82/6U8 valves in an unusual configuration.

The pentode section of the first ECF82, V1a, which has a low equivalent noise resistance, is employed as a straightforward r.f. amplifier. The mixer employs the triode sections of both valves, V1b and V2a, in a circuit originally developed for use on the v.h.f. bands and usually described as a cathode follower mixer. V1b operates as a cathode follower, the output of which is directly coupled to the cathode of V2a. The local oscillator is injected into the grid of V2a, and

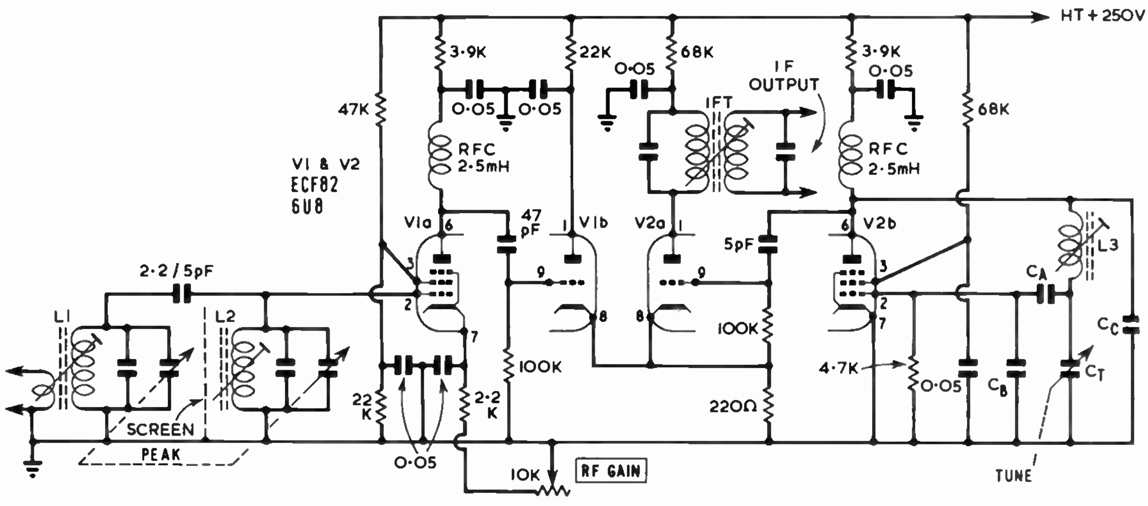


Fig. 16.14 Low noise high stability converter for frequencies between 1·8-30 Mc/s. V1 and V2 are both ECF82/6U8. Coils L1 and L2 are selected according to the tuning range to be covered, and for high Q. For the values of L3, CA, CB, CC, and CT see Table 16.1 and text.

mixing takes place in this section. The local oscillator is a high stability Vackar circuit using the pentode section V2b.

Due to the widely differing LC ratios in the oscillator and r.f. circuits, no attempt is made to gang these circuits, and thus the dial arrangement only operates on the local oscillator tuning. The r.f. tuning control functions as a peaking control in a manner similar to that employed in many s.s.b. transceivers. As the pentode section of the ECF82 has a short grid base, a.g.c. is not applied to the r.f. amplifier V1a. Experience has indicated that, under mobile conditions, unless the station being received is in close proximity, a.g.c. is not required, and where such a situation exists, operation of the r.f. gain control allows overload distortion to be removed.

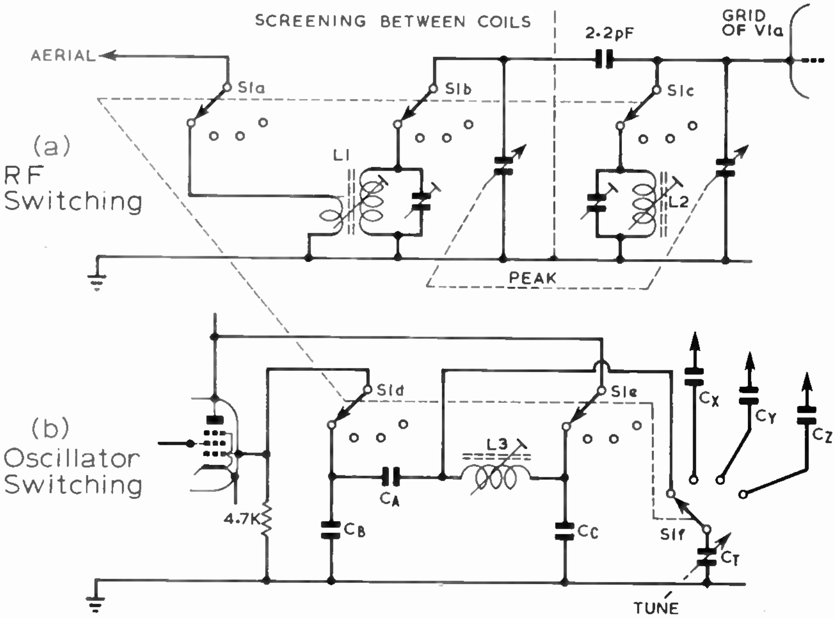


Fig. 16.15 Switching to be added to Fig. 16.14 for multiband operation. L3, CA, CB and CC, are selected from Table 16.1 according to the ranges required and the components for each ranged wired to the switch sections S1D and S1E as shown. For clarity only the components for one range are shown. CT is selected for the largest value required, and by use of the series capacitors X, Y and Z reduced to the effective capacity needed for the higher frequency ranges.

Coils used for L1 and L2 must be identical types and have a high Q. Details of the oscillator coil, together with values for CA, CB, CC, and an approximate value for CT will be found in **Table 16.1** for oscillator frequencies between 1 Mc/s and 39 Mc/s. In relation to the oscillator in general, attention is drawn to the sub-heading *Vackar Oscillator* in Chapter 6 on page 6.14.

While the diagram of Fig. 16.14 illustrates the unit as a single frequency tuner, there is no reason why the unit should not be bandswitched and so cover a number of ranges.

TABLE 16.1

Range in which oscillator is required to tune in Mc/s	L3 S.W.G. enam.	L3 TURNS close wound	CA	CB	CC	CT approx.
1·0— 1·5	34	Two layers each 70 turns	680	6800	680	10—300
1·5— 2·5	34	70	556	4700	556	10—250
2·3— 3·3	34	45	556	4700	556	10—250
3·2— 4·5	28	45	500	2700	500	10—100
4·5— 6·3	28	35	300	2700	300	10—50
6·3— 8·8	26	30	200	1800	200	10—25
7·8—11·0	26	20	200	1800	200	10—25
10·5—15·0	24	20	100	1000	100	10—35
13·5—19·5	22	15	100	1000	100	10—30
18·7—25·5	20	15	68	1000	68	10—25
25·0—33·0	20	10	68	1000	68	10—30
33·0—39·0	20	7	68	1000	68	10—30

All coils are wound on $\frac{5}{16}$ in. dia. formers and fitted with $\frac{1}{2}$ in. ferrite cores. Capacitor values for CA, CB, CC, and CT are in pF. The value quoted for CT is approximately that required to tune across the nearest amateur band. Larger frequency changes will require a corresponding increase in capacity swing, and it should be noted that tuning CT brings about a decrease in the oscillator frequency, and thus L3 should be selected for the highest frequency required. The 10 pF minimum value given for CT in each case includes stray circuit capacity. Oscillator frequency is signal frequency plus or minus intermediate frequency.

Such an arrangement is particularly useful where it is desired to employ it as a basic receiver on one or more of the h.f. bands, and to operate it with a v.h.f. converter having an i.f. output on a frequency which does not coincide with an amateur band.

The circuit of **Fig. 16.15** details the switching required, and the manner in which it is related to the circuit of Fig. 16.14. While the r.f. switching is straightforward, the oscillator switching requires some explanation in one particular respect. It will be noticed that on one frequency range the oscillator tuning capacitor is directly connected to the oscillator coil, while on the other ranges, it is connected via a series capacitor. Examination of Table 16.1 will show that as the frequency increases, so the value of CT decreases, and, in addition, to secure bandspreading over the full traverse of the tuning capacitor, different values are required for each band. The switching is therefore arranged so that the full value of the capacitor is connected to L3 on the lowest frequency band to be covered, but, by the use of series capacitors on the higher ranges, its capacity is effectively reduced to the value needed for the other inductances. These series capacitors are denoted as CX, CY and CZ on Fig. 16.15.

Crystal Controlled Converter

Where a broadcast receiver is already installed in the vehicle, it may be deemed satisfactory to employ a crystal controlled converter ahead of this receiver, and to use it as a tunable i.f. system. A converter suitable for the DX bands of 14 Mc/s, 21 Mc/s and 28 Mc/s is shown in **Fig. 16.16.** By suitably changing L1, L2, and modifying the oscillator

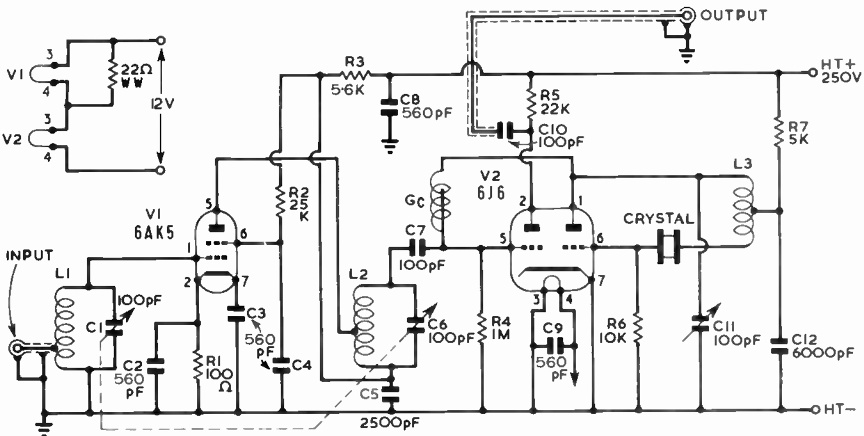
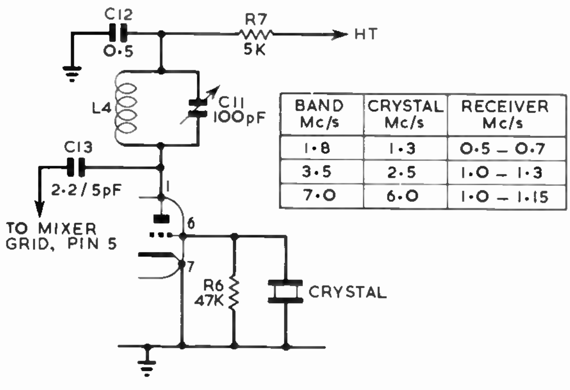


Fig. 16.16 Circuit diagram of the crystal controlled mobile converter. L1. 12 turns 16 s.w.g. ¾ in. dia., ⅞ in. long, tapped at 2½ turns; L2, 13 turns 16 s.w.g. ¾ in. dia. ¾ in. long tapped at 5½ turns; L3, 13 turns 16 s.w.g., ¾ in. dia., ⅞ in. long, tapped at 2½ turns. Crystal, see text and Table 16.2.

circuit, the same basic arrangement may be used for the lower frequency bands of 1·8 Mc/s, 3·5 Mc/s and 7 Mc/s.

Two valves are employed, a 6AK5 functioning as a r.f. amplifier, followed by a 6J6 mixer. One section of the 6J6 operates as a triode mixer, and the other as an overtone oscillator. In the original design of this circuit, injection from the oscillator to the mixer was achieved by positioning L2 and L3 relative to each other so that a degree of coupling existed between them. It has been found easier to set the oscillator injection to its optimum value by employing a small " gimmick " capacitor between the anode of the oscillator and the grid of the mixer, and to position the coils L2, L3 so that there is no interaction between them, i.e. at right angles to each other. This capacitor is shown as Gc on the diagram and consists of two 22 s.w.g. p.v.c. insulated wires twisted together for a length of ¾ in. On the lower bands, 1·8 Mc/s and 3·5 Mc/s in particular, Gc should be replaced by a capacitor of between 2·2 pF and 5 pF according to the notes which follow.



BAND Mc/s	CRYSTAL Mc/s	RECEIVER Mc/s
1·8	1·3	0·5 – 0·7
3·5	2·5	1·0 – 1·3
7·0	6·0	1·0 – 1·15

Fig. 16.17 Modified crystal oscillator circuit for use with Fig. 16.16 where the converter is to be used on the three lower H.F. bands. L4 is adjusted to resonate at the crystal frequency with C11.

Table 16.2 indicates some of the possible crystal frequencies, the oscillator output frequency, and the i.f. tuning range for the three higher h.f. bands. In Fig. 16.17 is shown the modified oscillator circuit for use when the converter is to be used on the 1·8 Mc/s, 3·5 Mc/s or 7 Mc/s bands with the corresponding crystal frequencies.

It will be noted that the anode of the mixer section of the 6J6 is not tuned by the primary of any i.f. transformer, the output being resistance capacity coupled to the following stages. This simplification aids the connection of the converter to following equipment, and gives a more level output over the band tuned on the broadcast receiver.

Provided that L1 and L2 are carefully constructed, and well screened from each other, it has been found possible to use a ganged tuning capacitor for C1/C6, and this aids operation by reducing the controls to a single " peaking " knob, the major use of which will be in the 28–30 Mc/s band. Where the receiver following the converter has appreciable gain, the output of the converter may be reduced by lowering the value of the output capacitor C10.

TABLE 16.2

Suggested crystal frequencies and outputs for the crystal controlled converter.

Band	Crystal frequency	Oscillator output	Receiver tuning
14 Mc/s	4333 kc/s	13 Mc/s	1—1·5 Mc/s
21 Mc/s	4000 kc/s	20 Mc/s	1—1·5 Mc/s
28—29 Mc/s	9166 kc/s	27·5 Mc/s	0·5—1·5 Mc/s
29—30 Mc/s	5700 kc/s	28·5 Mc/s	0·5—1·5 Mc/s

Layout of the converter follows standard practice, and provided that the grid coil L1, is well screened from the anode coil L2, no difficulties should be encountered. This is best achieved by placing a screen across the full width of the chassis so that it crosses the base of V1 separating the anode and grid pins. Bypass capacitors should be connected to the centre spigot, which is earthed by a wide strap, to a tag under one of the valve base retaining nuts.

After the wiring has been checked, the converter should be linked to its associated receiver, the crystal inserted, and power applied. The frequency of the crystal oscillator should either be verified with a sensitive absorption wavemeter, or checked by listening on the appropriate frequency on a wide range receiver. It is essential to verify the frequency where the converter is to be used on the three higher bands since in some instances the oscillator operates in the third harmonic mode, and in others on the fifth harmonic. After it has been ascertained that the crystal oscillator is oper ting in a satisfactory manner, a signal, preferably from a generator, should be applied to the input socket. Where ganged tuning has been incorporated, C1/C6 is adjusted to peak the signal, and then the turns of L1 are squeezed or spread to bring about any further increase. Following this, the capacity Gc should be adjusted, first larger, and then smaller, to determine whether the injection level is correct. Over a particular range of the size of Gc, the output will remain substantially constant. The size of Gc finally fitted should be the smallest capacitance at which the output remains at this level. With the modified circuit the value of C13 is changed to bring about the same effect.

On the lower frequencies, commercial coils and their associated link windings may be substituted for L1 and L2.

Low Consumption Communications Receiver for 1·8 Mc/s, 3·5 Mc/s and 28 Mc/s

This design will mainly be of interest to the more experienced operators since it provides circuit information only. It is primarily intended for those who do not wish to venture into the field of transistors, but who require a sensitive receiver for mobile or portable operation. In view of the rapid advances in transistors, it could well be that receivers of this type are considered archaic, but the design fills an information vacuum.

The receiver is designed to operate with either a 12/14 volt heater line supply unit, or from a mains power unit providing a 6·3 volt heater supply. This is accomplished by series-parallel connecting the valve heaters in the manner previously described. At 12 volt nominal, the heater circuit consumes 1·1 amp., while on 6·3 volts this rises to 2·2 amps. These currents are without optional dial lamps fitted. The 250 volts h.t. supply line requires a current of just under 90 mA under no-signal conditions, and thus to accommodate the stabilizer for the 105 volt h.t. line, the power unit has to be capable of delivering a maximum of 120 mA at 250 volts. In mobile operation where the h.t. supply may be obtained from a transistorized d.c. to d.c. converter, the total battery consumption does not exceed 4·5 amps. With a vibrator supply, and its lower efficiency, a higher overall consumption must be expected.

The circuit diagram is shown in **Fig. 16.18.** V1, a 6BJ6 functions as a conventional r.f. amplifier, with a.g.c. Its cathode is returned to the manual r.f. gain control line. V2 is a 12AH8 and operates as a triode-hexode mixer. A.G.C. is not applied to this section in the interests of freedom from oscillator pulling when strong signals are received; as a further aid to oscillator stability, the oscillator h.t. is taken from a stabilized h.t. line of 105 volts.

A half lattice crystal filter is positioned between the output of the mixer and the grid of the first i.f. stage. The crystals are surplus FT241 types having adjacent channel numbers of X1 = Channel 335 (465·27 kc/s) and X2 = Channel 334 (463·88 kc/s). The separation of these crystals amounts to 1·89 kc/s, and if this is found to give unacceptably sharp tuning, or to make the audio quality poor, then change X2 *only* to Channel 337 (468·05 kc/s). This will give a separation of 2·78 kc/s, and provided that the transformers specified for T1 and T2 are employed, will be entirely satisfactory for a.m. phone transmissions assuming that the receiver is correctly tuned.

With such a simple filter, undesirable side responses may be encountered, as may a dip in the centre of the bandpass. It has been found that both of these effects may be minimised to a point where they are no longer troublesome by careful selection of the crystals used at X1 and X2, and care in the alignment of T1 and T2. Starting with six crystals of each of the channel numbers to be employed, it has been found possible to select a pair which give a satisfactory performance without becoming involved in shifting any of the crystal frequencies. With six sets, two pairs will probably be found, but despite this, it is prudent to commence with six of each type.

The overall selectivity of the filter is adjusted by varying the impedance of the crystal loading by means of VR3. This carbon track control should be adjustable from the front panel. If required, VR3 could be replaced by a switch giving four pre-selected sensitivity levels, as illustrated in **Fig. 16.19.**

The first i.f. amplifier, V3, a 6BJ6, has a.g.c. control, manual adjustment of gain is obtained by returning its cathode lead to the r.f. gain control circuit. The second i.f. amplifier V4, again uses a 6BJ6, but in this case, neither a.g.c. nor manual gain control is applied.

A meter is connected to measure the difference in potential between the cathodes of V3 and V4. Since the voltage measured will vary according to the signal level, the meter can be calibrated in terms of signal strength.

Under no-signal conditions, the cathodes of V3 and V4 will both be at a potential of approximately 1 volt and there will be no reading on the meter. When receiving very strong signals, assuming that sufficient a.g.c. will be developed to cut off V3, causing its cathode to fall to zero volts, the potential across the meter will rise to 1 volt. Potentiometer VR4 is included in series with the meter to convert it into a voltmeter with a full scale deflection of 1 volt. In practice, however, V3 will never be cut off completely, and for this reason even the strongest signal will not drive the meter pointer against the full scale end stop. The cathode fixed resistor of V4 is a potentiometer, the slider of which is taken to the metering circuit. With the aerial and earth terminals shorted together, adjustment of VR5 enables the meter to be set to zero by allowing point R to be set to the same potential as that developed at point Q.

Semiconductors are employed for detection and a.g.c. A delay bias of approximately 20 volts is applied to the a.g.c. diode CR2 by returning its cathode to the junction of the potentiometer chain R21/R22. A.G.C. delay is to some extent a matter of personal choice, and if this is considered to be too high, it may be reduced to approximately 10 volts by replacing R22 by a 10 K ohm resistor. As the diodes specified do not have an infinite reverse resistance, under low, or no-signal conditions, the a.g.c. line will swing slightly positive. This is overcome by fitting the clamper diode CR3 directly across the a.g.c. line connected in such a manner that it conducts in the presence of positive voltage.

V5 is a self-adjusting noise limiter circuit employing a 6AL5/EB91 double diode. While silicon diodes could have been used, consideration of the heater wiring circuit will show that there was no point in doing so unless some other valve could also be removed, for, with V5 in circuit, each arm of the series parallel heaters balances to within 50 mA, whereas without V5 a larger unbalance would result.

The double triode V6, 12AU7/ECC82, combines the functions of b.f.o. and the audio pre-amplifier. The output stage is a conventional one using a 6AM5/EL91.

The receiver covers the three most popular mobile bands of 1·8 Mc/s, 3·5 Mc/s and 28–29·7 Mc/s with the constants of the tuned circuits proportioned so that each band is spread over virtually all of the swing of the tuning capacitor.

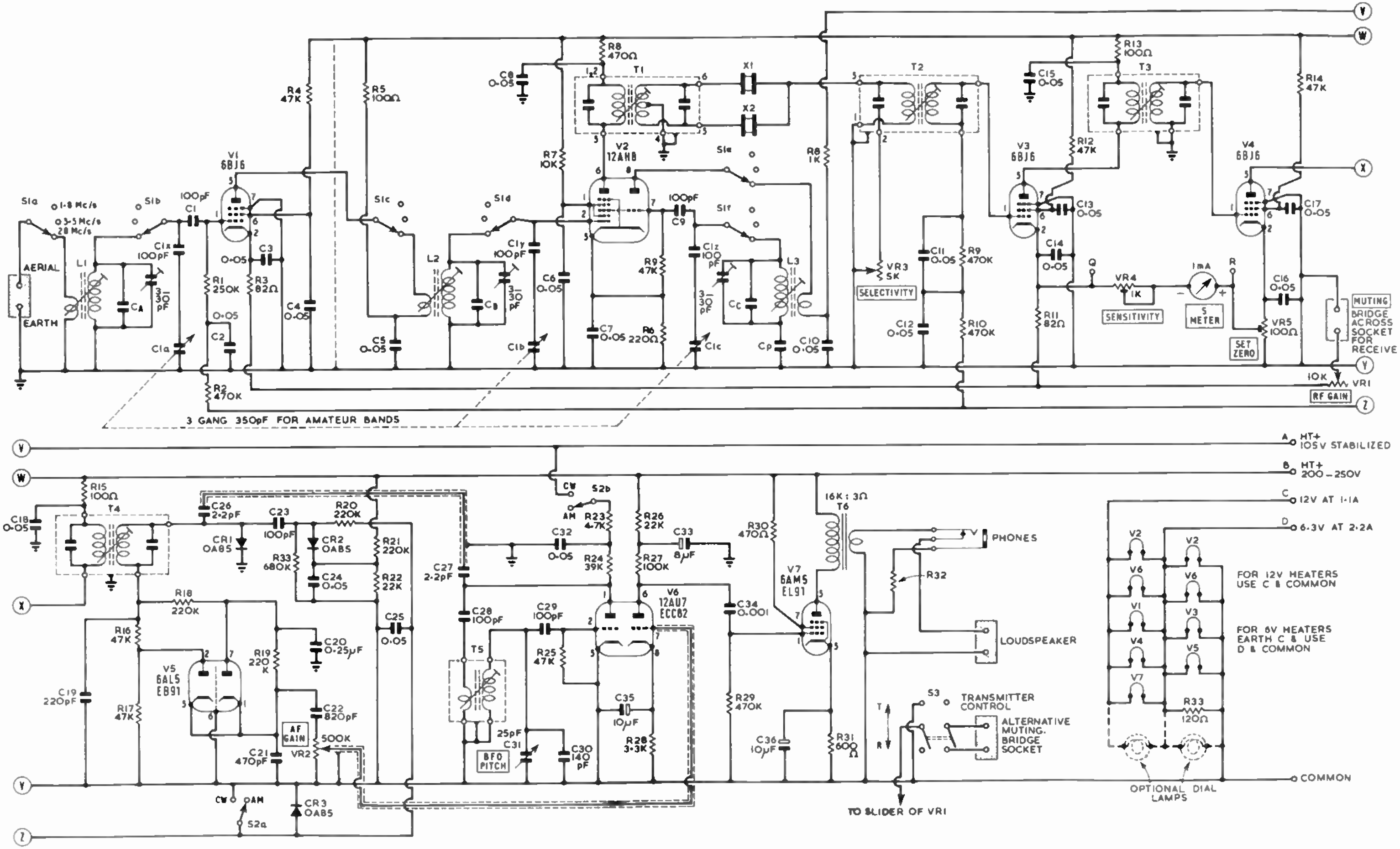


Fig. 16.18 Low consumption communications receiver for the 1·8 Mc/s, 3·5 Mc/s and 28-29·7 Mc/s bands. Each band covers the full sweep of the tuning capacitor. The receiver features a half lattice variable selectivity crystal filter, self adjusting noise limiter, and an S meter circuit. Other ranges may be added by suitable changes in coils and padding capacitors, but the tuning will be restricted to that given by 50 pF which is the effective capacity of each section of the ganged tuning capacitor.

TABLE 16.3

Coils, capacitors and i.f. transformers used in Fig. 16.18

Band	R.F. Coils L1	Mixer Coils L2	Osc. Coils L3	Value of CA, CB, CC	Cp
1·8 Mc/s	Blue 3	Yellow 3	Red 3	340pF	1100pF
3·5 Mc/s	Blue 3	Yellow 3	Red 3	170pF	1100pF
28 Mc/s	Blue 5	Yellow 5	Red 5	50pF	Nil

All coils are Denco miniature dual purpose. The number following the colour denotes the manufacturer's range number.

Transformers

	T1 and T2	T3 and T4	T5
Denco	IFT.11/465/CT	IFT.11/465	BF02/465

Range switch

Low loss ceramic, two-pole three-way per bank, three spaced banks.

I.F. Frequency

Normally 465 kc/s. Correct frequency centred on crystal filter.

To ensure accurate tracking, especially on the higher frequency range, and to ensure that the coverage is duplicated, it is essential to use the coils specified in Table **16.3**, and to use 1 per cent tolerance capacitors in positions CA, CB and CC, and also in positions C1X, C1Y, and C1Z. The tuned circuits are designed for a total stray capacity of 40 pF, to ensure this value will be attained, 3–30 pF trimmers are fitted across each coil. While this may seem to be high, it must be remembered that the residual capacity of the tuning gang is also included in this figure. The coils should be switched by a low loss switch, preferably ceramic insulated, and the tuning capacitor must also be a high quality unit.

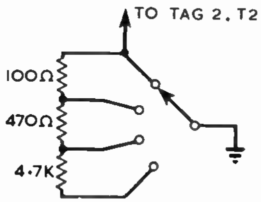


Fig. 16.19 Four stepped switched selectivity controls which may replace VR3 in Fig. 16.18.

Alignment follows standard practice for this type of receiver. That is to say, the i.f. amplifier is aligned first, starting from the detector and working back to the mixer. When the i.f. is injected into the grid of the mixer, the selectivity control should be set to its sharpest position and the ganging oscillator tuning swung to find the sharp output peak which will occur as its frequency coincides with that of the lattice filter. Carefully tune about this point until the oscillator is precisely in the middle of the lattice bandpass, and then readjust the i.f. transformers. For further information on the alignment of crystal lattice filters, see Chapter 10 page 10.12. Finally, the oscillator, mixer and aerial circuits are adjusted, bearing in mind that the cores are employed at the l.f. end of the range, and the trimmers at the h.f. end. In view of the limited frequency coverage, there will be no difficulty in achieving a high order of tracking between the relative circuits. Dial calibration is best undertaken by the use of a 100 kc/s sub-standard bar once the signal generator has verified the approximate frequency coverage. Take particular care that the oscillator is set on the correct side of the signal frequency.

1·8 Mc/s and 3·5 Mc/s Transistorized Converter

This converter is designed to function in conjunction with a standard car radio used as a fixed i.f. and audio chain, the output from the converter being at 1500 kc/s (200 metres). The converter is tuned, and this permits direct calibration eliminating the need for mental conversion from the car radio dial readings.

The converter, shown in **Fig. 16.20** makes use of miniature plug-in coils which fit B9A valve bases, and while changing bands does involve changing coil sets, the simplification which this allows in the wiring is adequate compensation. If required, the converter can be redesigned for switched coil operation.

Transistor TR1 functions as a normal earthed emitter amplifier, in which the base and collector circuits are tuned by L1 and L2 respectively. To avoid the need for neutralization of the r.f. amplifier, a high frequency type of transistor must be used for TR1.

TR2, in conjunction with L3, operates as a self-oscillating mixer in which the oscillator frequency is higher than the signal frequency by the i.f. of 1·5 Mc/s. The intermediate frequency is extracted by the transformer T1. It is important that the lead between the output socket and the car radio aerial socket is fully screened.

Self-oscillating transistor mixers can be highly unstable as far as communications applications are concerned, and in addition, can exhibit frequency pulling in the presence of strong signals. The stability of the oscillator depends very much on the design of L3, and if the coil details are closely followed, operation will be satisfactory. Pulling effects are minimized by stabilizing the supply to both the r.f. and mixer stages. It is interesting to note that stabilization of the supply to the r.f. stage gives a substantial improvement over results obtained when only the mixer supply is stabilized.

In view of the low consumption of the converter, a 250 mW zener diode, ZD1, is adequate, and as it is operated in the region of 60 mW, there is no need to mount it on a heat sink. By adjusting the value of R9, the converter may be operated on a 9 volt supply, or on a supply varying between 12 volts and 14·5 volts as is likely to be encountered when the car battery is used as the source.

In construction, care should be taken to screen L1 and L2 from each other. This is best accomplished by employing the canisters in which the coils are supplied, as screens fitted over the coils. Information on this procedure is supplied with the coils. It will be noted that a three gang 50 pF tuning capacitor is specified. As an alternative, a two gang 50 pF capacitor may be used for C7a and C7b and a separate capacitor of 50 pF for C7c, the latter being coupled to the tuning dial. The two gang unit will then function as a *peaking* control for the r.f. circuits. Where the unit is to be used under mobile conditions, it should be constructed within a fully screened enclosure provided with a tight fitting lid. For this an Eddystone die-cast box is ideal. In addition, the filter circuit shown in **Fig. 16.21,** should be included in series with the power supply lead.

Transistorized I.F. Strip

Working on the premise that, as far as the h.f. bands are concerned, it is not particularly easy to design a multiband aerial which shows any degree of efficiency under mobile conditions, it can be argued that while a bandswitched receiver might be desirable, it is not essential. If this line of reasoning is acceptable, then the r.f. section shown in

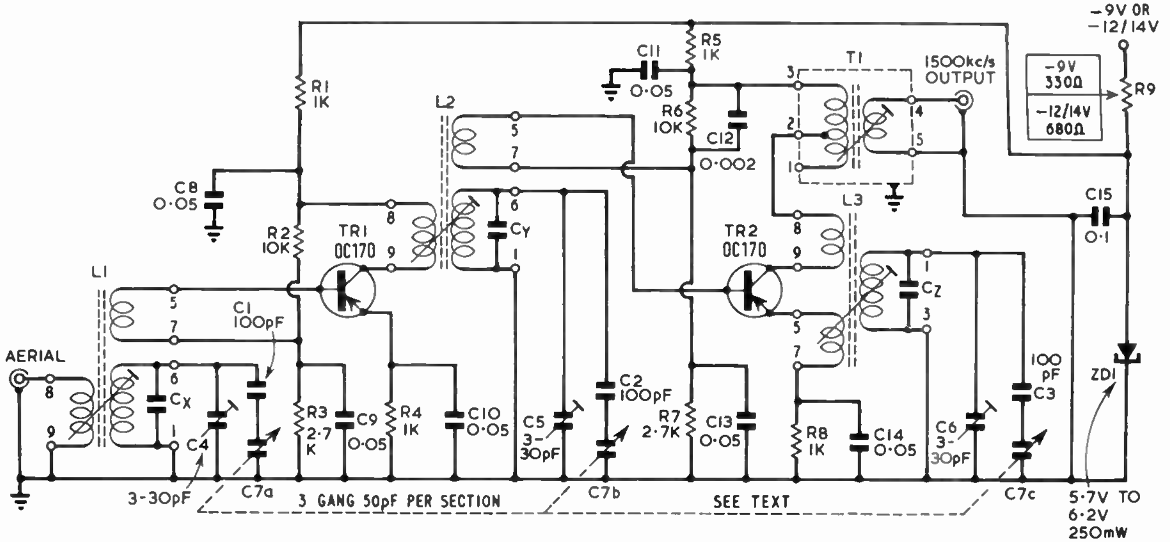


Fig. 16.20 1·8/3·5 Mc/s transistorized converter for use with a standard car radio as a fixed i.f. chain. The car radio should be tuned to 1500 kc/s (200 metres). The converter employs miniature plug-in coils fitting B9A valve bases. Capacitors CX, CY and CZ are fitted directly to the coils so that their values are automatically changed with the coils. The numbers on the coil connections correspond to the pin connections on the B9A bases. See text and Table 16.4. T1 is a Denco IFT.16.

Fig. 16.20 can be employed as the initial stages of a communications receiver, bandchanging being effected by changing coil sets.

Acceptance of this approach not only simplifies construction but allows a considerable reduction in the overall size of the unit since, in any multiband receiver, a large proportion of the space is taken up by the coil banks and their associated switching.

The circuit diagram of an i.f. amplifier, detector, a.g.c. circuit and noise limiter unit for use with the tuning head of Fig. 16.20 is shown in Fig. 16.22. To provide adequate selectivity at the intermediate frequency of 1·6 Mc/s, a simple crystal filter is included together with a phasing control and a selectivity control, the latter altering either the width of the nose or the depth of the rejection slot. Provision is made, although it is not obvious from the circuit diagram, to remove the crystal filter from the circuit. The a.g.c. circuit is highly effective under conditions of flutter which are often encountered during mobile working. Interference suppression is by the G3GFN Noise Limiter which has proved to be considerably superior to diode limiters under mobile conditions. In respect of this circuit it should be mentioned that it is protected by a number of patents held by G3GFN, and while it may be used in amateur constructed equipment, it may not be incorporated in commercial equipment without prior arrangements.

Following the crystal filter unit, transistors TR1 and TR2, which are high frequency alloy drift types showing very high gain at the intermediate frequency, operate as a.g.c. controlled amplifiers, control of their gain being effected by variation of the base-emitter current. TR3 is a further i.f. amplifier operated under fixed gain conditions. The i.f. output transformer, T5, feeds a full-wave detector circuit, and to secure as much negative voltage as possible for the a.g.c. control circuit, T5 is connected for high impedance feed to the detector circuit. The noise limiter circuit consists of TR4 and TR5. VR4 functions as the limiter control, while S1 allows the limiter circuit to be bridged across when it is not required. It is essential that S1 is a low loss switch in order to prevent noise pulses bypassing the limiter.

The a.g.c. varies the potential, and hence the forward bias current, between the emitter and base of the controlled transistors TR1 and TR2, in proportion to the signal. The less the signal, the higher the base potential, the forward bias, and hence the gain. The disadvantage that this system needs a positive supply source is more than offset by its functional superiority and the fact that a manual gain control may be incorporated. Only one precaution has to be observed, and this is that the positive source and the main negative supply must be switched on and off together. It is

TABLE 16.4

Coil data for Fig. 16.20.

Band	R.F. coil, L1	Coupling Coil L2	Osc. Coil L3	CX, CY & CZ	CP
1·8 Mc/s	Blue 3T	Yellow 3T	White 3T	330pF	330pF
3·5 Mc/s	Blue 3T	Yellow 3T	White 3T	180pF	330pF

As the same coils are used for each range, the only difference being in CX, CY and CZ, each set should be provided with some additional identification. The coils are produced by Messrs. Denco of Clacton.

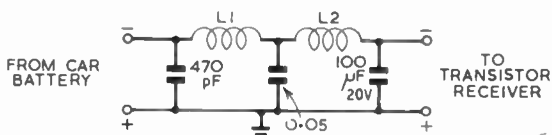


Fig. 16.21 Filter for use in series with supply from car battery to transistor receiving equipment. L1/L2 are a single winding of 50 turns 22 s.w.g. enam. wound on a 2 in. length of ⅜ in. dia. ferrite rod tapped at 10 turns. The section 0 to 10 turns is L1, and the balance L2. 100μF capacitor is an electrolytic type.

not sufficient to include a single switch in the common return line of the supply.

It will be seen that the bases of the controlled transistors are returned to the earth line and thus to achieve forward bias, the emitters have to be made positive with respect to earth. With the full bias voltage between base and ground, gain will be at maximum. The bias voltage is obtained via the series transistor TR7, the emitter of which is connected to the positive supply, and the collector provides the output terminal. Since the base of this control transistor is returned to earth through R19, and earth is negative with respect to emitter, TR7 is fully forward biased and the collector voltage is nearly +4·5 volts. However, R19 is not only the base resistor of TR7, but also the collector load of TR6. Without any signal present, TR6 has no forward bias and is therefore in the cut-off condition. Once the detector circuit produces a negative output voltage due to a signal, TR6 receives forward bias and commences to conduct. The resulting current through R19 increases the voltage drop across it, and this in turn reduces the forward bias to TR7. As the forward bias to TR7 decreases, its collector current falls, the bias to the controlled stages drops and their gain is reduced. TR7 may, for all practical purposes, be considered as an electronic variable resistor the value of which increases with increasing signal strength.

This i.f. amplifier has a very high level of overall gain, and for complete stability it is essential that the interstage screens are fitted in the manner indicated, and that no attempt is made to omit any of the decoupling in the collector and emitter circuits. The screens are arranged so that they divide the input and output connections of the transformers T2, T3 and T4, and so that TR1 is contained within one screened compartment, and TR2 within another. Providing that these requirements are met, and the circuit laid out in a straight line, no difficulties will be encountered.

Transformers T1 and T2 have to be slightly modified. In the case of T1, the tap on the secondary winding is removed from connection five and taped back to the secondary to secure its position. Following this, the fixed capacitor across the secondary winding is removed, and replaced by two capacitors of twice its value connected in series. The junction of these capacitors is connected to tag 5 on the transformer. If difficulty is experienced in obtaining capacitors which will fit inside the screening can, the two capacitors may be mounted on the base connections of the transformer after it is fitted to the chassis.

In the case of T2, the primary tap is removed from tag 2 and taped back to the primary to secure its position. The primary tuning capacitor is then removed from tag 3 and connected to tag 2.

The phasing capacitor also functions as a

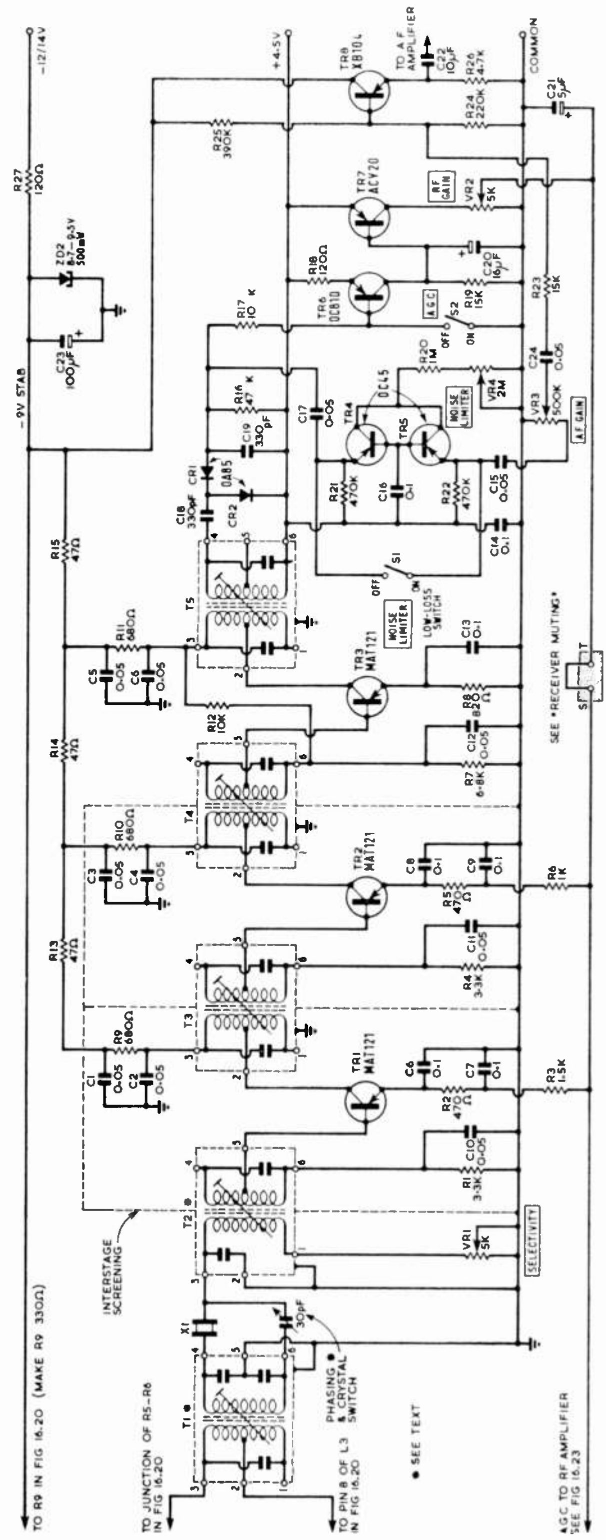


Fig. 16.22 I.F. amplifier, detector, a.g.c. and noise limiter circuits for use with Fig. 16.20 or any similar converter system with an output of 1·6 Mc/s. While the frequency of this unit is normally 1·6 Mc/s, the actual alignment frequency of the i.f. amplifier depends on the frequency of the crystal used at X1. With the transformers specified, the crystal frequency may lay between 1550 kc/s and 1650 kc/s, and units in this range are available "surplus." Transformers T1 to T5 inclusive are Denco IFT.18/1·6. T1 and T2 are modified according to the text. A BFO may be added by incorporating the circuit of Fig. 16.24. The zener diode ZD2 should have a free air dissipation of not less than 350 mW.

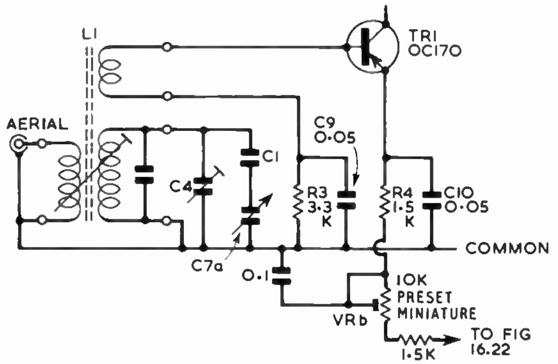


Fig. 16.23 Modification to r.f. amplifier of Fig. 16.20 to apply a.g.c. from the circuit of Fig. 16.22. With the lowest possible signal VRb should be set to the position which gives maximum signal output. It is important that the input signal is below that which results in a.g.c. action.

switch to connect and disconnect the crystal filter as required. This is achieved by the simple expedient of bending the tip of one of the moving plates so that it makes contact with the fixed plates in the full capacity position. It should be noted that the phasing capacitor requires a fully insulated mounting, and an insulated coupler to the panel control.

The crystal may be of any frequency between 1550 kc/s and 1650 kc/s. In alignment, the i.f. amplifier is aligned to the crystal frequency, and then the signal and local oscillator circuits of the converter adjusted to produce the intermediate frequency.

If the r.f. amplifier of Fig. 16.20 is modified in accordance with **Fig. 16.23,** a.g.c. may be applied from the circuit of Fig. 16.22. Additionally, a b.f.o. may be added to the i.f. amplifier by incorporating the circuit shown in **Fig. 16.24.** An " S " meter circuit may also be added to Fig. 16.22 by connecting a meter with a full scale deflection of 4·5 volts between the collector of TR7 and the common earth line, *positive* to collector. Under no-signal conditions, the meter will read the full value of the forward bias, that is 4·5 volts, but as the output from the detector circuit operates on TR6 and TR7 and so reduces the level of the forward bias, so the meter reading will fall. For signal strength measurements, the " S " scale is provided with a zero on the right hand side. Although such a meter might be considered a luxury for mobile equipment, in this case it will also indicate the state of the 4·5 volts positive supply. To reduce the loading on the a.g.c. line, the meter movement should have a sensitivity of 1 mA or better, and if part of the series resistor is made variable, this can function as a zero set control for signal strength readings.

The main power for the circuit of Fig. 16.22 is derived directly from a source of between 12 volts and 14·5 volts, and this is regulated by the zener diode ZD2 to the 9 volts required by the unit. Where the circuit of Fig. 16.20 is employed in conjunction with that of Fig. 16.22, the supply is derived from Fig. 16.22. In addition to the negative voltage of between 12 volts and 14·5 volts, the unit requires a positive supply of 4·5 volts. This is best provided by a dry battery built into the complete equipment. In view of the low current demand on this battery its life will approximate to the shelf life. A leak-proof battery should be fitted. As has been previously noted, the on-off switching must be arranged so that both the positive and the negative supplies are disconnected in the " OFF " position.

Where the units of Fig. 16.20 and Fig. 16.22 are to be used in conjunction with a transmitter, the receiver should be muted in the transmit condition by breaking the connections between the points marked as S and T on the a.g.c. control line in Fig. 16.22. To ensure protection of the r.f. amplifier, the circuit of Fig. 16.20 must be modified in accordance with Fig. 16.23. Any current flowing in the base-emitter circuit of the r.f. amplifier will be limited to a safe level unless there is severe leakage across the aerial relay during transmission, and further, it is unlikely that the reverse voltage rating of the junction will be exceeded. This method has been found satisfactory with transmitters operating at 50 watts d.c. input, which is somewhat higher than the usual mobile transmitter power level.

Transistor Protection

The r.f. stage transistor of a receiver operating in conjunction with a transmitter may be damaged by stray r.f. from the transmitter during transmission periods unless certain precuations are taken.

It is essential to employ an aerial relay in which the capacitance between the various sections of the contact set is as small as possible, and also to arrange for the contact which carries the receiver aerial load to be shorted down to earth during the transmit period. Additionally this lead must be fully screened, and earthed *only* at the relay end, and to the same point as the earthing contact. These points are shown in **Fig. 16.25(a).**

A low v.s.w.r. on the transmission line is most important, and thus the aerial matching must have close attention. Where the aerial is tuned by a remote a.t.u. then the coupling line between the transmitter and the a.t.u. must also be adjusted for the lowest possible v.s.w.r.

On the lower frequency bands, it is permissible to connect diodes across the first tuned circuit in such a manner that if any voltage is developed across this circuit due to stray r.f. from the transmitter, then it is shorted to ground. On the v.h.f. bands however, such diodes, unless biased by a polarizing voltage, may result in an unacceptable increase in receiver noise. This is illustrated in **Fig. 16.25(b).**

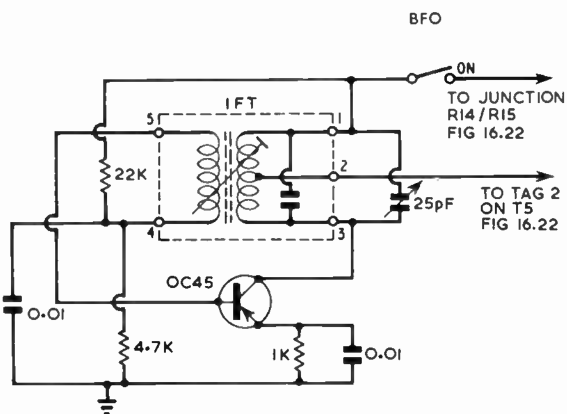


Fig. 16.24 B.F.O. circuit for use with i.f. amplifier unit of Fig. 16.22.

Greatest protection of the r.f. transistor will result if the base-emitter connection is open circuited during the transmission period, for the limiting factor is not then any current passing across this junction, but rather the potential developed across it due to the stray r.f. This is the course which is adopted in the case of the combined circuit given by Fig. 16.20–16.23 and Fig. 16.22.

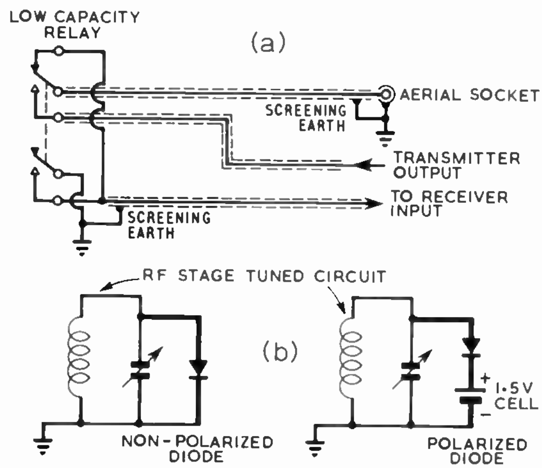


Fig. 16.25 Diagram (a) shows the recommended wiring of a low capacity relay, while diagram (b) illustrates non-polarized and polarized diodes connected across the r.f. stage tuned circuit.

MOBILE TRANSMITTERS

Due to the power restrictions imposed by the vehicle's electrical system, mobile transmitters are usually operated with a d.c. input to the power amplifier not exceeding 25 watts. Higher powers can be used, but unless special provisions are made, transmitter operating times needs to be restricted to the capabilities of the storage and charging system, particularly during the hours of darkness when the vehicle lighting circuits are in operation.

While no hard and fast rules can be laid down concerning how much current can be consumed by the transmitter and receiver installed in a specific vehicle, certain assumptions can be made, from which a reasonable maximum loading for that vehicle may be determined. The first, and most important assumption is that the storage and charging system is in good order and maintains a state of balance between supply and demand when the load of the vehicle is at maximum. For all practical purposes this load may be considered as the sum of the consumptions of the following items (*a*) Tail lights; (*b*) Side-lights and (*c*) Headlights dipped. In a typical case these amounted to (*a*) 12 watts; (*b*) 12 watts; and (*c*) 90 watts which added together produce a total of 114 watts, or a current of approximately 10 amperes in a 12 volt system. With an equivalent load, under daylight conditions, the electrical system of the vehicle should be capable of maintaining a state of balance. In fact, allowing for full headlight conditions, and adding in the loading of the ignition system, rear number plate light and a heater blower motor, the peak will be in the region of 160 watts for the vehicle cited. Thus, to assume a total available consumption of 120 watts maximum, i.e. a current of 10 amperes on a 12 volt system, will err on the right side.

In relation to this loading it will be appreciated that the electrical system will be overtaxed if the installation is used during darkness, and either the storage battery will have to be charged from an external source while the car is at rest, or alternatively, sufficient mileage covered during daylight without the equipment operating to allow the deficiency to be made up. To some extent, the extra loading can be compensated for by increasing the output of the generator, but where the loading is of the order of 120 watts, full compensation is not normally possible without the risk of overcharging the battery.

Where a great deal of mobile working is envisaged, and so long as there is adequate space under the bonnet or in the boot, consideration should be given to fitting an additional battery to run the installation. While this involves purchasing an additional battery, its cost may be preferable to being faced with a flat main battery. Where such a course is adopted, the auxiliary battery may be linked into the charging system of the car in the manner shown in **Fig. 16.26.** If this circuit is installed, the output of the generator may be increased without running into difficulty since the charging current is shared by both batteries when their state of charge is equal, but when unbalance arises, the lower battery will take the greater current.

In the case of valve operated equipment, not only is there the question of the current consumed by the heaters, but also the efficiency of the d.c. to d.c. converter system. Here, highest conversion efficiency is obtained by the use of a transistorized converter, with which an efficiency of at least 80 per cent is to be expected.

The majority of the power supply problems in mobile equipment can be solved by the use of transistorized installations. Although transistorized receivers are practical, transistor transmitters running any appreciable power are at present only an economic proposition to the radio amateur on the lower frequency bands. While transistors suitable for high frequency power amplifiers are available, their cost is high at the present time. There can be little doubt however that, as the prices fall, transistors will be preferred to valves.

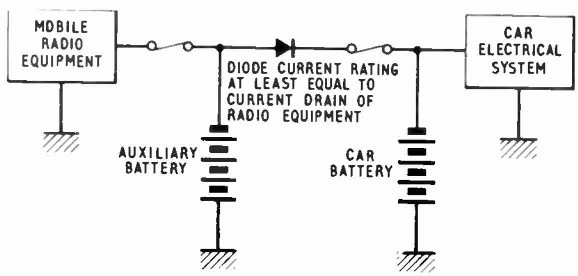


Fig. 16.26 Diagram showing how auxiliary battery can be added to the charging system of a car using positive earth. For negative earth systems, polarity of diode should be reversed.

Attention is currently being turned to hybrid transmitters, in which transistors are employed in the early stages, and valves are used only as high frequency multipliers, drivers and power amplifiers. In the main this approach has been used in the field of s.s.b. transceivers, but there is no reason why the same course should not be adopted for a.m. transmitters and transceivers.

10 Watt 1·8 Mc/s and 3·5 Mc/s Transmitter

This dual band transmitter is particularly suited to mobile operation in view of its low current consumption, the heater circuits taking 0·9 amperes at 12 volts, and the h.t. about 130 mA at 250 volts. Assuming that the h.t. converter shows an efficiency of 80 per cent, the total demand on the car system will be about 50 watts, or 4 amperes, during actual transmissions.

The circuit diagram is given in **Fig. 16.27.** It will be seen that the arrangement is similar to that of **Fig. 6.91** in Chapter 6 on page 6.55.

V1 which may be an EF91, Z77 or a 6AM6 functions in a Vackar v.f.o. circuit covering the fundamental range of 1·75 Mc/s to 2·0 Mc/s. Since, in order to cover both bands fully, the v.f.o. overtunes each band, separate scales should be provided on the v.f.o. for each range, and the band edges clearly marked. The usable v.f.o. tuning range in the case of Top Band, is from 1·8 Mc/s to 2·0 Mc/s, and for the UK 80m band, 1·75 Mc/s to 1·9 Mc/s corresponding to 3·5 Mc/s to 3·8 Mc/s.

V2, which is also an EF91/Z77/6AM6, operates as an untuned buffer when the output is on Top Band, and as a broadly tuned frequency doubler when the output is in the 80m band. In the original version of this transmitter, the drive to the p.a. was too high on 1·8 Mc/s when it was adjusted to the correct level on 3·5 Mc/s. This was corrected by wiring a damping resistor of 33 K ohms in parallel with the r.f. choke load of V2. If a discrepancy in drive is found between bands, a value of resistor should be chosen to equalize the drive.

The 6BW6 p.a. (V3), functions as a straight amplifier on both bands, and its tank circuit is a pi-network. The constants of this circuit have been selected to match down to impedances of 35 ohms and to cover the standard impedances of 50 ohms and 75 ohms. On the basis that the maximum ratio of input to output impedances in a pi-network should not exceed 100:1, the lowest actual output impedance which can be accommodated will depend on the impedance of the p.a. which is in turn related to the applied h.t. and the current drawn. At 250 volts h.t. and an input of 10 watts (40 mA anode current) the r.f. impedance is approximately 2900 ohms, and under these conditions, the network will accommodate load impedances of 35 ohms without difficulty. However, with an h.t. of 300 volts and an input of 10 watts, the p.a. r.f. impedance is of the order of 4100 ohms, and 35 ohm load impedances may not match. Thus where the load is of the order of 35 ohms, the p.a. h.t. should be kept down to 250 volts where the input has to be limited to

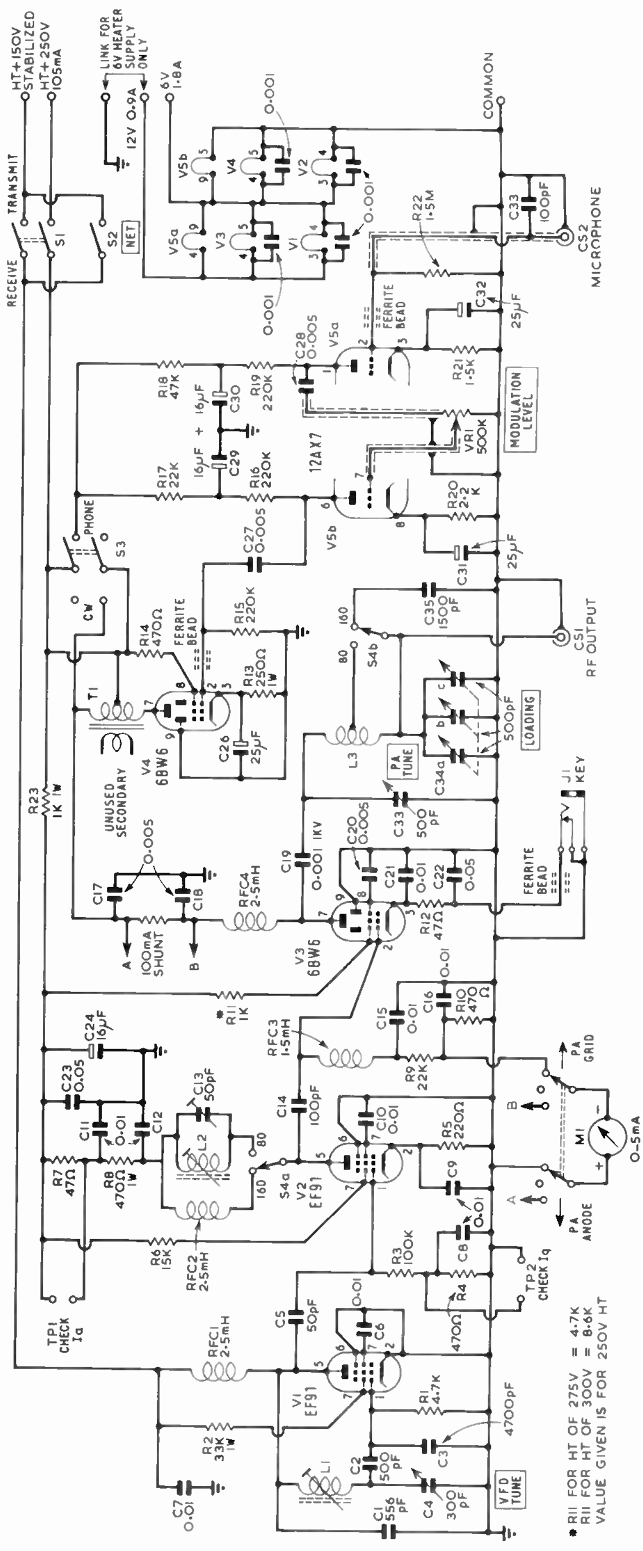


Fig. 16.27 1·8 Mc/s and 3·5 Mc/s dual-band mobile transmitter for AM/CW operation. The pi-output network will accommodate loads down to about 35 ohms impedance, and covers the standard impedances of 50 ohms and 75 ohms on both bands.

10 watts as in the case of the UK 1·8 Mc/s allocation.

Modulation is by V4, a 6BW6, in conjunction with the transformer T1. This transformer is a centre tapped output transformer, the loudspeaker winding of which is not used. This operates as a 1 : 1 modulation transformer in which one end of the secondary, and one end of the primary, are connected to a common h.t. supply. It should not be confused with Heising (sometimes called choke) modulation.

The microphone amplifier consists of both halves of V5, an ECC83/12AX7, the modulation gain control being positioned between the two sections. Particular attention should be paid to the steps which have been taken in the design to reduce the effects of stray r.f. entering the modulator. The ferrite beads should not be omitted from the positions shown, and if these are not available, they may be replaced by hollow ferrite cores intended for small coil formers.

While the control system indicates the use of switches S1 and S2, this may be modified to relay control by substituting **Fig. 6.92.** To determine the conditions in the p.a., a switched meter is fitted which monitors the grid current in one position and in the other the anode current. Where a yaxley type switch is used, a blank position should be left between the grid and anode current positions, in view of the potential difference between them.

So long as the layout follows good r.f. practice, no difficulties will be encountered, and, as a general basis, that shown in **Fig. 6.93** may be adopted. It should be noted however, that T1, together with V4 and V5 will have to be moved to the right to make sufficient room for the larger p.a. components. The three section loading capacitor, C34, is mounted below the p.a. tank tuning capacitor, C33, by cutting a suitable hole in the chassis within the area bounded by the p.a. compartment screen, and fitting it so that it is substantially below the level of the top of the chassis. The a.m./c.w. switch is moved to a position adjacent to the netting switch, and the bandswitch, S4, is mounted on the panel so that its spindle is in line with the centre of V3. The two sections of this switch are separated by a metal screen shaped like a step. One side of this screen runs up to the front panel, and the other arm runs past the right hand side of V3 when viewed from the top. The front section of S4, that is S4A, is fully screened from the output stage, and to this are connected C13, L2, and RFC2. The rear section, S4B, which is placed as near to the screen as possible, connects to the tap on the p.a. coil, the output connections and the fixed loading capacitor C35.

Details of unspecified components will be found in **Table 16.5.**

TABLE 16.5

Coil and transformer data for Fig. 16.27

Coil No.	Dia. in.	S.W.G. enam.	Turns close wound	Core fitted	Tapped
L1	$\frac{5}{16}$	34	70	Yes	Nil
L2	$\frac{3}{8}$	34	70	Yes	Nil
L3	1	18	56	No	28 turns

Transformer T1. Primary impedance to lay between 8 K ohm and 10 K ohm centre tapped and rated at not less than 50 mA. "Radiospares," "Hygrade" or "EL84" output transformers are suitable.

10 Watt Transistor Transmitter for 1·8 Mc/s

As indicated in the introduction to this section, transistorized transmitters show a substantially higher power conversion ratio than their valve operated counterpart. Not only is there no need to waste valuable current in valve heater circuits, but in addition, the power losses of a d.c. to d.c. conversion system are avoided.

The design of a 10 watt transistorized transmitter for the 1·8 Mc/s band, complete with modulator, will be found in **Fig. 6.87** of Chapter 6 (*H. F. Transmitters*). This equipment is an example of an "ideal" transistorized transmitter, since, operating directly from the 12 volt supply line, the percentage of the total input power realized as usable power output is at its highest.

15 Watt 28 Mc/s Transmitter

The circuit of a 15 watt transmitter for the 28 Mc/s band which incorporates optional crystal or v.f.o. control, a.m./c.w. operation, netting facilities, and a built-in aerial changeover relay is shown in **Fig. 16.28.** The heater circuits consume 1·2 amperes at 12 volts, and the h.t. approximately 120 mA at 250–300 volts, and thus this transmitter, and that described for the 1·8 Mc/s and 3·5 Mc/s bands could be interchanged between a common power supply unit.

V1 is an EF91/Z77/6AM6 used in a high C Colpitts v.f.o. circuit with a fundamental frequency of 7 Mc/s. The anode circuit is tuned to 14 Mc/s. A built-in stabilizer regulates the h.t. to the v.f.o. at 150 volts.

The second valve, V2 is a 6AH6, but this may be replaced by a 6AK6 without changing the performance, although if this latter type is used the heater wiring will have to be modified in the following manner. The 12 volt pilot lamp shown on the circuit diagram is removed and replaced by one rated at 6·3 volts 0·3 amps connected across the heater pins of the 6AK6.

V2 functions as a doubler from 14 Mc/s to 28 Mc/s, the anode circuit of which comprises a self-resonant coil L3. With this coil, the drive to the p.a. remains sensibly constant over a fairly wide frequency range, but not over the entire 28 Mc/s to 29·7 Mc/s allocation. For mobile operation where large excursions in frequency are not usually encountered, the specification for L3 will usually be satisfactory. However, if it is desired to move from one end of the band to the other, then L3 should be reduced to 8 turns and tuned by a 25 pF capacitor, the spindle of which should be coupled to a panel control.

A 5763 functions as the p.a., the tank circuit of which is a pi-network. As shown, the p.a. is not neutralized, and as long as the layout is carefully planned, regeneration should not be encountered. Due to their high slope, 5763 valves may sometimes show signs of parasitic oscillation. If this is experienced, a 10 ohm carbon resistor connected directly at the anode pin invariably effects a cure.

The modulator commences with one half of a 12AX7 connected as an earthed grid amplifier with the carbon microphone inserted in series with the cathode lead. This arrangement shows rather less gain than a conventional triode circuit, but has the advantage that the current flowing through the cathode circuit provides the energizing voltage for the microphone. This is followed by a 12AU7, the first half of which operates as a voltage amplifier, and the second half as a driver for the output stage. The modulator employs a 12AX7 in a zero bias Class B configuration, the resting

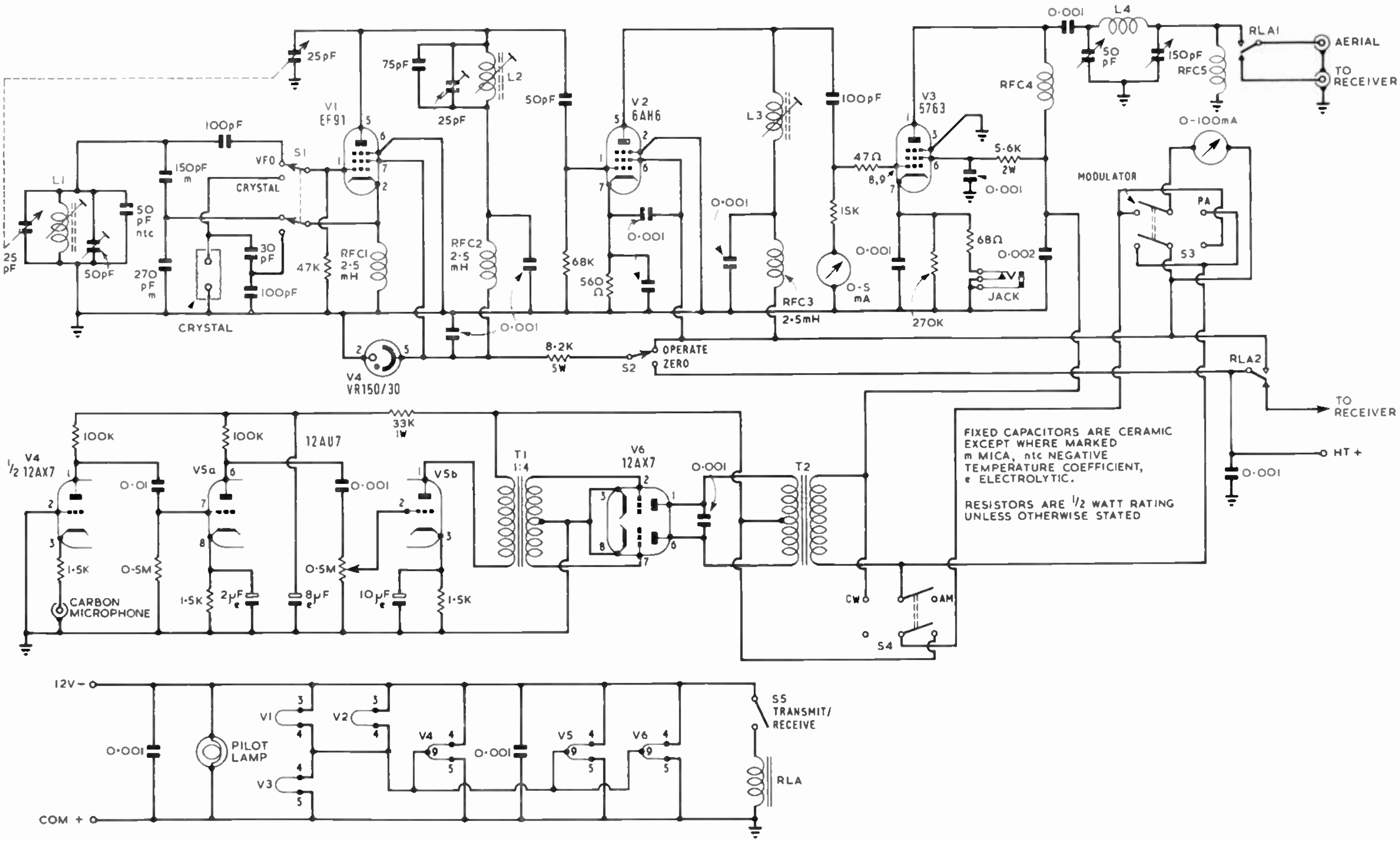


Fig. 16.28 Circuit diagram of a mobile transmitter for 28 Mc/s. The coils are wound as follows: L1, 22 turns 18 s.w.g. enam. on 1 in. dia. former, turns spaced diameter of wire; L2, 12 turns 24 s.w.g. on ⅜ in. dia. dust-iron former, length 3 in.; L3, 15 turns 24 s.w.g. on ⅜ in. dia. dust-iron former, length ¾ in.; L4, 8 turns 18 s.w.g. 1 in. dia. former, spaced diameter of wire; RFC4, 30 s.w.g. enamelled close wound 1½ in. long on ¼ in. polystyrene rod; T2, modulation transformer to match 10,000 ohms to 6000 ohms. Unless the leads to the 0·5mA meter can be kept very short, a 0·001μF capacitor should be connected across the meter to avoid instability.

current of which is 10 mA. On modulation peaks this rises to about 40 mA. This circuit is especially useful for mobile working due to its low idling current, and, provided that the h.t. smoothing capacitor has a value of not less than 50 μF, the h.t. voltage will remain substantially constant.

While individual meters are shown in the p.a. anode and grid circuits, a switched meter system similar to that of Fig. 16.27 could be substituted, although in this case it would not be possible to measure the modulator anode current. When switched to " Modulator " the anode current meter measures the sum of the currents drawn by V4, V5 and V6.

Correctly set up for a.m. operation, the p.a. grid current should run close to 3 mA, and when matched into an 8 ft. whip, the p.a. anode should load up to 45 mA.

By changing the v.f.o. frequency range, and making the necessary alterations to the tuned circuits, this basic transmitter design may be readily adapted for other frequencies.

Hybrid 70 Mc/s Transmitter

It has already been observed that while it is possible to construct a fully transistorized transmitter for the lower frequency bands, for the average radio amateur the cost of r.f. power transistors becomes too high as the frequency increases. In the v.h.f. region and higher, apart from transmitters which employ receiving type transistors with d.c. inputs in the milliwatt region, attention is currently directed to the hybrid designs. In these, valves are normally employed in the r.f. section, and transistors in the modulator and d.c. to d.c. converter.

The circuits of a hybrid transmitter, power supply, and control unit are shown in **Figs. 16.29, 16.30** and **16.31.** While this circuit specifically relates to the 70 Mc/s U.K. band, it may be adapted for the 50 Mc/s band by making appropriate changes in the crystal and tuned circuit frequencies.

Transmitters employed in v.h.f. mobile operation are invariably crystal controlled, and in the case of 70 Mc/s, operation within a particular area is confined to a specific frequency. Under such conditions it is possible to emulate the course adopted by the majority of mobile business radio systems, and to position the transmitter unit in the boot of the vehicle and govern its operation by a compact control unit at the driving position. Since space is always at a premium at the driving position, this method has much to commend it, and furthermore, as it is now common practice to mount v.h.f. aerials somewhere on the canopy surrounding the lid of the boot, it permits a shorter lead length between the aerial and the transmitter.

This transmitter is essentially two units, although shown as three diagrams. The first is the transmitter itself, which employs valves in the r.f. section, a transistor modulator, and a transistor d.c. to d.c. converter. Two frequencies are provided for the transmitter, and these are selected at the control position. The second unit is the control box, which, in addition to governing the function of the transmitter, contains a microphone pre-amplifier.

The design of the r.f. section was primarily governed by the p.a. valve and the desired d.c. input which was set at 15 watts maximum. A 5673 valve will operate at very close to this d.c. input at 70 Mc/s, and provide some 7 watts of r.f. output. The heater current of this valve is 0·75 amps. at 6·3 volts, and it was decided that this should be the total heater current consumed at 12 volts. The other types selected were an EF91/Z77/6AM6 and a 6BW6 wired in parallel, these also take 0·75 amps., and when this combination is placed in series with the 5763, the heater circuit may be directly connected across a 12–14 volt supply without the use of any ballast resistors.

The circuit of the r.f. and modulator sections is shown in Fig. 16.29. V1, operates as a Colpitts crystal oscillator of conventional design. Crystals in either the 7 Mc/s or 11 Mc/s range may be employed, multiplying by nine or six respectively. Whichever frequency crystal is employed, the anode of the crystal oscillator is always tuned to 23 Mc/s, and therefore with 7 Mc/s crystals this resonates at the third harmonic, and with 11 Mc/s crystals, at the second harmonic. The drive to the following stage will depend on the harmonic of the crystal used. The value given for C4 in Fig. 16.29 is the nominal value which gives correct drive to the succeeding stage when 11 Mc/s crystals are employed. Where 7 Mc/s crystals are used, depending on the activity of the crystal, C4 may have to be increased to 150 pF. During the setting up of the transmitter, C4 is adjusted in value until the drive to V2 is not less than 0·8 mA, not more than 1·5 mA.

V2, a 6BW6, operates as a frequency tripler, the anode circuit being tuned to 70 Mc/s. To attenuate unwanted crystal harmonics, link coupling is employed between this circuit and the grid of the p.a.

The p.a. is V3, a 5763, which operates as a neutralized amplifier. Although neutralization is shown on the circuit diagram, and provided for in construction it may not be needed. Nevertheless the provision should not be omitted. Due to its very high slope, the 5763 is prone to parasitic oscillation, and for this reason R11 is shown connected between the anode and the p.a. tank circuit. Initially this resistor should be omitted and only wired into position if parasitic oscillation is experienced. While a value of 10 ohms is shown, a greater or lesser value may be required. The lowest possible value resistor should be used and it must be fitted to the anode connection of the valve base with nil lead length. The resistance should not exceed 47 ohms.

A dividing screen must be fitted across the base of the 5763 so as to shield the anode and grid connections from each other. This screen should form one side of a box enclosing the p.a. tank circuit components, and be connected to the centre spigot of the valve base. Increased stability of the p.a. will result from using a valve base with a skirt although the valve itself must not be fitted with a screening can. To further reduce possible coupling between the p.a. tank and grid circuits, the coil assemblies, L3 and L4, are mounted so that they are at right angles to each other.

To make provision for neutralizing, the coil L4 is run parallel with the screen across the base of the 5763. In this screen, and adjacent to the end of L4 remote from the anode, a small hole is drilled and fitted with grommet. Neutralizing, if required, is achieved by running a well insulated wire from L4 through this hole, back to the proximity of the grid pin of the 5763. Adjustment of the position of this wire relative to the grid pin, and its wiring, varies the neutralizing capacitance.

Three metering positions are provided. Test point TP1 measures the drive to the 6BW6 tripler from the crystal oscillator, and the current measured at this point should lie between 0·8 mA and 1·5 mA. The p.a. grid current, and the p.a. anode current are measured across points Q and R, the circuit being checked is selected by a d.p.d.t. toggle switch.

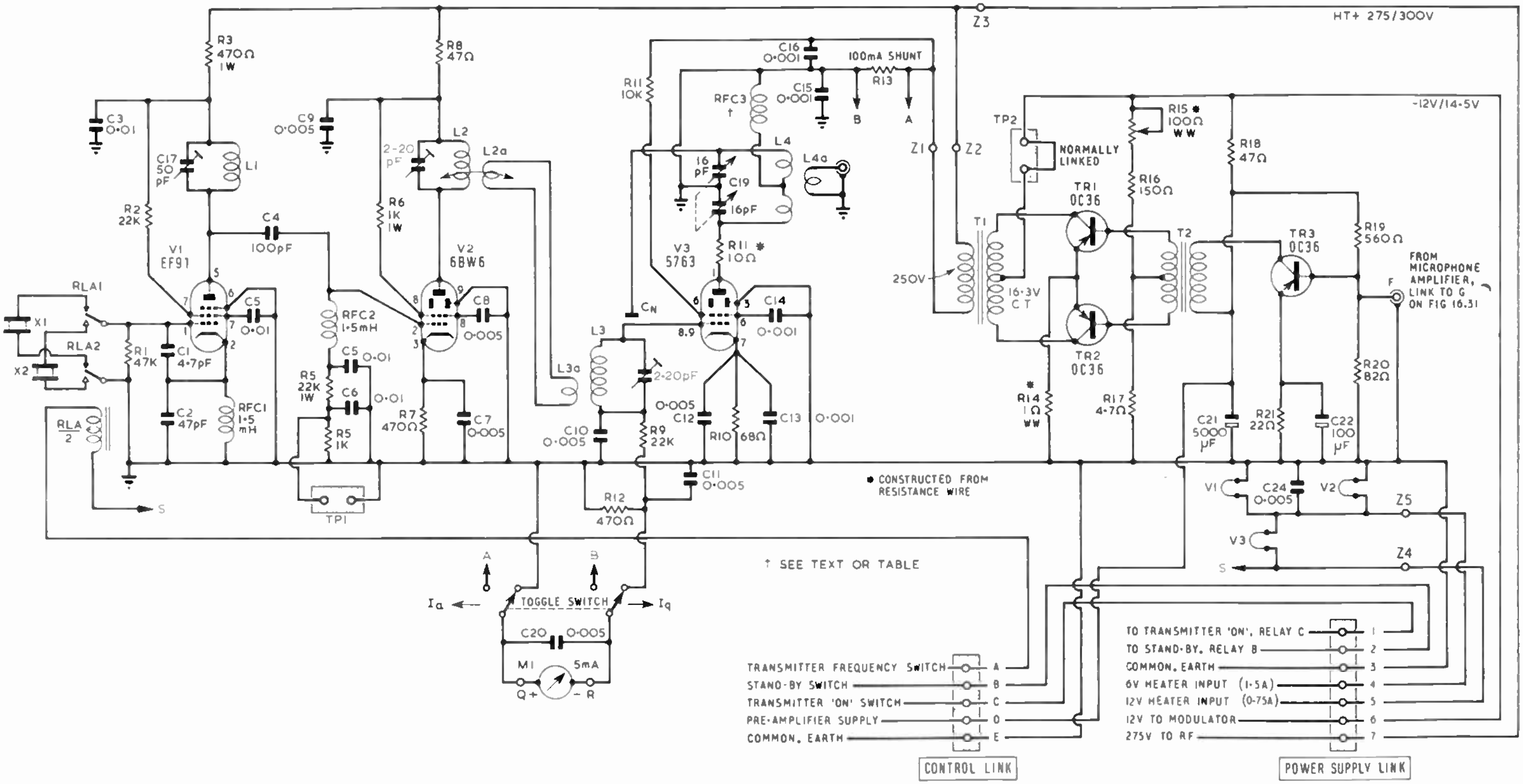


Fig. 16.29 Hybrid transmitter for 4 metres. This unit is designed for remote control and features the choice of two operating frequencies and a fully transistorized modulator. Normally the power supply unit would be constructed in with the transmitter but for clarity, it is shown as a separate unit in Fig. 16.30. The microphone pre-amplifier is housed in the control unit shown in Fig. 16.31. Points identified by the letter Z relate only to the 144 Mc/s hybrid transmitter. See relevant text.

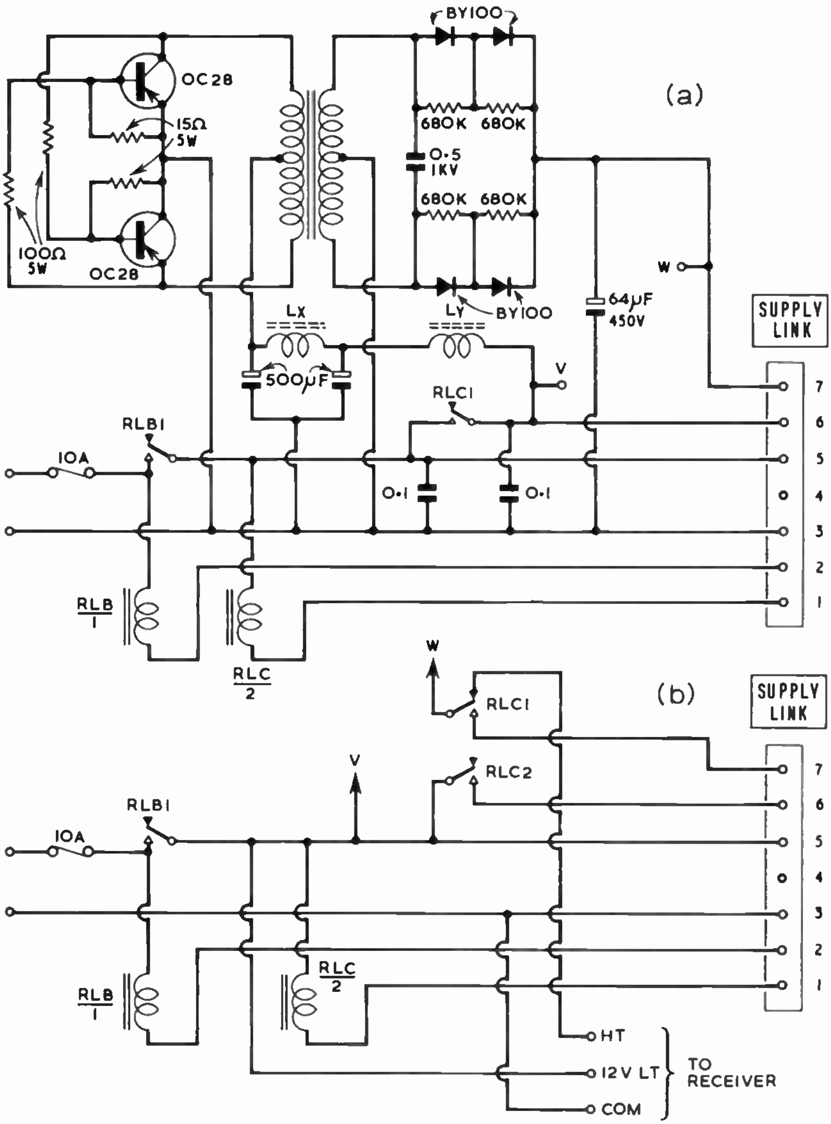


Fig. 16.30 Transistorized d.c. to d.c. converter for use with 4 metre hybrid transmitter. In Fig. (a) above the converter operates the transmitter only. If it is desired to employ the output to operate a receiver, the control circuits should be modified to that shown in (b), circuit (a) being broken at the points V and W and the lower arrangement substituted. In the interests of economy, the circuit employs a standard vibrator transformer.

The meter used has a basic sensitivity of 5 mA which is shunted to 100 mA when reading the p.a. anode current. This meter may either be fitted to the transmitter enclosure, or alternatively, points Q and R terminated on small sockets. Which course is adopted will depend on personal choice, and whether the transmitter is constructed as a unit mounting in the vehicle boot, or as a more conventional assembly.

The tag strip shown as " Power Supply Link " need not exist where the power supply is built in with the transmitter assembly, although the provision of a main power distribution tag strip does simplify checking the various supply voltages. The control link, on the other hand, may be any convenient plug and socket arrangement.

The modulator circuit is capable of developing 12 watts of audio power, which is far greater than needed by the r.f. section of this transmitter. Two OC36 transistors operated in Class B are supplied direct from the 12–14 volt line. The modulation transformer is a low voltage mains transformer the primary of which is wired in series with the h.t. supply to the p.a. The modulating transistors are connected across the low voltage secondary. This transformer will provide a reasonable match to power amplifiers with a d.c. impedance of between 4 K ohms and 6 K ohms. The driver transformer does have to be hand wound, but as will be seen from the components table, this is not a difficult task.

The modulating transistors, TR1 and TR2, must be mounted on an adequate heat sink. The actual chassis itself can be used for this purpose, but where this course is adopted, they should be positioned away from the p.a. valve to reduce the heat which they receive by conduction through the chassis. Under normal conditions the transistors operate at a temperature which makes them only just warm to the touch.

The driver transistor is a further OC36, and since this operates well within its power rating, it could be free air mounted. Mechanically however it is simpler to attach it to the chassis. The pre-amplifier stages of the modulator are contained within the control box.

H.T. for the transmitter r.f. section is derived from a transistorized d.c. converter, the circuit of which is shown in Fig. 16.30. This unit employs a standard vibrator transformer as these are not so expensive as toroid units, and in addition may be obtained as " surplus." The transformer used in the B44 transmitter-receiver is particularly suitable for use in this circuit. It will be seen that two control arrangements are given for the power unit. That illustrated in Fig. 16.30(a) is for use when the converter only supplies h.t. to the transmitter. In this case, relay RLB applies power to the heater circuit of the transmitter when it closes and places it in the stand-by condition. Either the Transmit or Press-to-Talk switches will actuate RLC, causing power to be

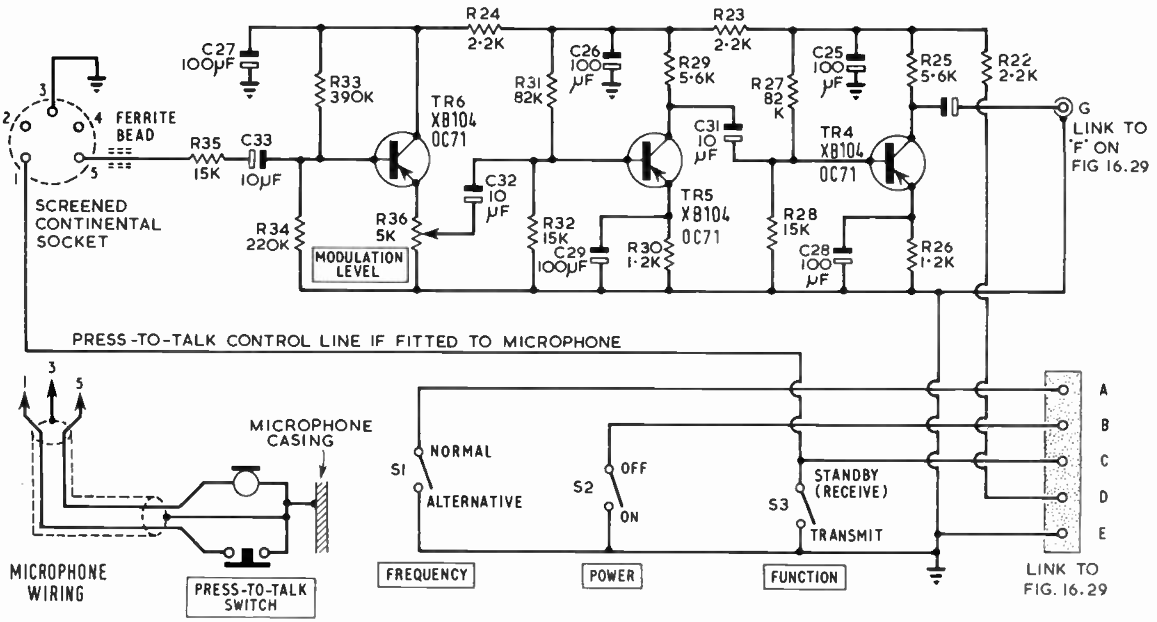


Fig. 16.31 Control unit and microphone pre-amplifier for 70 Mc/s hybrid transmitter. The pre-amplifier is designed to operate with a high impedance crystal microphone. The twin screened microphone cable is required if p.t.t. operation is incorporated. This may be omitted, and in which case, S3 operates the transmitter.

applied to the d.c. to d.c. converter in order to provide h.t. to the transmitter circuits.

Where the same converter is to be used to supply h.t. to an associated receiver, the control circuits are modified to those shown in **Fig. 16.30(b).** In this case, when RLB operates, power is applied to the heater circuits and the d.c. to d.c. converter at the same time. RLC, which is changed to a d.p.c.o. type, switches the h.t. between the transmitter and the receiver and also applies 12 volts to the transistorized modulator circuits on " Transmit."

TABLE 16.6

Hybrid 70 Mc/s Transmitter component data

L1	18 turns 24 s.w.g. enam. close wound on 5/16 in. dia. former.
L2	6 turns 20 s.w.g. enam. close wound on 5/16 in. dia. former.
L2A	2 turns 20 s.w.g. enam. at earthy end of L2 arranged so that it slides along former so varying spacing from L2.
L3	3½ turns 20 s.w.g. enam. close wound on 5/16 in. dia. former.
L3A	2 turns 20 s.w.g. enam. adjacent to earth end of L3.
L4	3½ turns either side of centre wound on ¾ in. mandrel with 16 s.w.g. tinned copper wire. Turns spaced wire diameter. Gap in middle for L4A should be ⅜ in. Link L4A 2 turns 16 s.w.g. sleeved.
RFC3	39 in. 34 s.w.g. enam. close wound on 1 Megohm watt resistor.
T1	"Radiospares" heater transformer. Primary 240V. Secondary 16·3V at 0·3 amp.
T2	Core size ⅜ in. square laminations. Wind secondary first which consists of 5 bifilar turns—that is 10 turns in total of 24 s.w.g. Insulate with two layers p.v.c. tape. Wind primary over secondary using 200 turns 30 s.w.g. Wire is enam.

In power pack. Fig. 16.30

Lx and Lz. "Radiospares" 3 amp. TV suppressor chokes.

Transmitter control and modulator pre-amplifier circuits are shown in **Fig. 16.31.** These can be constructed as a very compact unit, for which space may easily be found adjacent to the driving position. The microphone pre-amplifier is designed for a crystal microphone input and uses an emitter-follower arrangement. The modulation gain control is in the emitter of this stage. This is followed by a two stage amplifier, the output of which is linked by a single screened lead to the input of the modulator driver transistor in the main transmitter unit. Three single-pole toggle switches control the various operations.

Setting up the transmitter follows standard practice. Drive to the p.a. should lie between 2·5 mA and 3 mA and is governed by adjusting the position of L2A in relation to L2, while the p.a. anode current should be between 45 mA and 50 mA when the stage is fully loaded. Prior to the application of power to the transmitter, the tuned circuits should be approximately set to their correct frequencies by the use of a g.d.o. Where such an instrument is not available, it is essential to check the frequencies of the various circuits under operating conditions with an absorption wavemeter.

In matching the output to the aerial, the use of an s.w.r. bridge is strongly advised. If it is difficult to arrive at satisfactory loading, link reactance may be tuned out by fitting a 50 pF variable capacitor in series with the earthy end of L4A, but such a course should only be adopted after experimenting with the position of L4A in relation to L4.

The standing current of the OC36 modulators should be checked by removing the link from TP2, and connecting across this test point a meter set to its 500 mA range. R15 is adjusted until the current lies between 50 mA and 60 mA with no input to the modulator. At the full capability of the modulator, which is not required for this transmitter, the

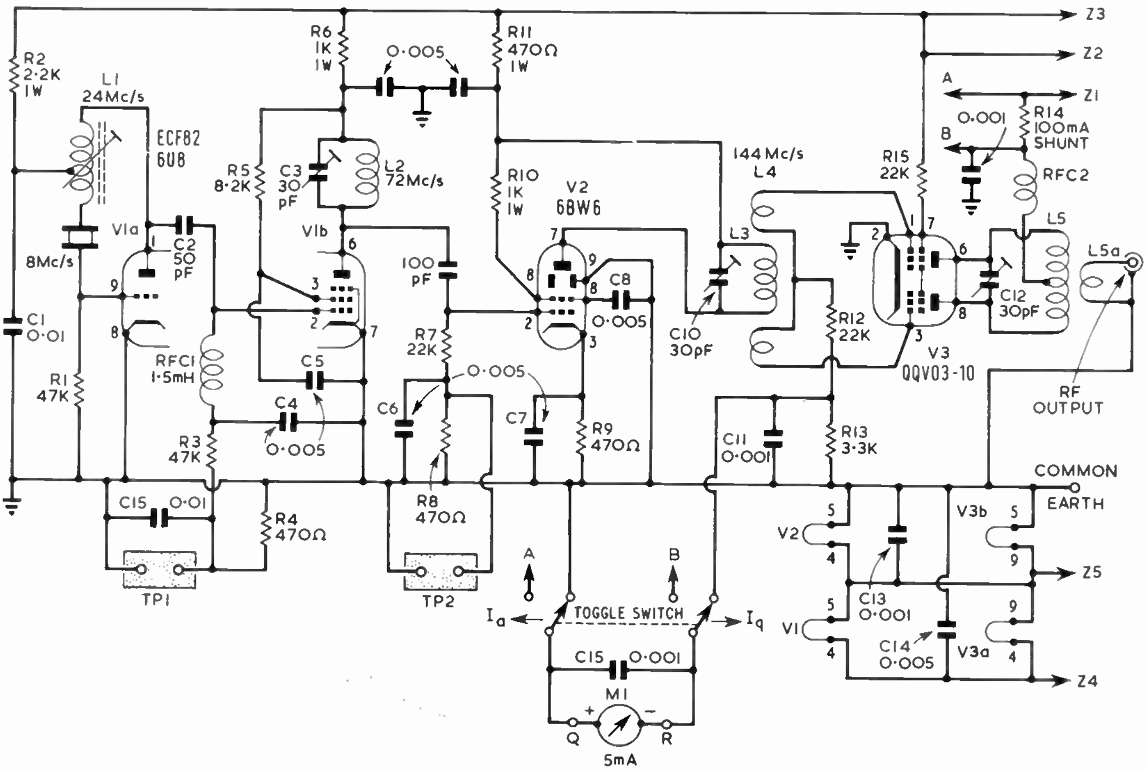


Fig. 16.32 144 Mc/s hybrid transmitter. The complete transmitter is given by substituting this r.f. section for that shown in Fig. 16.29. Terminations coded with the letter Z are connected to the points identified in a similar manner in Fig. 16.29.

peak current swings up to 1·75 amps., at which the audio output will be approximately 15 watts.

Hybrid Transmitter for 144 Mc/s

Apart from the changes which are necessary in the r.f. section, the modulator, power supply and control circuits of the 70 Mc/s hybrid transmitter may be utilized for a 2m transmitter of similar design and power output. With a little ingenuity in the method of mechanical construction, matters can be so arranged that the two r.f. sections are interchangeable almost on a plug-in basis. For mobile applications this is useful since it permits the construction of equipment having dual band capability, and while the changing of bands involves the exchanging of the sub-assemblies, this arrangement is worthy of consideration since it is rare that twin aerial installations are fitted to a vehicle.

In the prototype of a transmitter using this method, the area of the main chassis covered by the r.f. sections was entirely removed. Each r.f. section was constructed on a 16 s.w.g. aluminium plate that overlapped the chassis aperture by ½ in. all round. To the chassis, and along the overlap, were fitted captive 2B.A. bolts with their heads on the underside, and matching holes were made in the r.f. plates. These plates were secured in position by 2B.A. wing nuts, and the electrical connections made by a free plug from the r.f. plates to an octal socket on the main chassis. Exchanging r.f. sections can be accomplished within two minutes comfortably. If this course is adopted, the underside of each r.f. plate must be fully enclosed to prevent damage to the components when a particular r.f. unit is stored.

The circuit of the 144 Mc/s r.f. section is shown in **Fig. 16.32.** In this, the points marked Z are connected to the similarly identified points on the circuit of Fig. 16.29. Control line A on the Contol Link Panel of this circuit is not used.

TABLE 16.7

Hybrid 144 Mc/s Transmitter component data

L1	31 turns 28 s.w.g. enam. close wound on ¼ in. dia. former tapped at 7 turns from grid end. Former fitted with ferrite core.
L2	6 turns 18 s.w.g. close wound on ¼ in. dia. mandrel.
L3	2 turns 18 s.w.g. 7/16 in. dia. turns spaced wire diameter. This winding is placed in the centre of L4.
L4	2½ turns either side of centre of 16 s.w.g. Overall length, including gap at centre to accommodate L3, ¾ in. Diameter 7/16 in.
L5	4 turns of 14 s.w.g. tinned copper/silvered copper ¾ in. dia. Place 2 turns either side of a small centre gap to take L5A. Overall length, including gap 11/16 in.
L5A	2 turns 18 s.w.g. close wound insulated with thin p.v.c. sleeving wound on ⅜ in. dia. mandrel.
RFC2	19 in. 34 s.w.g. enam. close wound on 1 Megohm 1 watt resistor.

One half of V1, an ECF82/6U8 functions as a Squier overtone oscillator producing output on the third harmonic of the crystal frequency. The second half of this valve operates as a tripler to 72 Mc/s. V2 operates as a doubler to 144 Mc/s, and while there are better valves than the 6BW6 for this purpose, in this instance it provides adequate drive to the p.a., and its heater current is the same as that of V1 which permits the two to be wired in series across the 12 volt supply. The p.a. employs a QQVO3-10 in a push-pull circuit in which it will be seen no provision is made for neutralization. The QQVO3-10 valve is internally neutralized during manufacture, and thus no external circuits are needed.

L4 resonates with the valve capacitance. The spacing between its turns may be varied to give maximum drive. If an iron dust core partially inserted in L4 increases the drive the spacing should be reduced if, on the other hand, the drive increases when a brass slug is inserted, the spacing should be increased. Drive may also be improved by varying the coupling between L3 and L4.

Tuning up is achieved by monitoring the grid current of the following stage, and at the same time, verifying the frequency in the stage being tuned by means of an absorption wavemeter. Typical currents to be expected are: TP1, not less than 0·5 mA and TP2 not less than 1 mA nor more than 2 mA.

Drive to the p.a. under loaded conditions should run at not less than 2 mA, nor more than 2·5 mA. Under no circumstances must the anode current exceed 80 mA, nor the supply line exceed 260 volts. At an h.t. of 250 volts, the p.a. should be loaded to 60 mA. At these figures, the p.a. will match the modulator impedance.

Dual Band Transmitter for 1·8 Mc/s and 144 Mc/s

While multiband mobile operation is not usual, it is practical in the case of the 1·8 Mc/s and 144 Mc/s (or 70 Mc/s) bands in view of the compact nature of the aerial installation on the v.h.f. band.

For mobile operation, where space is at a premium, the increase in size made necessary by bandswitching can, at times, be unacceptable. An ingenious method of coupling h.f. and v.h.f. tuned circuits due to D. W. Furby G3EDH makes possible a transmitter which will operate on either

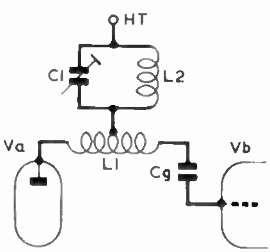


Fig. 16.33 Combined v.h.f. and h.f. interstage coupling arrangement employed in the dual band transmitter.

1·8 Mc/s or 144 Mc/s without the complications of mechanical band switching in the main circuits. By using this system, a compact dual band transmitter may be constructed.

The basic circuit coupling arrangement is shown in Fig. 16.33. The v.h.f. tuned circuit is L1 which in conjunction with the output capacity of Va, and the input capacity of Vb, operates as a balanced coupler. Since L1 is resonated by

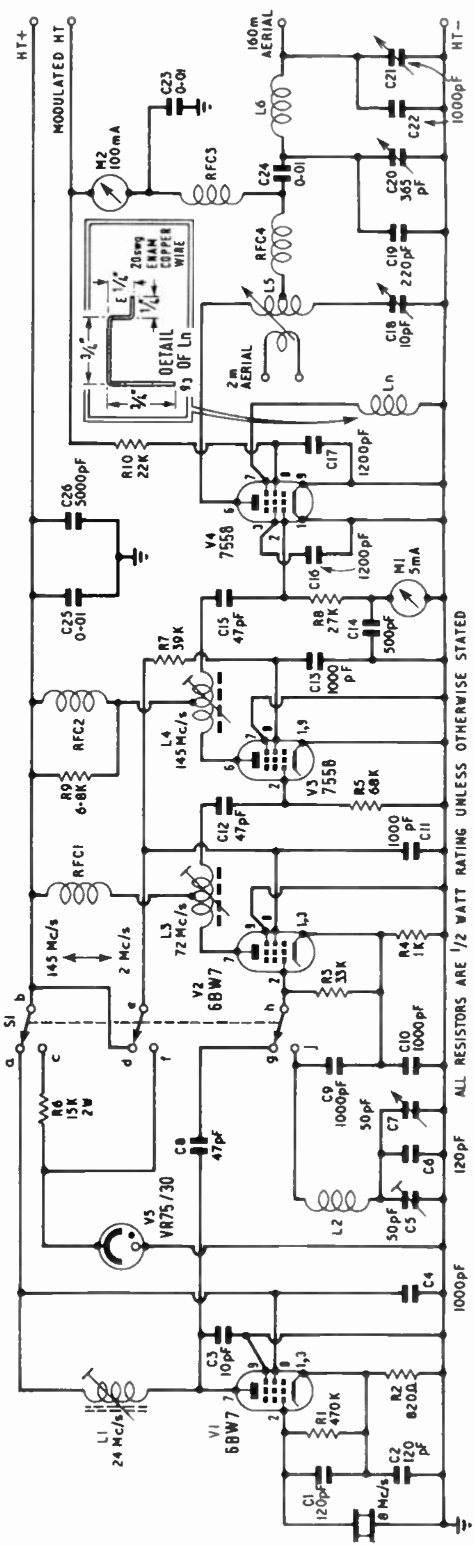


Fig. 16.34. Circuit diagram of the dual band transmitter for operation on 1·8 Mc/s and 144 Mc/s. See text for alternative crystal oscillator circuit.

these capacitances in series, the net capacitance will be very small and this in turn will allow the use of a comparatively large coil at L1. The result of this will be that the circuit will be relatively broad-band at v.h.f., consequently tuning by a manual capacitor is not needed. At approximately the centre of L1 will be a point of zero potential, and to this point is connected the h.f. tuned circuit C1/L2. At h.f., L1 behaves as a piece of wire connecting the anode of Va to the circuit C1/L2 and to Cg, the coupling capacitor to the following stage. When Va is fed at v.h.f., the L1 network becomes effective, C1/L2 having no effect since they are connected to its point of zero r.f. potential.

The circuit of the r.f. section of a transmitter employing this principle is shown in **Fig. 16.34** and this will run up to 15 watts on 144 Mc/s.

Taking the operation of the circuit on 2m first, V1 functions as a Colpitts crystal oscillator using 8 Mc/s crystals and giving output on 24 Mc/s. With S1 in the 144 Mc/s position this output is coupled to V2 which functions as a frequency

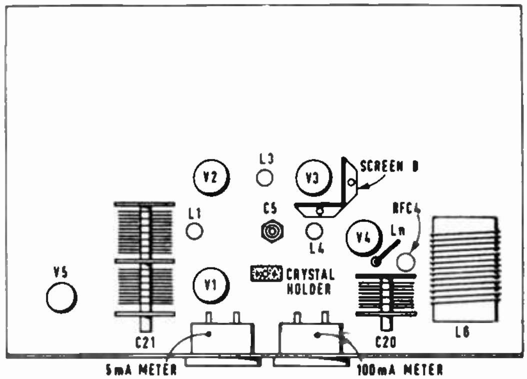


Fig. 16.35 Layout of the principal components of dual band transmitter.

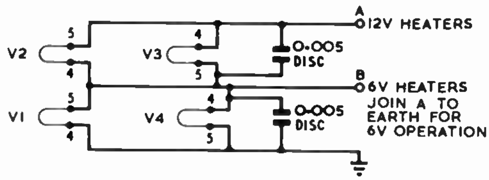


Fig. 16.34A. Heater circuit for the dual band transmitter.

tripler to 72 Mc/s. In the 1·8 Mc/s position of S1, V2 operates as a v.f.o. and this accounts for the somewhat unusual cathode circuit. In fact, when V2 is fed from the crystal oscillator, it operates without any cathode bias since the grid leak, R3, is returned to the cathode, and it therefore depends on the drive from the crystal oscillator to generate drive bias. Because of this, care should be taken to ensure that when the switch S1 is in the 144 Mc/s position, a crystal is always in position in the grid of V1. In the event of the drive to V2 failing under 144 Mc/s conditions, the high value resistor R4 in the d.c. path to the cathode will give a limited degree of protection for a brief period, for, as the cathode current increases the voltage drop across R4 will also increase, and effectively reduce the h.t. across the valve. Nevertheless, loss of drive to V2 is to be avoided.

Output from V2 is coupled to V3 and this operates as a doubler. The p.a. V4 functions as a straight amplifier on

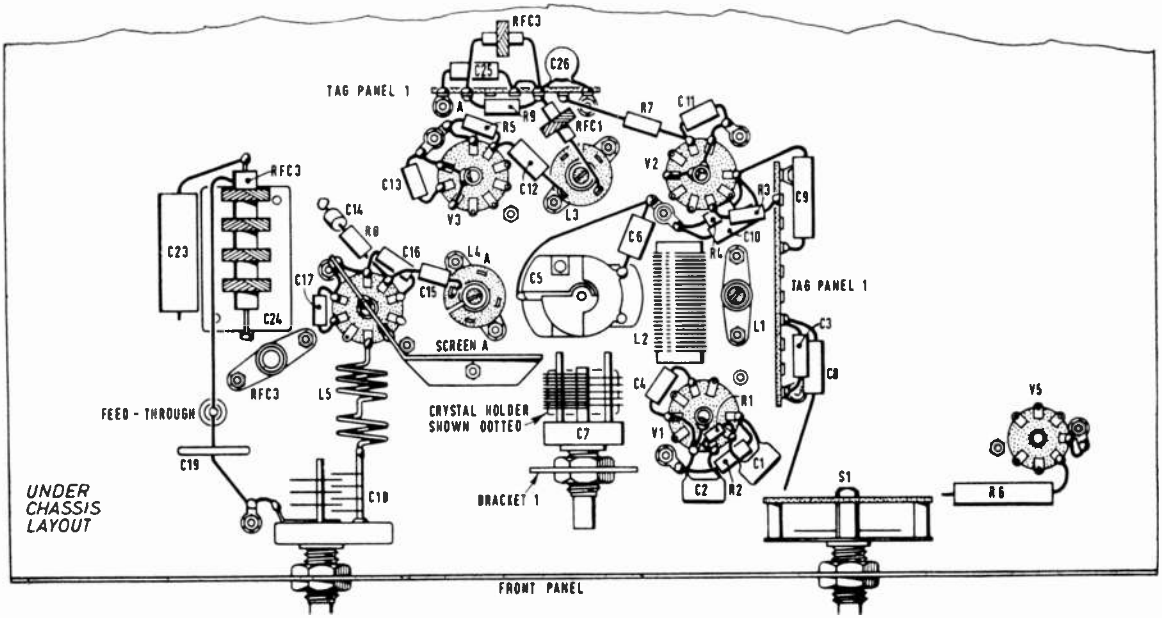


Fig. 16.36. Layout of under-chassis components.

both bands. On 144 Mc/s this is screen grid neutralized by means of the inductance Ln in the screen grid lead. The construction of Ln is shown in mechanical detail on the diagram, and this must be carefully copied.

With S1 set in the 1·8 Mc/s position, the stabilizer V5 is brought into operation and h.t. applied to V2, but not V1. V2 functions as a Clapp v.f.o. covering the frequency range 1·8 Mc/s to 2·0 Mc/s. The anode load for this valve is the r.f. choke RFC1 connected to the centre of the 72 Mc/s coil L3. V3 operates as an untuned buffer, again loaded by an r.f. choke, this time with RFC2 connected to L4. In order to adjust the level of the drive to the p.a. this choke is parallel loaded by the resistor R9. If chokes other than those specified are fitted, it will probably be necessary to experiment with the value of R9 in order to arrive at the correct grid drive to the p.a.

On 1·8 Mc/s the p.a. output circuit is a conventional pi-network connected to the centre of the v.h.f. output circuit, L5, but isolated from it by the v.h.f. r.f. choke, RFC4. Low frequency isolation for the 1·8 Mc/s pi-network is given by the r.f. choke RFC3. Anode and grid circuits of the p.a. are metered in the usual manner.

Examples of the layout employed in the original transmitter are shown in **Fig. 16.35** and **Fig. 16.36**, but for mobile operation a more compact arrangement might be desirable, and in this connection, the circuit lends itself to mechanical adaption always provided than L3, L4 and L5 are positioned to avoid interaction. It may be of interest to note that the d.c. to d.c. inverter shown in Fig. 16.30(a) will power this transmitter in a satisfactory manner, and in addition, that the transistorized modulator of Fig. 16.29, in conjunction with the pre-amplifier of Fig. 16.31 will modulate it fully.

Since self-bias is not used throughout this transmitter, initial tuning must be carried out stage by stage. The first step is to set up the transmitter for 144 Mc/s operation. A 6 volt heater supply must be used during the following adjustments. Turn S1 to the 144 Mc/s position, insert a suitable 8 Mc/s crystal, and fit V1 and V2. Disconnect the earthy end of R3, and fit a 5 mA meter between this and the cathode of V2. Apply power to V1 only. Tune L1 for maximum current in the meter, about 1·2 mA, and with an absorption wavemeter, verify that L1 is tuned to 24 Mc/s. Reconnect R3, and restore power to V2. Insert V3, and disconnect *(a)* the h.t feed to V3 and *(b)* the earthy end of R5. From the free end of R5 connect a 5 mA meter to earth. Apply power to V1 and V2. Tune the slug of L3 to give maximum indication on the meter, about 1 mA, and verify that L3 is tuned to 72 Mc/s. Restore R5 and the h.t. supply to V3. Insert V4 and remove the h.t. feed. Apply power to V1, V2 and V3. Tune L4 to give maximum indication on the p.a. grid current meter, about 2 mA, and verify that L4 is tuned to 144 Mc/s. Restore the h.t. to V4.

Apply h.t. to the whole transmitter, and rapidly tune C18 for maximum dip in the p.a. anode current. Remove the h.t. Bring an absorption wavemeter close to L5, apply h.t. again, and verify the frequency as 144 Mc/s. With either an 80 ohm dummy load, or a 144 Mc/s aerial connected to the link of L5, apply power, and with an insulated tool, move the link into L5, periodically checking the dip in the p.a. anode current by adjusting C18. Correctly loaded the p.a. current stands at about 60 mA. Final adjustment of the link should be carried out with the transmitter installed in the vehicle. Finally L4 is re-adjusted for maximum attainable grid drive to the p.a.

In the case of 1·8 Mc/s operation, initially only V2 and V3 are fitted, power applied, and the v.f.o. tuning range checked on a suitable receiver. With C7 at minimum capacitance, adjust C5 until the v.f.o. is precisely on 2 Mc/s. Tune C7 to maximum capacity and verify the lowest frequency to which the v.f.o. tunes. This will be below 1·8 Mc/s. Reduce the inductance of L2 by removing one turn at a time until the v.f.o. is on 1·8 Mc/s. As turns are removed from L2, so the highest frequency to which the v.f.o. tunes will rise. After each adjustment to L2, tune C7 to minimum capacity, and with C5 restore the frequency to 2 Mc/s. Continue the process until the 1·8 Mc/s–2 Mc/s band covers most of the tuning range.

Fit all valves, switch to 1·8 Mc/s, set C21 to maximum capacitance, apply power, and tune C20 for maximum dip in p.a. anode current. Verify that the p.a. grid current is 2 mA. If r.f. chokes other than those specified have been fitted in positions RFC1 and RFC2, it may be necessary to experiment with the value of R9 to secure this grid current.

In the crystal oscillator circuit, it will be noted that the effective capacitance in parallel with the crystal is somewhat higher than usual, being approximately 60 pF. If difficulty is experienced due to the final frequency being lower than the calculated value on 144 Mc/s, the circuit may be amended to the more conventional arrangement by taking the following steps. Reduce C1 to 30 pF. Reduce C2 to 100 pF. Reduce R1 to 47 K ohm and connect between grid (pin 2) of V1 and earth. Replace R2 by a 1·5 mH RFC. Take R2 and place it in series with the h.t. to L1, and at the junction of L1 and R2 connect a 0·005 μF disc ceramic capacitor to earth.

There is no reason why the interstage coupling method used in this transmitter should not be applied to other frequencies. Experiments have indicated that it works in a satisfactory manner with the following combinations *(a)* 144 Mc/s–3·5 Mc/s; *(b)* 70 Mc/s–1·8 Mc/s; *(c)* 70 Mc/s–3·5 Mc/s. In addition, a tuned circuit at the lower frequency may be connected to centre of the high frequency coil as shown in Fig. 16.33, and the system is not restricted to the r.f. chokes shown in Fig. 16.34.

H.T. FOR MOBILE EQUIPMENT

There are three methods of obtaining the h.t. required for valve operated equipment from the 12 volt vehicle system, and these are, in ascending order of conversion efficiency, (1) Dynamotor; (2) Vibrator; and (3) Transistorized inverter.

Dynamotor

A dynamotor is a motor driven d.c. generator in which the motor and generator windings are combined on a common shaft. In some cases the generator section may be double

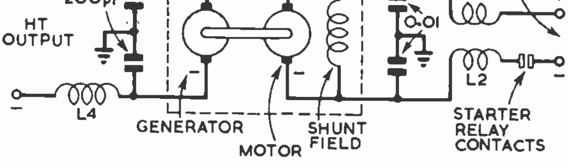


Fig. 16.37 Typical dynamotor suppression circuit. The values shown are for equipment operating between 60 Mc/s and 150 Mc/s. L1/L2 each 11 turns 16 s.w.g. cotton covered on ½ in. dia. former. L1 is wound between the turns of L2. L3/L4 v.h.f. r.f. chokes each being a ¾ in. long winding of 32 s.w.g. enam. on ¼ in. dia. former.

wound so as to provide two output voltages, one of which is usually in the range 250–300 volts, and the other 450–600 volts.

Since current has to be provided from the 12 volt source to run the motor under all conditions, the power conversion efficiency is not as high as is possible with other methods, particularly under low loading conditions. The idling current of a typical 12 volt motor generator may range between 2 amps to 4 amps., this current increasing as the load is applied to the generator section. Highest efficiency in a dynamotor is usually secured when the output is in the region of the maximum load specified, and with a well designed unit usually lies between 50–55 per cent.

Where the output voltage and current of a dynamotor are known, it is possible to approximate the maximum current that the unit will take from the d.c. source. The output wattage rating should be calculated, and assuming a conversion efficiency of 50 per cent, doubled to give the maximum input wattage.

The prime advantage of the dynamotor is its high reliability factor, always provided that it is correctly serviced, and for this reason, and despite the availability of more efficient systems, it is still favoured under certain circumstances. However, with the increasing reliability of other systems, particularly transistorized inverters, the days of the dynamotor are numbered.

Like all motors, the dynamotor can generate substantial interference unless correctly suppressed. The circuit of an HT31 dynamotor and its suppression circuits are shown in Fig. 16.37, and this is typical of the arrangements for this type of unit.

Maintenance of a dynamotor should be limited to replacing the carbon brushes when they wear down to about ¼ in. Oil should never be applied to the bearings. If the bearings appear to be running dry, they should be dismantled, the old grease removed, and repacked with high pressure grease. Oil will destroy the properties of the lubricant, and produce faster wear than that which would be experienced if the bearings were left untouched in most cases.

Vibrator Supplies

These units rely on a vibrator reed in series with the d.c. supply to the primary of a transformer. This reed carries one contact which makes and breaks with a fixed contact, and, as a result, the primary of the transformer receives pulsed d.c. The secondary of the transformer steps up this pulsed d.c. in the usual manner.

The vibrator units themselves fall into two general categories. The first, and the simplest, consists of the simple reed arrangement together with its driving coil. The second type is fitted with an additional contact set wired in such a manner that they reconvert the pulsed d.c. of the transformer secondary into a supply of only one polarity, that is to say, the vibrator is self-rectifying. The two types are known as non-synchronous and synchronous respectively.

Compared to the dynamotor, the vibrator supply shows a higher order of efficiency, usually between 60–70 per cent, since the idling current under no load conditions is substantially smaller. However, since the reed contacts are virtually a switch operating in the region of 100 times per second, contact wear reduces the life of a vibrator well below that of a dynamotor. Additionally, the vibrator is not as tolerant of overloads since these invariably give rise to arcing at the contacts which further reduces their life, and may in extreme cases result in the contacts becoming welded together.

Despite the foregoing observations a vibrator power supply will give good service when used within its current ratings and, in view of its higher efficiency, it is probably to be preferred to a dynamotor.

The output of a vibrator is prolific in harmonic interference since its waveshape approximates to a square wave. For this reason, the input and output circuits must be well filtered, and the whole unit contained within a fully screened and well earthed enclosure. The circuit of a typical vibrator supply is shown in Fig. 16.38.

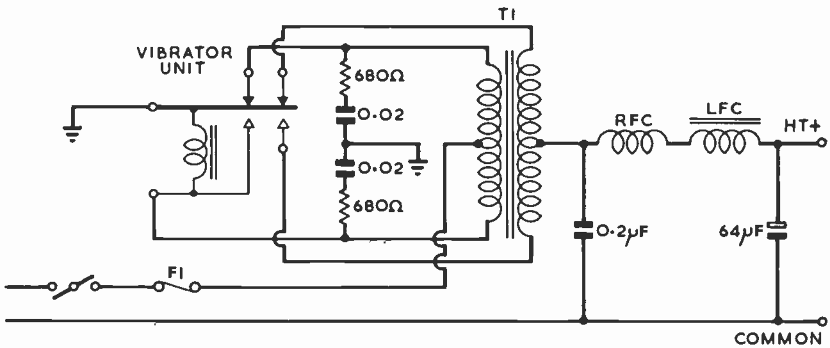


Fig. 16.38 Typical self-rectifying (synchronous) vibrator power supply.

Transistorized D.C. Converters

As the frequency of the supply to a transformer increases, so, depending on the nature of the core material, does the efficiency of the transformer increase. At frequencies of between 1000 c/s and 2500 c/s modern materials make possible transformers in which the core losses are negligible, and due to the increased magnetic efficiency, have a high coupling coefficient between primary and secondary windings. The net result is that a transformer having a certain rating for use at, say, 1000 c/s, will be considerably smaller than one of the same rating designed for 50 c/s operation.

For mobile operation, where the highest power conversion efficiency is always sought, the use of such transformers can drastically reduce power losses. With suitable circuits, the full load efficiency will not be less than 80 per cent, and may well be of the order of 85 per cent. Compared to a dynamotor this represents a saving of 30–35 per cent.

With a vibrator unit, some increase above mains frequency is possible, but due to mechanical considerations related to the reed and the manner in which it is energized, the frequency is usually of the order of 100 c/s. However, sub-

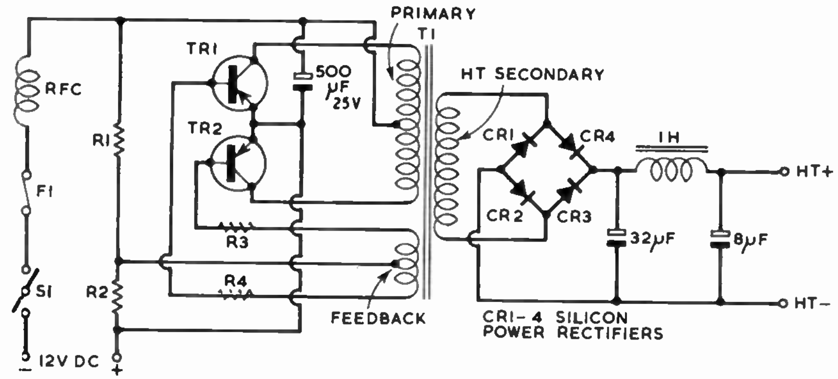


Fig. 16.39 Typical example of the circuit of a high efficiency transistorized d.c. converter operating at a frequency of about 1000 cps. With circuits of this type, efficiencies better than 80 per cent are to be expected.

stituting power switching transistors for the vibrator unit enables higher frequencies to be used with consequent simplification of smoothing.

Where transistors are employed, the frequency of the switching may be determined by arranging the primary of the transformer, in association with a feed-back winding, as an oscillatory circuit, and under these circumstances, the frequency can be fixed at that corresponding to highest transformer efficiency. Since there are no moving parts in a transistor, the potential reliability of converters using them is substantially higher than those which employ vibrator units. However, to realize the full reliability factor, certain precautions have to be taken, especially with regard to the ambient temperature and transients.

The circuit of a typical transistorized d.c. converter is shown in Fig. 16.39. The supply voltage is applied to the centre of the primary winding, the outside ends of which are connected to the collectors of TR1 and TR2 respectively. Switching within the transistors takes place between the collector and emitter connections and depends on the forward bias applied to the base; when the transistor is driven into saturation, the resistance between collector and emitter is effectively zero. The feedback winding feeds the switching currents to the bases of the transistors, and a small amount of forward bias is derived from the potentiometer chain R1, R2.

Since transistors TR1 and TR2 are connected in push-pull they conduct on alternate half cycles of the primary oscillation. When the base of TR1 is driven negative, the base of TR2 will be driven positive, and hence TR2 becomes non-conducting. During the next half cycle TR2 is driven negative, and hence conducts, while TR1 is cut-off.

The circuit as a whole needs to be designed with a number of points in mind. Firstly, the transistors selected must be able to switch the primary current without difficulty. Secondly, since the amount of feedback will decrease as the load on the transformer is increased, the feedback circuit must be designed so that when the transformer is supplying its rated load, the feedback level is sufficient to maintain oscillation. Thirdly, under lightly loaded conditions, the feedback voltage to the bases of the transistors will be at its highest, and the transistors must be able to withstand the reverse base-emitter voltage which will appear across this junction during the cut-off condition. Fourthly, the transistors must be rated to withstand any peak transient voltage occurring during the switching process, as any such spikes will cause failure of the transistors if they are not adequately rated. A good rule of thumb is to employ transistors in which the collector voltage is specified as at least three times the maximum supply voltage. In the case of mobile operation where the line voltage may rise to 14·5 volts in a 12 volt nominal system when the generator is charging fully, this implies transistors with a Vce of at least 45 volts.

From the foregoing it will have been noted that the feedback conditions are equated for full load, and from this it follows that under reduced loading conditions, the efficiency will be lower since TR1 and TR2 are being switched for a greater current than that which the primary will actually need to handle to meet the demand. If overall efficiency is of paramount importance, it is sometimes the practice to reduce the drive to the transistors under low loading conditions by switching in series resistors in the positions R3/R4 so reducing the drive to the transistors and thus, their standing current. Such switching is usually controlled by a relay linked to the transmit/receive switch so that the resistors are in circuit during receiving periods when the current demand is usually less.

For any transformer, but especially one which has a higher rating than the load to be supplied, resistors R3 and R4 may

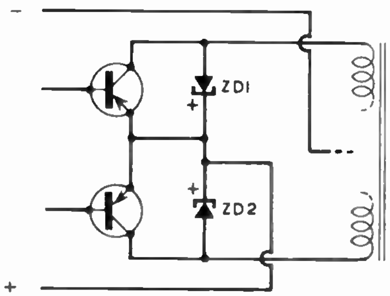


Fig. 16.40 Use of zener diodes across the collector-emitter junction to limit transient voltage spikes. The diodes should be rated for a breakdown voltage slightly less than the Vce rating of the transistor.

be determined by trial and error. With the supply operating, various values are connected in positions R3 and R4 with the object of fitting the largest possible resistor which does not reduce the power output to the load. In general terms, these resistors will have values below 20 ohms, and may be as low as 3 ohms.

It is sometimes thought that transistorized d.c. converters do not cause interference, this conclusion being based on the premise that since there are no moving parts, and no contacts opening and closing and arcing in the process, there is nothing to be suppressed. This is a fallacy. While it is true that there is no interference due to contact arcing, the waveform within the generator is basically a square wave. Thus harmonics are present and these can produce a buzz similar

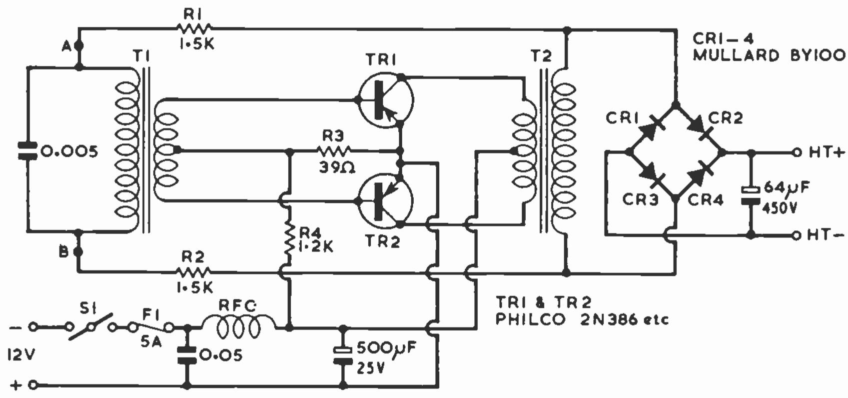


Fig. 16.41 Transistor d.c. converter employing two similar standard mains transformers and suitable for loads of up to 40 watts. The transistors must be mounted on heat sinks of 9 square inches minimum. Transformers T1 and T2, Radiospares Hygrade filament transformers. The two 6·3V windings on each transformer are connected in series to produce a 12·6V winding. Check by applying mains voltage to the transformers, that the 6·3V windings have been connected as series aiding. See text for adjustments to R1/R2 in relation to output load. Power transistor types are not so critical so long as Vce is not less than 48V.

to that of a vibrator unit. For this reason r.f. suppression is essential. Furthermore, there is a strong magnetic field around the transformer, and audio circuits should be positioned well clear of the converter if pick-up of the converter frequency is to be avoided.

Standard vibrator transformers may be used in conjunction with transistors to produce a transistorized d.c. converter, and while the efficiency will only be slightly higher than that obtained when the transformer is operated with a vibrator unit, suppression is easier due to the absence of arcing contacts. A typical design is shown in Fig. 16.30(a). Here the feedback, and hence the efficiency is determined by the value of the series resistors between the outer ends of the primary and the bases of the transistors, and to achieve the optimum conditions for a particular transformer may mean changes in their value. The process of determining these resistors is the same as that previously described; to start with a 100 ohms resistor is connected in each arm in this case, power and load applied. The resistors are then changed equally in value, in small increments, until the largest value is found with which the desired power output is secured.

Temperature is all important to transistors, and those used in d.c. converters must be mounted on an adequate heat sink, and operated in such a manner that under full load conditions their cases feel only warm to the touch. If there is any sign of overheating, either the load is too heavy for the type employed, or they are being overdriven. In the first case a larger transistor should be fitted, and in the second, the degree of base feedback should be investigated with the object of reducing it to a lower level.

Sometimes a transistor d.c. converter may operate happily for months, and then, for no accountable reason, fail. Such failures are invariably due to transients on the 12 volt vehicle system created by some other electrical device. To absorb such spikes, it is essential to include a high value capacitor across the supply line to the d.c. converter, as shown on the diagram Fig. 16.39. Another line of approach is to clamp the collector-emitter potential with a zener diode having a breakdown voltage somewhat less than that of the Vce rating of the transistors used. While this is more costly than the use of capacitors, it is unbeatable for efficiency in respect of the job it has to do, and certainly far less costly than having to replace transistors. The arrangement is shown in **Fig. 16.40.** It should be noted that a zener diode must be placed across the collector-emitter connections of each transistor.

For constructors having suitable transformers on hand, one way of reducing the cost of a transistorized d.c. converter is shown in **Fig. 16.41.** This employs two similar 12 volt centre tapped mains transformers, one to provide the voltage step up, and the other for matching to the bases of the switching transistors and to provide the feedback.

If the unit fails to oscillate on switching on, the primary winding of T1, marked A.B., should have its connections reversed to correct the phase of the feedback.

This converter will supply up to 40 watts at voltages ranging between 250 volts and 300 volts depending on load. At 30 watts output, the terminal voltage will be approximately 280 volts. For maximum efficiency, the values of R1 and R2 should be varied, and those fitted have the largest value consistent with maintaining the desired output and reliable starting under load. When correctly adjusted, conversion efficiency of the order of at least 65 per cent is to be expected.

One effect which is sometimes noted with transistorized d.c. converters is that they show a reluctance to start under load, but once running, perform in a satisfactory manner. This is invariably due to the feedback being set at a critical level. If there is no way of increasing the level of the feedback in a convenient manner, sequential switching of the supply and the load will usually provide a solution to erratic starting. This entails employing a double pole relay, one set of contacts of which are in series with the supply, and the other set in series with the h.t. output to the load. The contacts in series with the supply are carefully adjusted so that they close before the h.t. contacts. With this arrangement even the most reluctant circuit should start without difficulty.

MOBILE AERIAL INSTALLATIONS

The success of any transmitter not only depends on the power which it has available, but the ability of the aerial system to radiate this power efficiently. A station with a 50 watt capability may in fact radiate a weaker signal than that achieved by a 10 watt capability station simply due to an inefficient aerial.

Aerial fixing must be mechanically strong enough not to be damaged when the vehicle is travelling at high speeds, or succumb to a blow caused by a low hanging branch, or entering the garage before dismantling the aerial. Further, if the vehicle is not to lose some of its resale value, careful thought must be given to any ideas of making fixing holes in the bodywork. Whatever fixing system is employed, it should be so arranged that the main support bracket can be removed, and the aerial system easily demounted.

While on the low frequency bands aerial length does have some effect, the difference between a 16 ft. loaded whip, and an 8 ft. loaded whip will be so marginal that it is hardly worth being bothered with the mechanical complications of the longer aerial.

Low Frequency Vertical Aerials

When operating on frequencies of 28 Mc/s and higher, it is practical to use a quarter-wave aerial; on the lower frequency bands, such aerials are impractical. However, by the use of a suitable loading coil, a short aerial can be made to behave like a quarter-wave. Such loading coils may be positioned either at the base of the whip section, in the middle of the whip, or at the top. Top loading is rarely employed, due to mechanical difficulties, but centre and base loading are commonly used.

Every aerial system exhibits a characteristic radiation resistance. Briefly radiation resistance may be defined as that value of resistance which, if substituted for the aerial at a current maximum would dissipate the same amount of energy as heat as the aerial radiates into space. If power is

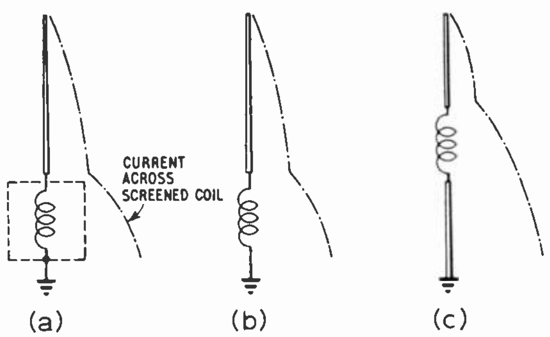


Fig. 16.42 Current distribution along loaded whips. In (a) the coil is fully screened from the whip section. In (b) the whip is base loaded, and in (c) the whip is centre loaded.

to be transferred from the transmitter to the aerial system with maximum efficiency, then the transmitter feed system must match the radiation resistance of the aerial. For normal values of feed point impedance this may be accomplished without difficulty, but where the value is very low, as may be the case with a loaded whip, a substantial proportion of the transmitter power may be lost in the matching system unless care is taken. Generally, the higher the radiation resistance of an aerial, the easier it is to match, and the higher will be its efficiency.

The position of the loading coil of a low frequency vertical aerial has a pronounced effect on the radiation resistance. Where the coil is placed at the base of the aerial, and is fully screened as in **Fig. 16.42(a)**, the radiation resistance can be of the order of 0·2 ohms. If the coil is now made a component part of the vertical system and allowed to radiate, the radiation resistance of the system rises to about 2 ohms. This is illustrated in **Fig. 16.42(b)**. If the coil is now positioned in the centre of the whip as in **Fig. 16.42(c)**, both the lower section of the whip and the coil will have current flowing in

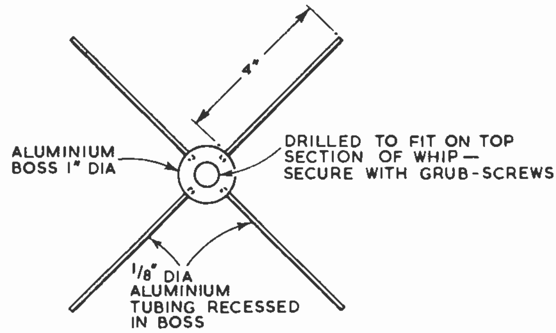


Fig. 16.43 A capacity hat suitable for use with vertical whip aerials.

them, and this again raises the radiation resistance, this time to about 4 ohms. These figures relate to an aerial operating on 3·8 Mc/s.

As is to be expected, the position of the loading coil has a pronounced effect on the current distribution in the aerial system as a whole, and the dotted lines running down the various arrangements shown in Fig. 16.42 represent the current distribution. Since the field radiation is proportional to the current flowing, and the length of the aerial carrying this current, then it will be apparent that the best radiator will be the one which has the highest current length product. From Fig. 16.42 this will be seen to be the centre loaded aerial.

While the centre loaded aerial may be the best radiator, on the lower frequency bands, particularly 1·8 Mc/s and 3·5 Mc/s, mechanical considerations may have to influence the loading coil position. On these two bands it is more usual to employ a base loaded whip. However, as will be seen later it is possible to introduce some compensation by attention to the form of the loading coil.

Loading Coils

The statement that centre loaded whips are the best radiators is not strictly correct since, for a given whip length, and depending on frequency, there is an optimum position for the loading coil. The radiation resistance of an aerial, but more important, its capacitance, are not things confined to one point on the whip, but rather are spread along its entire length. When the coil is placed at the base of the whip it is resonated by the total capacitance of the whip, but as it is raised, the capacitance of the section of whip decreases. As the capacitance decreases, so the loading coil size has to be increased to restore resonance. Increasing the size of the loading coil also increases the losses in the coil, and eventually as the coil is further raised the law of diminishing returns takes over and the current-length product falls.

This effect may be offset to some extent by deliberately introducing capacity into the top section of the whip above the coil. Such a method is commonly described as a capacity hat, and is shown in **Fig. 16.43**. By introducing capacity in this manner, the size of the loading coil may be reduced, and its losses also. However this system is not without its disadvantages. As capacity is added above the coil, so the aerial becomes increasingly frequency selective. As an example, with a centre loaded coil on 3·5 Mc/s it may be possible to vary the transmitter frequency over about 25 kc/s before it becomes essential to re-adjust the loading. If this coil is replaced by another which is tuned by a capacity hat, a frequency shift in excess of 10 kc/s may well entail retuning the aerial.

An aerial which is short of a quarter-wave at the operating frequency exhibits a capacitive or negative reactance to the tuning device, and the shorter the aerial is in terms of a quarter wave, the higher will be the value of this reactance. Take for example a standard 8 ft. whip without any loading coil, the resonant frequency of which is about 29·5 Mc/s. At this frequency the feed impedance will be 35 ohms, and since it is resonant, there will be no negative reactance. At a frequency of 21 Mc/s, the aerial is electrically short, the radiation resistance drops to about 11 ohms, and the negative reactance effect appears, in this case showing a value of −250 ohms. At 3·9 Mc/s the radiation resistance has dropped to a bare 0·2 ohms, and the negative reactance rises to approximately −1650 ohms.

From the foregoing it will be appreciated that the loading coil is literally the heart of the loaded whip for it has to cancel out the negative reactance by adding to the aerial an equal positive reactance, so leaving only the radiation resistance as the load for the aerial tuner.

Loading coils are never perfect, and in addition having reactance, must, by virtue of their construction, also exhibit resistance. The goodness of any coil is expressed by the term Q, and this is given by the ratio of the reactance to the r.f. resistance. Consider a loading coil for the 3·5 Mc/s band which has a reactance of 1400 ohms, and an r.f. resistance of 10 ohms. Such a coil would have a Q of 140.

Loaded Whip Power Distribution

The relationship between the radiation resistance of the whip and the r.f. resistance of the loading coil have a direct bearing on the manner in which the power applied is distributed across the whole of the loaded whip, sometimes called the coupling efficiency. The percentage of the power in the top section is given by:

$$\frac{Rr}{(Rr + Lr)} \times 100$$

where Rr is the radiation resistance of the whip, and Lr the r.f. resistance of the loading coil. The resulting figure subtracted from 100 gives the percentage of the power dissipated in the loading coil.

Using the coil previously detailed, assume that the aerial has a radiation resistance of 6 ohms, then the power in the section will be:

$$\frac{6}{6 + 10} \times \frac{100}{1} = 37{\cdot}5 \text{ per cent,}$$

leaving 62·5 per cent appearing in the loading coil. If however the radiation resistance is 3 ohms, which is a likely figure for base loading, then the power in the top section will be:

$$\frac{3}{3 + 10} \times \frac{100}{1} = 23 \text{ per cent,}$$

so leaving 77 per cent in the coil.

Although the loading coil does radiate, by far the greater majority of the power which appears across it is dissipated in the form of heat, and as such may be considered as wasted power.

Key Requirements

It is now possible to specify the requirements which have to be met to secure maximum efficiency in a loaded whip. Firstly the radiation resistance must be made as high as possible, and as the frequency decreases this implies moving the loading coil up the length of the whip and away from the base. Since as the loading coil is positioned further up the whip the top capacity decreases, moving the coil in this manner will also require that its size is increased unless capacity is deliberately added to the top section by fitting a capacity hat. On 1·8 Mc/s, the physical size of the loading coil may make it mechanically impossible to move the coil very far up the whip, and from the point of view of the appearance of the aerial installation, it may not be desirable to do so. In this event, as will be seen later, it is possible to reach a compromise by forming the coil in such a manner that it radiates a fair degree of the energy appearing across it.

The loading coil must have the highest possible Q, or put another way, the resistance and losses of the coil must be as low as possible. As the Q of the coil is increased for a given frequency, so the losses in it will decrease, and the strength of the radiated signal will increase.

Loading Coil Construction

The success or failure of an h.f. mobile aerial hinges almost completely on the amount of care and attention paid to the construction and adjustment of the loading coil.

Ideally, and to reduce dielectric losses, the loading coil should be self-supporting, but from the point of view of achieving a high degree of mechanical strength, this is rarely possible except perhaps on the 14 Mc/s and 21 Mc/s bands. Where such a course is adopted, the strength of the actual whip must be maintained through the length of the coil winding. The most satisfactory method of achieving this is shown in **Fig. 16.44** and consists of inserting an insulated rod into the bore of the whip top and bottom of the distance

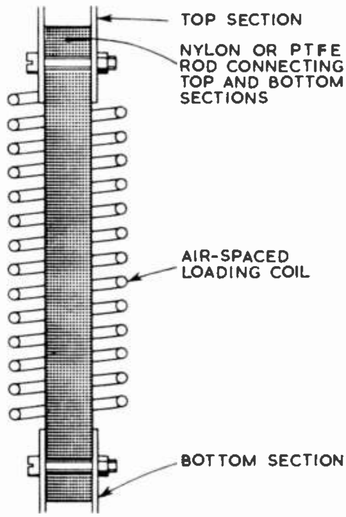


Fig. 16.44 Method of bracing and connecting the upper and lower sections of a whip when air spaced loading coils are employed.

occupied by the loading coil, and running this rod through the centre of the coil. As the rod runs through the centre of the coil, it must have a low dielectric loss. P.T.F.E. or nylon are the most suitable materials.

It is essential to protect any loading coil from the effects of moisture, and where the coil is a self-supporting unit, this is best accomplished by enclosing it in a shield constructed from polythene and which is fully sealed. The shield should clear the turns of the coil by about ½ in. Freely available polythene bottles make a good basis for such shields.

Where formers are employed for the loading coil, ideal materials are nylon, because of its high mechanical strength, or p.t.f.e. Paxolin can be used on 1·8 Mc/s always provided that its relatively high dielectric losses are acceptable. Using the first two mentioned materials, coils with a Q in the region of 400 are practical.

Wire gauge will be a compromise on the lower frequency bands between the final size of the coil and the acceptable r.f. resistance. On 1·8 Mc/s the smallest wire gauge contemplated should be 18 s.w.g., on 3·5 Mc/s 16 s.w.g., and 7 Mc/s and higher 12 s.w.g. Whenever possible the wire should be *Formvar* insulated. This is a tough plastic coating impervious to moisture, and unlike enamel, will not chip or flake.

After the coil has been finally adjusted to resonance, it must be thoroughly doped with a suitable compound to prevent moisture getting at the turns. Polyurethene varnish seems to be the most suitable material currently available, and by using a coat on coat process, a tough and completely waterproof layer can be built up over the coil. This coating must include the coil terminations, and special attention should be paid to the top of the loading coil which will receive all the water running down the whip when the vehicle is stationary.

Finally, there is the question of the form factor of the loading coil. For optimum Q the length of the coil should be about twice its diameter. If there is no objection to a fairly large diameter coil, this requirement can just about be met on the 3·5 Mc/s band. On 1·8 Mc/s however such proportions are impractical.

Loading Coils in Practice for 1·8 Mc/s and 3·5 Mc/s

The 1·8 Mc/s and 3·5 Mc/s bands pose special problems in relation to loading coils for use with an 8 ft. whip, for in nearly all cases the size of a coil designed for optimum Q would not only be difficult to construct, but also look ridiculous when fitted to a vehicle.

In a number of experiments conducted with loading coils for these bands, the conclusion was reached that if the form factor requirement was disregarded, and instead, steps taken to produce a coil which exhibited a high degree of radiation, then the overall radiation efficiency was improved in quite a remarkable manner. These conclusions were not based solely on the readings of an adjacent field strength meter, but were also confirmed by stations at varying distances. In many respects the loading coil appeared to have the attributes of a helical whip, an aerial system at one time popular on the higher frequency bands.

As the radiation from the coil increases, as has been indicated earlier, the radiation resistance of the whole system increases, and this in turn makes matching to the transmitter easier to accomplish, so reducing the losses which are inevitable where the feed impedance is very small. Quoting

TABLE 16.8

Starting details of loading coils for various frequency bands and positions on the whip. Overall length of whip is 8 feet.

Band	Position	Dia. in.	Turns	S.W.G.	Length in.	Notes
1·8 Mc/s	Bottom	3	140	18	7	Q 300
1·8 Mc/s	Bottom*	1½	260	16	17	Max radiation
1·8 Mc/s	Centre	1½	475	22	27	Max radiation
3·5 Mc/s	Bottom	2½	75	14	6	Q 300
3·5 Mc/s	Bottom	2½	55	18	2½	Q 350
3·5 Mc/s	Bottom*	1½	155	16	16½	Max radiation
3·5 Mc/s	Centre	2½	105	16	10½	Q 300
14 Mc/s	Bottom	2	10	14	1½	Q 380
14 Mc/s	Centre	2	16	14	2	Q 350
21 Mc/s	Bottom	1¾	6	12	2	Q 300
21 Mc/s	Centre	2	8	12	2	Q 280

* Coil mounted 12 in. above base of aerial.

from the results of experiments with a whip tuned to 3·750 Mc/s, the measured impedance was 16 ohms, and the calculated overall efficiency of the aerial 60 per cent. This used a coil 1½ in. dia. and approximately 16½ in. long wound with 155 turns of 16 s.w.g., and had a measured Q of 200. The coil was positioned 12 in. up from the feed point. Compared to a coil wound to the optimum form factor with a measured Q of 400, the long loading coil produced almost twice the field strength.

Although the effects of centre loading with such coils was not evaluated, it is evident from the base loading experiments that this line of approach should be fruitful. It seems likely that the performance of a particular commercial centre loaded whip, and in which the loading coil has a long length compared to its diameter, arises from the radiation of the loading coil.

Table 16.8 gives approximate details of loading coils for the lower frequency bands. The details must be approximate since every aerial system will need to be tuned in relation to the vehicle with which it is to be employed, for it is impossible to estimate such factors as earth-loss resistance and car capacitance.

Tuning Loaded Whips

Just as spending time on the construction of the loading coil will be worthwhile, so the care taken in tuning the system will pay dividends.

Tuning to frequency will be made a great deal easier if the section of the whip above the loading coil is constructed so that its length can be varied. If this cannot be arranged, then a variable size capacity hat will almost certainly be required to achieve precise resonance on any particular frequency.

The lower end of the whip should be earthed to the car body through a two turn link, the length of wire between the whip and the link being made as short as possible. A grid dip oscillator is then offered up to the link and tuned to find the resonant frequency. The coupling between the g.d.o. and the link should be as small as possible otherwise some inexplicable dips may be found. During this operation, the top section of the whip should be extended to within about 6 in. of its full length.

If the frequency is found to be very low, turns should be removed from the loading coil. If on the other hand the frequency is found to be high, this indicates inadequate turns. Where Table 16.8 is used, the frequency will be low in each case, for the specifications are on the basis that it is easier to remove turns rather than have to rewind the coil.

Accuracy in the g.d.o. is essential, and if it is an unfamiliar instrument borrowed for the purpose of setting up the whip, prior to use, its calibration should be checked against a receiver of known accuracy, and over the frequency range of interest. A mental note can then be made of any calibration error.

As the resonant frequency of the whip approaches the low side of the desired frequency, if it is possible, the frequency of the g.d.o. should be continuously monitored on a receiver. Normally, the receiving equipment employed in the vehicle may be used. When within 10 kc/s of the low frequency side of the desired resonant frequency, adjustment of the length of the top section of the whip should permit the whip to be precisely resonated to the desired frequency. To increase the frequency of resonance, the top section of the whip is reduced in length.

Loading Tuned Whips

Since the feed impedance of a loaded whip will be substantially lower than the normal transmitter output impedance of 75 ohms, the base impedance of the aerial must be matched to the feed line impedance. Experience indicates that this is best accomplished by the use of a variable series loading coil as shown in **Fig. 16.45**. This should consist of a roller wheel making contact with the turns, which, by rotating the coil, travels along the winding. Units employing this principal are sometimes available on the surplus market. To construct such a device is usually hardly a practical proposition for the average radio amateur, and this being the case, a tapped coil will probably have to be employed.

If the coil is constructed from 16 s.w.g. tinned copper wire, the turns of which are spaced by one wire diameter, connecting lugs may be soldered into place every one-third of a turn. Closely adjacent taps are essential to achieve correct loading, and while the coil will not be very beautiful when it is completed, it will be efficient. Typically such a coil consists of 30 turns on a 1½ in. dia. former.

As the frequency of operation is changed from the resonant frequency of the whip, the whip should really be re-resonated by adjustment of the length of the top section. On 1·8 Mc/s a change in excess of ±10 kc/s may well reflect on the transmitter loading, and a change of 20 kc/s produce a serious drop in performance. To some extent such changes may be taken up in the loading system. With the tapped coil, it is possible to incorporate a relay actuated from the driving position which changes the tap on the coil and so compensates for the change in frequency. This is shown in Fig. 16.45. If this course is adopted, the control switch is best marked with the actual resonant frequencies as a reminder of the range to which the aerial is tuned.

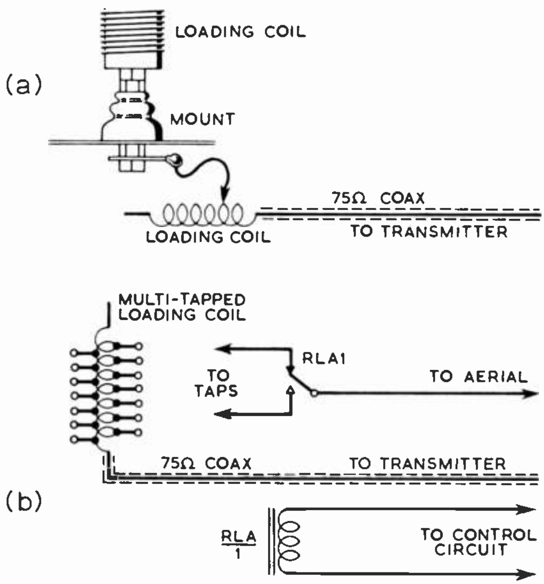


Fig. 16.45 (a) shows the use of a series inductance to match the base of the aerial to the transmitter feed line. For limited coverage a tapped inductance may be used in conjunction with a relay to adjust the inductance as the frequency of operation is changed. This is shown in (b).

V.H.F. Aerials

On frequencies of 70 Mc/s and 144 Mc/s aerials become simpler since it is possible to fit a full quarter wave. Under ideal conditions, and with the aerial mounted in the centre of the roof, it will function as a ground plane, the feed impedance of which will be of the order of 35 ohms. However, and understandably, there is usually considerable objection to cutting holes in the centre of the roof of a saloon car, and to having an aerial standing up in that position.

For all practical purposes, a quarter wave aerial may be mounted either on the apron surrounding the lid of the boot, or forward of the windscreen, without degrading the performance to any appreciable degree. If it is close to rising metalwork of the saloon, its polar diagram might be very distorted, and for this reason it should be positioned as far away from the rise of the saloon as possible. The polar diagram will still deviate substantially from a circle, but it will not have so many peaks and nulls in unexpected directions. A typical polar diagram of a 70 Mc/s aerial installation is shown in **Fig. 16.46.**

Where possible, 35 ohm co-axial cable should be used to feed the whip as this will give an almost perfect match. As an alternative the length of the aerial may be adjusted to present an impedance of 75 ohms at its base. As an aerial is made longer than a quarter wave, the base impedance rises, and at some point can be made to look like 75 ohms.

To arrive at this situation with any accuracy two instruments are required. Firstly an s.w.r. bridge for a line impedance of 75 ohms, and a field strength meter covering the band of interest. The s.w.r. bridge is connected in the feed line as near to the base of the aerial as possible, and the field strength meter placed at a convenient distance. Do not stand the field strength meter on the vehicle itself, as r.f. currents in the car will result in misleading indications. Set the aerial to a calculated quarter wavelength: 19 in. for 144 Mc/s or 41 in. for 70 Mc/s. Switch the s.w.r. bridge to *Forward*. Apply power to the transmitter, and adjust the transmitter output circuit for maximum forward reading.

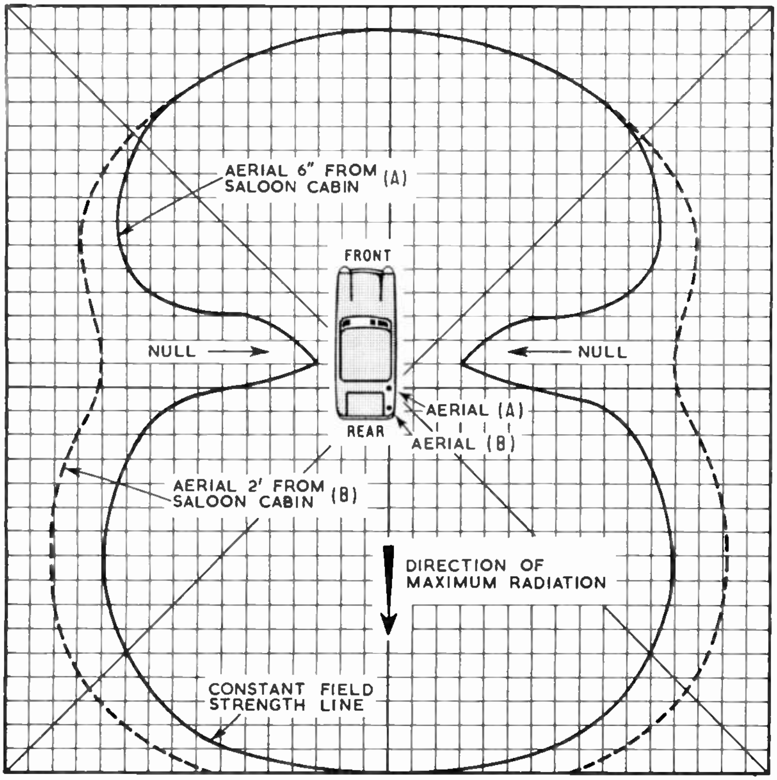


Fig. 16.46 Measured polar diagram of a 70 Mc/s ¼ wave vertical aerial mounted on the rear of a Triumph Herald. Maximum radiation takes place towards the rear, but that to the front is not greatly lower. Notice deep nulls off the sides of the vehicle when aerial is within 6 in. of saloon cabin, Moving further back so that it is 2 feet from saloon cabin substantially reduces side nulls.

Slowly *increase* the length of the whip to either (*a*) increase the forward reading on the s.w.r. bridge or (*b*) decrease the reflected indication. Once an apparent optimum point has been found, again adjust the transmitter output circuit for maximum forward indication of the bridge. Now carefully adjust the aerial length for minimum *reflected* power as indicated by the s.w.r. bridge.

When correctly adjusted, minimum reflected power will coincide with maximum forward power and maximum indication on the field strength meter. Using this method, the s.w.r. on the feed line can be expected to be better than 1·25:1.

During the course of these adjustments it will be found that opening and closing doors, and the lid of the boot, will have a pronounced effect on the readings, as will walking near to the aerial or in the proximity of the vehicle. This is a practical demonstration of how surrounding objects affect the operation of a mobile aerial. In all cases the readings should be made from inside the car and with all the openings closed, and if possible, all adjustments conducted with the car in an open space.

As an alternative to a vertical quarter wave, on 144 Mc/s a halo aerial may be employed. Being virtually a horizontal dipole, this aerial may show a performance superior to that of the quarter wave whip when working to base stations using horizontal polarization. Standard halo aerials tend to be uncomfortably large even on 144 Mc/s, and this led to the design of a miniaturized capacity loaded halo, the construction of which is illustrated in **Figs. 16.47, 16.48 and 16.49.**

The most important part of the design is the tuning capacitor, and this must be constructed with care. To allow sufficient threads to be cut in the wall of the outer adjusting sleeve, this is tapped at 6B.A. It should be noted that the inner of the two tube sections making up the capacitor is secured to one end of the circular element by a screw which force fits into the bore of the tube. The polythene or p.t.f.e. lining of this inner tube is fitted to the other end of the circular element by cutting a thread on its outside diameter and force screwing the lining over this thread. To assemble, the outer sleeve of the capacitor is slipped over the inner sleeve, the p.t.f.e. end of the element sprung away from the sleeve end, and then inserted into the bore of the capacitor inner sleeve.

The gamma match is of the same dimensions as would be used on a full sized version, that is approximately 4½ in. long. In setting the system up the outer sleeve of the capacitor is used to bring the aerial to resonance, and then, in conjunction with an s.w.r. bridge, the gamma match is set for minimum s.w.r.

Contrary to what might be expected,

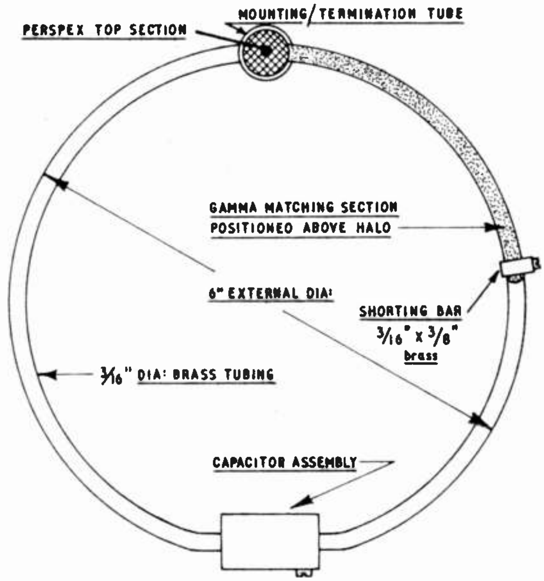


Fig. 16.47. Looking down on the top of the completed Mini Halo.

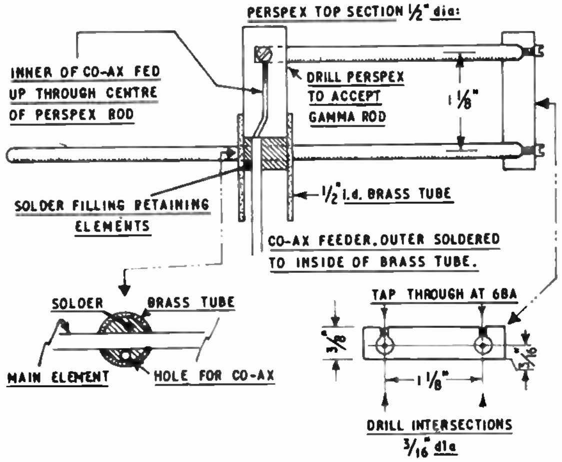


Fig. 16.48. Side elevation of the completed halo, with drilling details of the gamma match support and perspex mounting rod.

the mini-halo is not particularly frequency conscious, and tuning 2m does not show any decrease in signal strength from one end of the band to the other.

MONITORING FIELD STRENGTH

Under mobile conditions it is highly desirable to be able to monitor the field strength of the signal radiated by the transmitting aerial. Using a field strength meter within the saloon of a car can give completely misleading indications due to the r.f. currents circulating within the car body. Ideally a small external aerial is required, but the idea of fitting yet another aerial is generally unacceptable.

A neat solution to this problem is to use a wing mirror as the field strength meter aerial by insulating such a mirror from the frame of the car. The method is illustrated in **Fig. 16.50.** For best results the wing mirror metal parts should be made from brass as the base metal. The lead from the solder tag on the underside of the mirror to the F/S meter is a coaxial lead, the screening of which is either earthed at the indicator end in the case of v.h.f. operation, or connected to the inner for the h.f. bands.

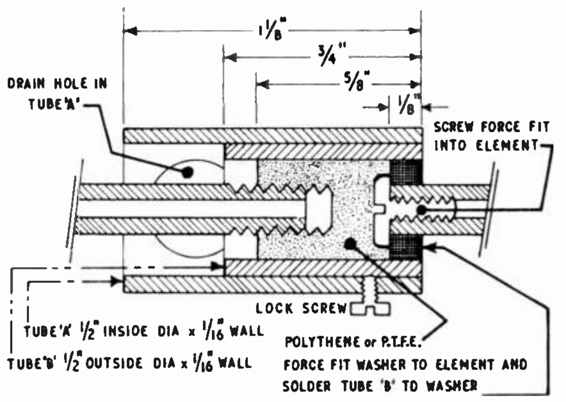


Fig. 16.49. Cut-away view of the capacitor.

BATTERY

The modern storage battery is inherently reliable but because of its very reliability, it frequently fails to give good service simply because it is taken for granted.

Just as a motor vehicle needs to be serviced regularly if it is to be maintained in good condition, so the accumulator needs a little attention now and then, but, compared to the vehicle as a whole, its requirements are minor. The most important of its needs is distilled watter.

Distilled water must always be used for topping-up the electrolyte, never tap water, nor boiled water. Both of these contain traces of mineral elements, some of which can react with the coating on the plates to produce either a semi-insulating compound, or a compound which is shed during the charging process.

Prolonged charging at an excessive rate is as damaging to a storage battery as is allowing the level of the electrolyte to fall below the level of the plates. For this reason, adjustments should never be made to the voltage regulator with a view to increasing the charging current without first determining that the higher rate is within the capabilities of the battery. It is usually permissible to increase the charging rate to some degree to compensate for the extra loading imposed by mobile radio equipment, but such adjustments should always be made by a competent automobile electrical engineer. Alternatively, if the extra loading is heavy, consideration might be given to the provision of an auxiliary battery in the manner described earlier on page 16.18.

Where radio equipment is added to the circuit loading and no compensating output from the generator arranged, it is essential that provision is made to trickle charge the battery overnight after the equipment has been used.

The battery terminals should be kept clean and coated with grease, and the connecting lugs well tightened.

Should it be necessary to remove and store the battery, it should first be *fully charged* before being placed into store. It should be given a weekly boost charge to keep it in the fully charged condition.

The off-load voltage of a battery may not give a correct indication of its condition. As the state of charge of a battery varies so does the specific gravity of the electrolyte. For this reason, the most reliable indication of the state of a battery is given by a hydrometer, and this is an instrument which every mobile operator should possess. Normal s.g. levels lay between 1·150 and 1·250 and may vary from battery to battery of the same type.

MOBILE SAFETY

While it is to the credit of mobile operators that no accidents have been attributed to the operation of mobile radio equipment, in the event of such an accident there could be no valid defence.

One item which could cause such an accident is the fist microphone and its trailing lead. Anyone who has foolishly tried to negotiate a sharp turn, or a roundabout while clinging to a microphone with a press-to-talk switch will readily verify that the hand held microphones can in these circumstances be particularly dangerous.

One microphone arrangement which goes a long way to solving this problem is that used in some radio-taxis and in which the microphone is fitted to a length of flexible tubing. Even this is not perfect for while it leaves both hands free for control of the vehicle, the head has to be maintained in

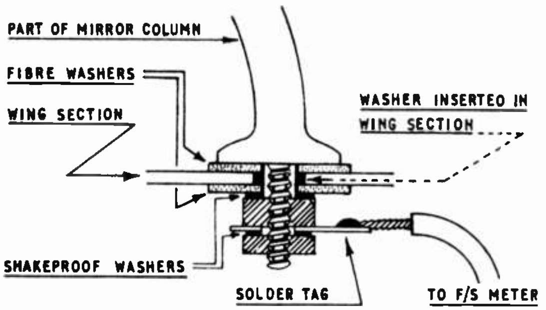


Fig. 16.50. Insulating a wing mirror to allow it to be used as the aerial for a field strength meter.

almost a fixed position for close speaking, resulting in a restricted field of vision for the driver.

A solution to this is to employ a halter as shown in **Fig. 16.51.** This fits over each shoulder and runs round the back of the neck of the wearer, and incorporates a boom upon which is mounted the microphone, and a section to carry the microphone lead clear of the user. Any of the usual microphone inserts may be fitted, and the lead run through the bore of the copper tube. If it is desired to incorporate transmit/receive switching with this halter, this can be accomplished by fitting a small box containing such a switch to the end of the halter where the microphone lead leaves the tube.

In use, as the driver/operator turns, so the microphone on the boom turns also, and over a very wide range of movements, the microphone will remain in close proximity to the mouth. Leaving both hands free, there is no interference with the process of driving, nor feeling of being encumbered.

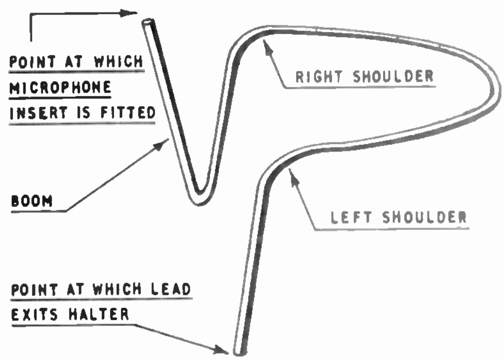


Fig. 16.51. General view of the combined halter and microphone mounting boom.

CHAPTER 17

POWER SUPPLIES

Almost all electronic equipment in which valves or semiconductors are used requires a source of power which is substantially free from ripple often described as pure direct current. Such supplies are generally derived from the a.c. supply mains by means of a power supply unit which may be simple or complex depending upon the particular requirements.

Where supplies are required for valve operated equipment, the power supply unit will be called upon to supply heater, grid bias and one or more high anode voltages.

The cathodes of most modern valves used in receiving and low power transmitting stages are indirectly heated and therefore the heaters may be supplied with either a.c. or d.c.; exceptions to this are low level audio amplifier stages where it may be essential to use d.c. for the heater power to ensure freedom from hum. This precaution is usually necessary only for the first amplifier stage in a high gain audio amplifier which is fed from low sensitivity microphones or similar devices.

In some types of valves used as r.f. amplifiers or modulators, the filaments are directly heated and these, because they are operating at high signal levels, can be supplied with a.c. without introducing hum into the output signal. It is desirable but not essential to use a centre tapped transformer with the centre tap earthed.

The various voltages commonly required may be obtained from a transformer, rectifier and filter system. Separate transformers may be required to supply the heater and grid bias supplies. High voltages, generally in excess of 500V, are usually obtained from a transformer which has low voltage windings for the rectifier valve heaters or filaments. Control and protection is simpler if the high voltage transformer does not supply any voltage other than that for a signal lamp.

Semiconductor equipment requires a low voltage, high current power unit, often having an output current of several amperes. Positive or negative earthing, depending on the type of device being used, may be required and must be borne in mind when constructing the power supply unit. A transformer, rectifier and filter system will be needed, but it is important to keep the internal resistance and impedance as low as possible.

Another type of power supply unit provides a constant voltage output which is virtually independent of the load. Shunt or series stabilizer valves maintain the output voltage constant within close limits set by the design.

Suitable circuits for all the common types of power supply units are described in detail later in this chapter.

Means are required for controlling the various voltages either by direct switching or remotely by means of relays or contacts. These may be associated with pilot or signal lamps showing which part of the equipment is alive. In addition suitable fuses or overload cutouts are necessary to protect the supply unit in event of overload or component breakdown in the equipment supplied from the power unit.

The actual overload and control arrangement will depend on individual requirements, but it is essential to ensure that at all times the equipment is safe to operate, and that a ready means is provided to remove all voltages rapidly in an emergency. In this connection, attention is drawn to the Amateur Radio Safety Recommendations in Chapter 20.

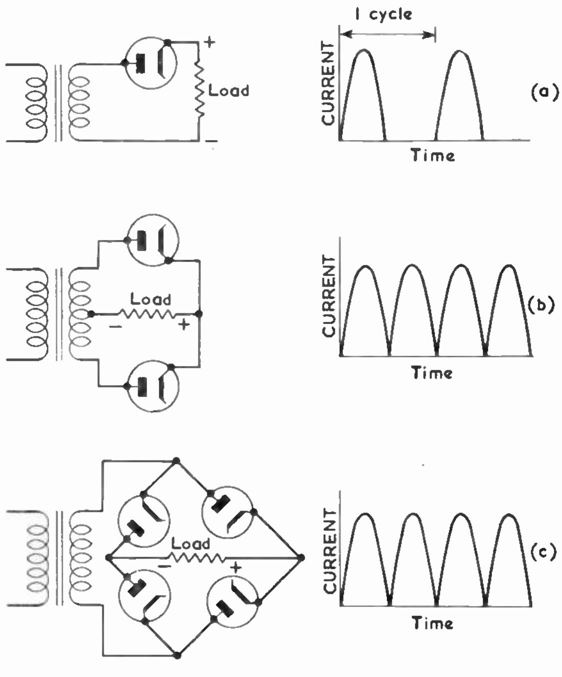
RECTIFIER CIRCUITS

A rectifier in one of its several forms is necessary when a.c. is to be converted to d.c. The rectifier functions by virtue of its property of unidirectional conduction, i.e., it passes current only when the applied voltage has the correct polarity to bias it to a low resistance condition. When the polarity of the input voltage is reversed the rectifier exhibits a very high resistance, approaching infinity in the case of a thermionic rectifier, and current cannot flow. At the present time both valve and semiconductor rectifiers are used in amateur practice, each with particular advantages in its own field.

Fig. 17.1 shows three types of rectifier circuit which cover most of the applications in amateur equipment, together with curves indicating the shape of the current wave delivered by the rectifier in a resistive load. The output current is seen to be unidirectional but pulsating in character and may be shown to consist of a d.c. component plus an a.c. component. The a.c. component is composed of the fundamental and harmonics of the supply frequency, the fundamental being predominant in the half-wave circuit and the second harmonic in the full-wave circuit. This a.c. component is termed *ripple* and may be removed or attenuated to any desired degree by the filter which follows the rectifier.

The half-wave circuit (**Fig. 17.1(a)**) is best suited to low current applications such as grid bias supplies and the e.h.t. supply for cathode ray tubes. When used to supply current of the order of tens of milliamperes and higher, the half-wave circuit has disadvantages in that the ripple content is high, necessitating a large filter; in addition the d.c. load current flows through the secondary winding of the transformer feeding the rectifier, and can saturate the core giving rise to low transformer efficiency. The regulation or variation of output voltage with load current is also poor.

The full-wave circuit (**Fig. 17.1(b)**), also known more correctly as the bi-phase half-wave circuit, is the most commonly used rectifier circuit suitable for voltages up to about 1500 volts and currents of the order of 1 ampere. A centre tapped winding is used to supply the rectifiers and must provide twice the a.c. voltage applied to each rectifier. The load is connected between the centre tap of the transformer and the strapped output terminals of the rectifiers. Each rectifier conducts during the time when the voltage applied



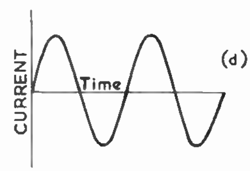


Fig. 17.1. Rectifier circuits showing the input and output current waveforms with resistive loads. (a) Half-wave. (b) Full-wave or bi-phase half-wave. (c) Bridge. (d) Input waveform.

to its input is positive with respect to the centre tap, i.e. on alternate half-cycles and contributes one half of the total load current. The d.c. component of the load current flows through each half of the transformer secondary in such a direction as to cancel the d.c. magnetization of the core, enabling a smaller and more efficient transformer to be used than is possible in a half-wave circuit providing the same output.

The bridge circuit (**Fig. 17.1(c)**) is preferred for higher voltage supplies since the peak inverse voltage across each rectifier is only half that of the bi-phase half-wave arrangement for the same d.c. voltage output. The peak inverse voltage, referred to as p.i.v. is the voltage appearing across the rectifier diode in the non-conducting condition. During each half-cycle of the input voltage, the rectifiers in opposite arms of the bridge are conducting and supply half the total load current. The d.c. magnetization of the transformer core is effectively cancelled, as the d.c. component of the load current flows through the h.t. secondary winding in opposite directions during each half-cycle.

The advantage of the bridge is seen in the simplification and reduction in cost of the transformer. Relative to the bi-phase half-wave circuit, the r.m.s. current rating is reduced by about one-third, and the turns on the winding feeding the rectifiers by half. In practice this results in a smaller transformer used more efficiently.

A disadvantage of the bridge circuit when valve rectifiers are used is that three separate heater windings are necessary since only two of the four valves have cathodes at the same potential. A separate rectifier heater transformer is usually required due to the high voltage insulation required between windings and core.

Table 17.1 gives the operating conditions for each type of rectifier circuit.

TABLE 17.1

Single Phase Rectifiers

	Half-wave	Full-wave or biphase half-wave	Full-wave bridge
Circuit	Fig. 1 (a)	Fig. 1 (b)	Fig. 1 (c)
Number of rectifiers	1	2	4
Current per rectifier			
—mean	1·0 I_{dc}	0·5 I_{dc}	0·5 I_{dc}
—peak (inductive load)	—	1·0 I_{dc}	1·0 I_{dc}
—peak (resistive load)	3·14 I_{dc}	1·57 I_{dc}	1·57 I_{dc}
—r.m.s.	1·57 I_{dc}	0·725 I_{dc}	1·11 I_{dc}
Peak inverse voltage across rectifier A	3·14 V_{dc}	3·14 V_{dc}	1·57 V_{dc}
Peak inverse voltage across rectifier B	2·83 V_{ac}	2·83 V_{ac}	1·41 V_{ac}
R.m.s. secondary voltage	2·26 V_{dc}	1·13 V_{dc} *	1·13 V_{dc}

V_{dc} = output voltage I_{dc} = output current V_{ac} = input a.c. voltage
* Each half of the h.t. winding.
A = choke input B = capacitor input

VOLTAGE MULTIPLIER CIRCUITS

The voltage multiplier is a form of rectifier circuit where the output voltage may be two or more times that obtained from the conventional half-wave or full-wave rectifier with capacitor input when supplied with the same a.c. input voltage. A combination of rectifiers and capacitors is used such that the capacitors are charged in parallel via the rectifiers and then discharged in series, the rectifiers acting as switches in addition to their normal function. The regulation of these circuits is poor as a result of this mode of operation, but the use of large value capacitors will assist in maintaining a reasonable regulation if the load varies appreciably.

Half-wave and full-wave voltage doubler circuits are illustrated in **Fig. 17.2 (a)** and **(b)**. The full-wave circuit has better regulation but has the disadvantage of not having a common input and output terminal. As in all full-wave circuits, the ripple component is at twice the supply frequency and is therefore easier to remove. At zero load, the output voltage is very nearly twice the peak value of the input a.c. voltage but falls rapidly to a value approximately twice the r.m.s. input voltage when appreciable current is drawn. The regulation may be improved by increasing the size of the capacitors, but care should be taken to ensure that the peak current ratings of the rectifiers are not exceeded. Semiconductor rectifiers are more convenient for use in these circuits although valve rectifiers may be used if desired.

Voltage tripler and quadrupler circuits are illustrated in **Fig. 17.2 (c)** and **(d)**. The tripler is a form of half-wave circuit where the output of a half-wave doubler is added to a normal half-wave rectifier. The quadrupler circuit is also known as the Cockcroft-Walton multiplier and if required, additional stages may be added to provide higher output voltages.

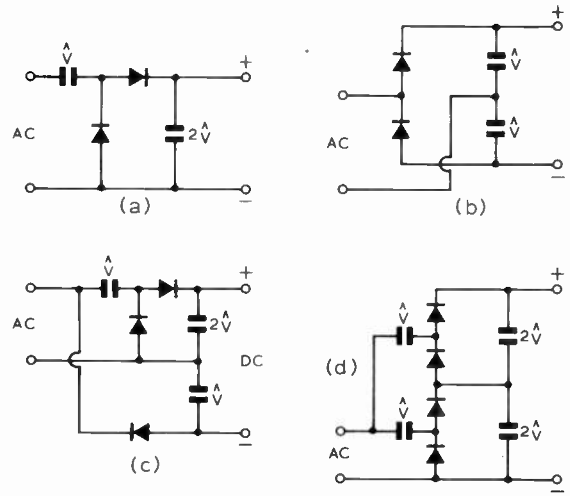


Fig. 17.2. Voltage multiplier circuits. (a) Half-wave voltage doubler. (b) Full-wave voltage doubler. (c) Voltage tripler. (d) Voltage quadrupler. $\hat{V}$ — the peak value of the a.c. input voltage. The working voltages of the capacitors should not be less than the values shown.

When used to provide appreciable currents (up to 50–60 mA) the capacitors shown in the circuits of **Fig 17.2 (a), (b)** and **(c)** should be at least 8 μF and preferably 16 μF, to provide reasonable regulation of the output voltage. Lower values (0·1 to 0·5 μF) will be satisfactory when the load is not more than 1–2 mA as in e.h.t. supplies. The polarity of the output voltage may be reversed in all the circuits shown by reversing the connections to each rectifier.

Types of Rectifier

Semiconductor and valve rectifiers may be used in the circuits described in the previous section. Semiconductor rectifiers have the advantage of low voltage drop and do not require heater power, but are more susceptible to breakdown when misused, particularly in respect of excessive operating temperature and peak inverse voltage. Generally speaking the semiconductor rectifier is preferred for use in low voltage high current supplies where the bridge connection is normally employed and in e.h.t. supplies, particularly when one of the various types of voltage multiplication circuits is employed. In passing it may be noted that the terminal of a semiconductor diode corresponding to the cathode of a thermionic diode is marked + or carries a red colour indication.

The valve rectifier is manufactured in a great variety of types and ratings suitable for almost any application in amateur equipment. They may be divided into two types—high vacuum and gas or mercury vapour filled, each having its particular field of usefulness.

The high vacuum rectifier is available in both half-wave and full-wave types, the latter consisting of two diodes in the same envelope with a common cathode and heater assembly. The indirectly heated type of low-power rectifier is preferred although the earlier directly heated types are still used. The indirectly heated type warms up at sensibly the same rate as the other indirectly heated valves in an equipment and so avoids the application of high no-load h.t. line voltages to the circuit. Some types of indirectly heated rectifiers will withstand heater-to-cathode voltages of up to 650 volts peak and are provided with 6·3 volt heaters, thus permitting the use of the same heater winding for all valves, e.g. in a receiver or test equipment. The vacuum rectifier is characterized by a relatively high internal resistance between anode and cathode which renders it less susceptible to damage from temporary overload.

The maximum d.c. load current available from the larger type of full-wave rectifier is 250 mA, and the maximum a.c. input voltage to each anode is 600–1000 volts dependent upon the type of filter circuit employed. The two anodes of a full-wave rectifier may be paralleled via current equalizing resistors of 50–100 ohms if higher current output is required. Two such valves used in the full-wave circuit will provide twice the output current of one valve, with some improvement in regulation.

Gas-filled or mercury vapour rectifiers are used when the current and voltage requirements are in excess of the ratings of high vacuum types. They are normally directly heated half-wave types so that two valves are required in the full-wave circuit. Both gas-filled and mercury vapour rectifiers have a low internal voltage drop of 10–20 volts independent of load current. In the case of mercury vapour types there are certain operating limitations which must be observed, e.g. the h.t. voltage must not be applied until all the liquid mercury within the bulb has been vaporized by pre-heating the filament and the bulb temperature must be kept within prescribed limits set by the manufacturer. Non-observance of these points entails premature failure of the rectifier. Gas-filled types are less critical in operation but the filament must be allowed to reach its normal operating temperature before applying h.t.

Gas-filled valves and to a lesser extent mercury vapour types can produce oscillation noise and for this reason block-bias switching of transmitters may, at frequencies below about 3·5 Mc/s give rise to excessive noise in a nearby receiver. This noise does not, however, appear on the carrier as it occurs only when the valves are virtually non-conducting.

Rectifier Ratings

Valve and semiconductor rectifiers are subject to limitations in respect of peak inverse voltage and peak current handling capability, both when delivering power to a load and when switching on. In this connection it should be noted that when h.t. voltage is applied to a rectifier (assuming the cathode is at normal operating temperature in the case of a valve) having a capacitor input filter, the capacitor when uncharged presents a virtual short circuit to the rectifier. Under these conditions a very large current flows, limited only by the internal impedance of the transformer and the rectifier itself until the capacitor is charged. In order to ensure that an acceptable life is obtained from the rectifier the manufacturers provide recommendations in respect of the minimum value of series impedance which should be present to limit this surge current to a safe value. This information is often presented in the form of rating charts, from which suitable operating conditions may be selected with a minimum of calculation. A study of the manufacturers' data on maximum ratings will be well repaid in long life when the rectifier is correctly chosen and the associated circuitry so designed that none of the ratings are exceeded.

Operation of Rectifiers

It is important to ensure that the filament or heater of the valve rectifier is operated at the correct voltage, measured at the valve pins. Reduced life expectancy and performance will occur if the heater or filament is under-run. The larger types of valve rectifier take considerable heater or filament current and an adequate gauge of wire should be used when wiring up the valveholder. The choice of valveholder is also important as in many instances each electrode of the valve is at a high potential to earth. Ceramic valveholders spaced away from the chassis if necessary are recommended for high voltage rectifiers operating at 500 volts or more.

It is desirable to mount rectifier valves, whether vacuum, mercury vapour or gas filled types, in the vertical position with base downwards, to assist ventilation and avoid short circuits due to a sagging filament touching the anode, particularly when low impedance directly-heated rectifiers are employed.

When mercury vapour rectifiers are first put into service or after a prolonged period of storage, they should be operated at normal filament voltage for at least 30 minutes prior to applying the anode voltage. This is to ensure that any liquid mercury which has become attached to the filament is evaporated, as otherwise arc-back may occur with consequent damage to the filament and its oxide coating. After the valve has been put into operation, a heater warm up time

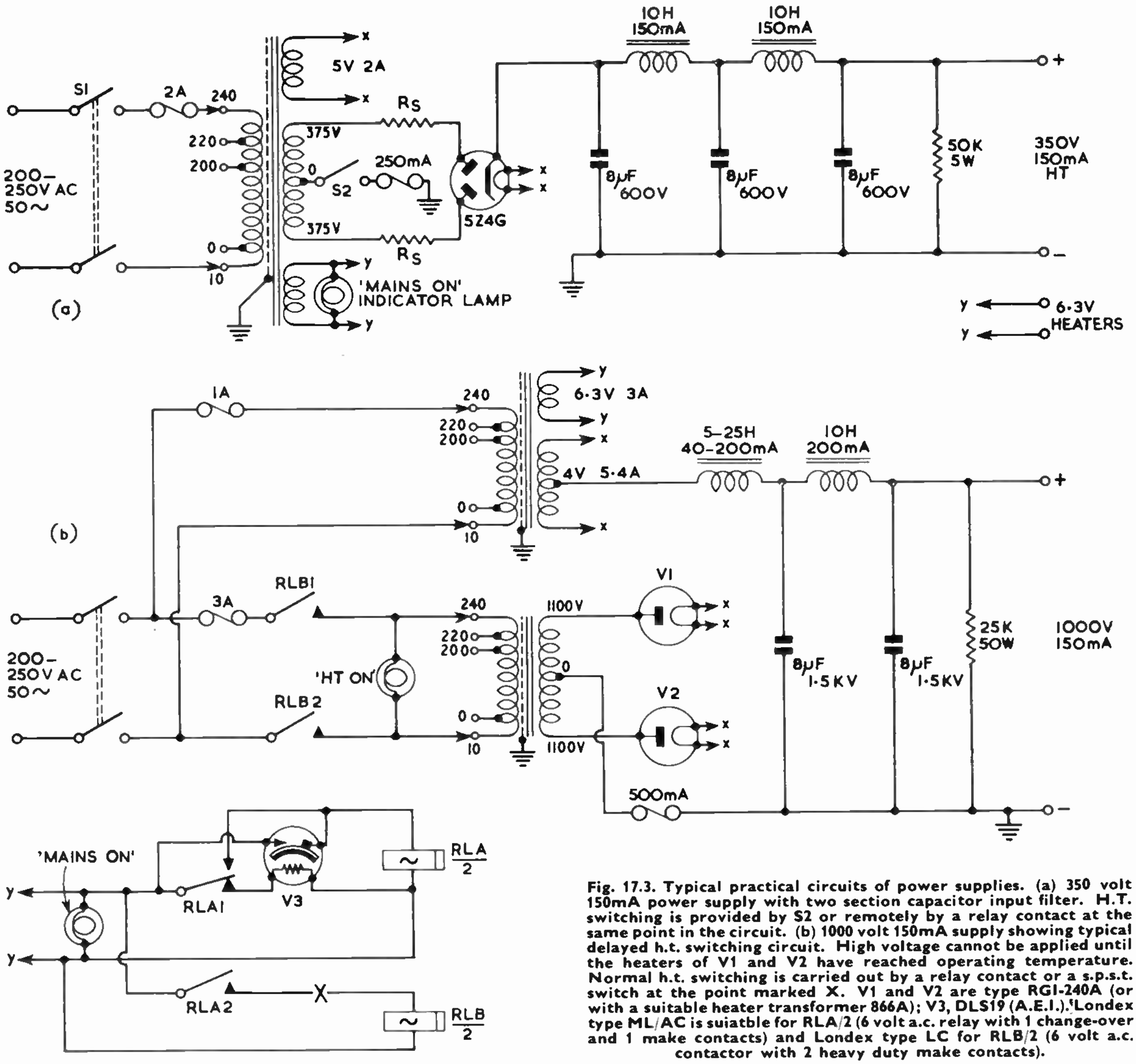


Fig. 17.3. Typical practical circuits of power supplies. (a) 350 volt 150mA power supply with two section capacitor input filter. H.T. switching is provided by S2 or remotely by a relay contact at the same point in the circuit. (b) 1000 volt 150mA supply showing typical delayed h.t. switching circuit. High voltage cannot be applied until the heaters of V1 and V2 have reached operating temperature. Normal h.t. switching is carried out by a relay contact or a s.p.s.t. switch at the point marked X. V1 and V2 are type RGI-240A (or with a suitable heater transformer 866A); V3, DLS19 (A.E.I.). Londex type ML/AC is suiatble for RLA/2 (6 volt a.c. relay with 1 change-over and 1 make contacts) and Londex type LC for RLB/2 (6 volt a.c. contactor with 2 heavy duty make contacts).

of 30–60 seconds will be adequate before anode voltage is applied. The use of a thermal delay switch to delay application of the h.t. voltage to the rectifier valves until the heaters have warmed up is recommended. A typical circuit employing a thermal delay switch for this purpose is shown in **Fig. 17.3.**

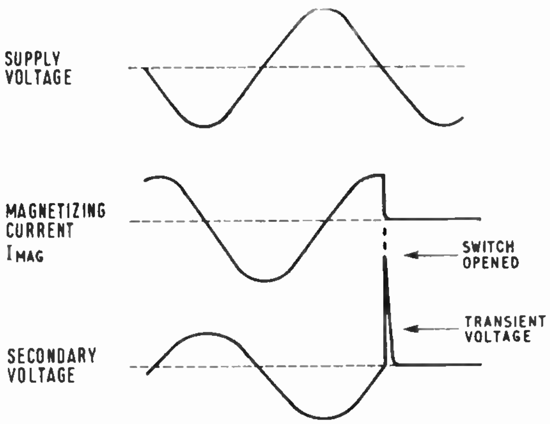


Fig. 17.4. Switch-off transient generated from transformer or choke.

Silicon Rectifiers

Silicon rectifier diodes are rapidly replacing thermionic rectifiers, especially as the small size and relative lack of heating results in smaller and cooler-running power supply units.

Silicon diodes as with all other types of semiconductors do not tolerate excessive voltage or current transient surges, and it is necessary to protect these rectifiers against transients which may arise from power system switching on the power company's distribution network, lightning striking the power lines directly or by induction, or switching on or off the supply to the transformer.

With any rectifier having a low forward resistance (e.g. silicon, gas-filled or mercury-vapour rectifiers) consideration must be given to the effect of operating into a capacitive load such as that presented by the reservoir capacitor of the conventional ripple filter. At the moment of switching on, the reservoir capacitor represents an almost direct short-circuit across the output of the rectifier. Additionally, if the value of the capacitor is high, then large peak currents will flow during the relatively short duty-cycle of the rectifier. The peak current which will flow at switch-on will depend upon the precise moment in the a.c. cycle when the switch is closed; the worst possible conditions occur when the switch closes exactly at the peak of the sine wave. A silicon rectifier has no protective warm-up time so that it is possible for a heavy surge of current to flow into the reservoir capacitor.

As with mercury vapour and other low impedance rectifiers, this problem can be largely solved by using a choke-input rather than a capacitive-input, ripple filter; this change will also improve voltage regulation, though at the cost of some lowering of d.c. voltage output. However, a choke is a bulky and fairly expensive component and there are many applications where this solution is inconvenient. In such circumstances it is highly advisable to fit a surge limiting resistor between the rectifier and the reservoir capacitor, even though this results in some loss of rectification efficiency and the generation of additional heat. An acceptable value for a surge limiting resistor is roughly 15 to 30 ohms per diode;

Transient voltages are generated whenever a magnetic component (such as a transformer or choke) is energized or de-energized. For example, transient voltages of a peak amplitude several times the normal value can be generated through the sudden interruption of current by switching (see **Figs. 17.4 and 17.5**).

There are two main design problems: to make sure that normal working conditions are well within the p.i.v. rating of each diode; and to limit high voltage and current surges to a safe working level.

Some makers distinguish between these two problems by adopting the term " crest working voltage " (c.w.v.) for the maximum peak voltage which occurs under normal repetitive conditions as opposed to the transient rating which is a higher voltage which can be accepted for a momentary period.

Calculating P.I.V. Ratings

The first step when considering the design of a power supply using silicon diodes is to calculate the basic peak inverse voltages (that is crest working voltages) which will appear across each diode. **Fig. 17.6** shows the situation for the most commonly used circuits, these theoretical figures ignore transformer and rectifier voltage drops, etc., but this is of little practical importance in the power units under discussion.

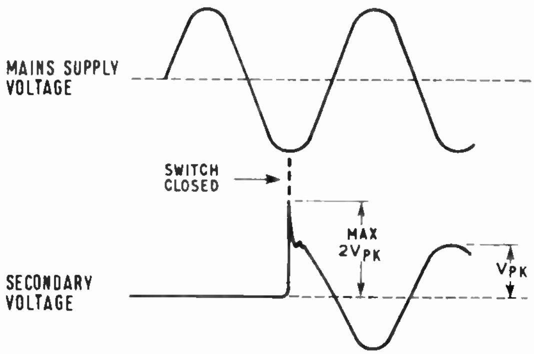


Fig. 17.5. Switch-on transient generated from transformer or choke.

Consider the bi-phase or full-wave circuit of Fig. 17.1 (b). The p.i.v. across each leg of the rectifier is approximately equal to $2\sqrt{2}$ (about 2·8) times the alternating voltage (measured in r.m.s.) across each half of the transformer secondary winding. Thus with a 350-0-350 volt transformer, the p.i.v. across each rectifier section will be roughly 2·8 × 350 volts, which is 980 volts. To provide a reasonable margin of safety, it will be advisable for each section to have a p.i.v. of about 1200 volts. A single diode with this rating would tend to be rather expensive; in practice, two 600 p.i.v. or 800V p.i.v. types would be used. The use of diodes of mixed ratings is not recommended.

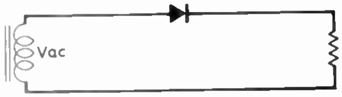
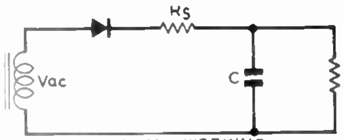
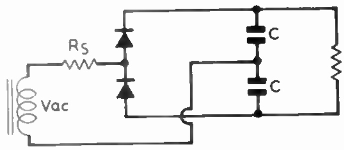
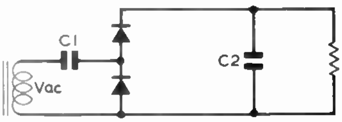
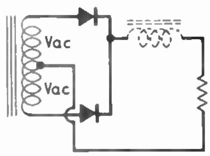
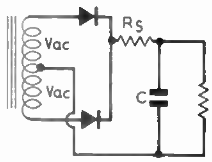
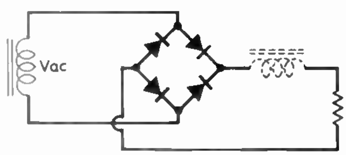
Circuit		P.I.V. Across Diode	Max. Vac Related to Diode P.I.V. Rating	Max. mean D.C. Output Current Related to Diode Rating	Approx. Mean D.C. Output Related to A.C. input voltage
Basic Single Phase Half-Wave Resistive Load		1·4 Vac	$\frac{PIV}{\sqrt{2}}$	I	0·45 Vac
Single Phase Half-Wave Capacitive or Battery Load	 C = 1·4 Vac WORKING	1·4 Vac	$\frac{PIV}{\sqrt{2}}$	0·8 I approx.	1·4 Vac* (Maximum)
Single Phase Voltage Doubler Symmetrical	 C = 1·4 Vac WORKING	2·8 Vac	$\frac{PIV}{2\sqrt{2}}$	0·8 I approx.	2·8 Vac* (Maximum)
Single Phase Voltage Doubler Cascade	 C1 = 1·4 Vac WORKING C2 = 2·8 Vac WKG	2·8 Vac	$\frac{PIV}{2\sqrt{2}}$	0·8 I approx.	2·8 Vac* (Maximum)
Single Phase Full Wave Resistive or Inductive Load		2·8 Vac	$\frac{PIV}{2\sqrt{2}}$	2 I	0·9 Vac
Single Phase Full Wave Capacitive or Battery Load	 C = 1·4 Vac WORKING	2·8 Vac	$\frac{PIV}{2\sqrt{2}}$	1·6 I approx.	1·4 Vac* (Maximum)
Single Phase Bridge Resistive or Inductive Load		1·4 Vac	$\frac{PIV}{\sqrt{2}}$	2 I	0·9 Vac

Fig. 17.6. Ratings in other connections related to basic ratings.

* Figures quoted are maximum, i.e. no load. The actual voltage depends on the load and capacitor value. Any voltage drop due to the diodes and the transformer impedance is ignored.

Note: Vac is in r.m.s. values. RS is a surge limiting resistor.

In this calculation it has been necessary to distinguish between the a.c. input to the rectifiers and the theoretical d.c. output. This is because the multiplication factor to obtain the p.i.v. from the theoretical d.c. output is 3·14, appreciably higher than the 2·8 used.

Voltage equalization is required when a number of diodes are connected in series as the reverse resistance of silicon diodes varies appreciably. This problem can be overcome by connecting an equalizing resistor across each diode (as should be done incidentally for series-connected VR-tubes and electrolytic capacitors) see **Fig. 17.7.** The value of the resistor must be relatively small compared with the reverse resistances of the diodes. The peak inverse voltages are then shared almost equally between the diodes despite any differences in the diodes themselves. A typical value would be 300 to 400 ohms per volt of the p.i.v. rating of the diode. These resistors will, of course, allow a certain amount of a.c. to bypass the diodes but this forms only a small part of the ripple current.

Even with equalizing resistors it is advisable not to operate series-connected diodes right up to their full p.i.v. ratings if this can be avoided. A good safety margin should be allowed to meet unavoidable current and voltage surges.

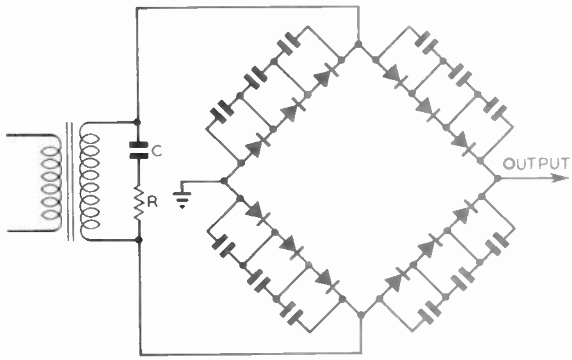


Fig. 17.8. Disc ceramic capacitors used as transient suppressors in high-voltage silicon diode power supply. For a 2 kV transformer C would be rated at 5 kV and the remaining capacitors (0·002 μF) 1·2 kV. The actual number of diodes in each arm would depend on their p.i.v. ratings.

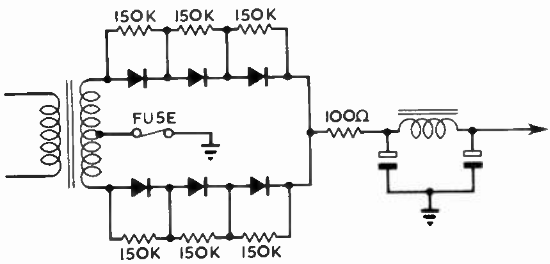


Fig. 17.7. Typical power supply using series connected silicon diodes illustrating some of the precautions discussed in the text.

As an alternative to resistors, equalizing capacitors (**Fig. 17.8**) may sometimes be recommended for this purpose, forming an a.c. potentiometer.

There are various circuit methods of reducing the effects of transients arising from switching surges and mains over-voltages.

One of the most satisfactory methods of restricting these is to connect suitable series connected capacitors and resistors across either primary of secondary of the power transformer.

Suitable values for these components can be calculated from the following expressions.

For use across the primary:

$$C = \frac{100 \text{ I mag}}{V} \mu F$$

$$R = \frac{150}{C_{\mu F}} \text{ohms}$$

For use across the secondary:

$$C = \frac{225 \text{ Imag } T^2}{V} \mu F$$

$$R = \frac{200}{C_{\mu F}} \text{ ohms}$$

where

Imag = Magnetising r.m.s. current.
For this purpose a value of 10 per cent of the primary or secondary as appropriate may be used.

V = Transformer primary r.m.s. voltage.

T = Transformer $\frac{\text{primary}}{\text{secondary}}$ r.m.s. voltage.

Controlled avalanche rectifiers are likely to be increasingly used, these have a carefully controlled non-destructive internal avalanche breakdown to protect the junction.

Temperature Effects

The characteristics of silicon junction rectifiers, like those of other semiconductor devices, are sensitive to changes in temperature. Although silicon diodes can be used at temperatures well above the boiling point of water (100° C), the maximum safe d.c. output current falls quite rapidly with increasing temperature, so the maximum output of silicon diodes must be reduced with increasing ambient temperature.

The extent of the derating necessary can be quite severe although the figures vary a little with type. Thus a 250 mA diode might have to be restricted to about 100 mA if required to operate at a high temperature.

The amount of heat generated in the rectifier itself is low, so that it should usually be possible to ensure that the diodes run cool. They should be mounted so that a reasonable amount of cool air can reach all outside surfaces; they should be kept away from heat producing valves and components. The temperature of any equalizing or surge-limiting resistors can be kept low by fitting generously rated types. Try to ensure that the diodes are not mounted in currents of warm air.

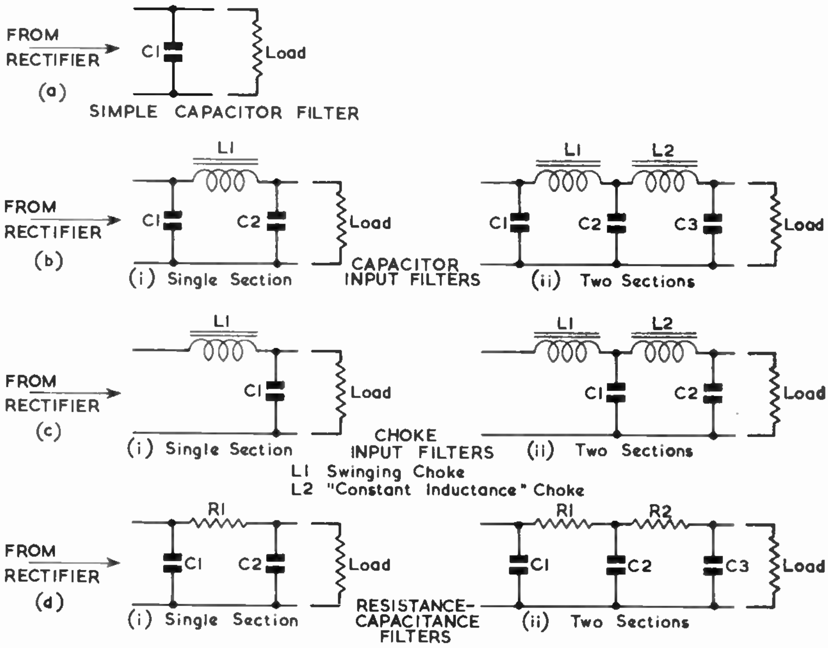


Fig. 17.9. Typical filter circuits. (a) Simple capacitor filter. (b) Capacitor input filters. (c) Choke input filters. (d) Resistance capacitance filters.

SMOOTHING CIRCUITS

The pulsating d.c. from the rectifier circuits illustrated in Fig. 17.1 is not sufficiently constant in amplitude to be used to operate equipment directly and means must be provided to smooth out the variations. This function is carried out by means of a combination of series inductors or resistors and parallel capacitors termed a filter or smoothing circuit. As described earlier, the output from the rectifier consists of d.c. plus a superimposed a.c. component and expressed in its simplest terms the action of the filter is to separate one from the other and allow the d.c. to pass to the load while the a.c. is bypassed and confined to the filter circuit. Apart from its basic purpose of separating the a.c. and d.c. components of the rectifier output, the filter has a profound effect upon the operation of the rectifier itself and the d.c. voltage output, the voltage regulation of the supply and the maximum load current which can be safetly drawn depend to a large extent upon the design of the filter.

Power supply filters fall into two classifications depending upon whether the first component of the filter network is a capacitor or an inductor (choke). Capacitor input filters

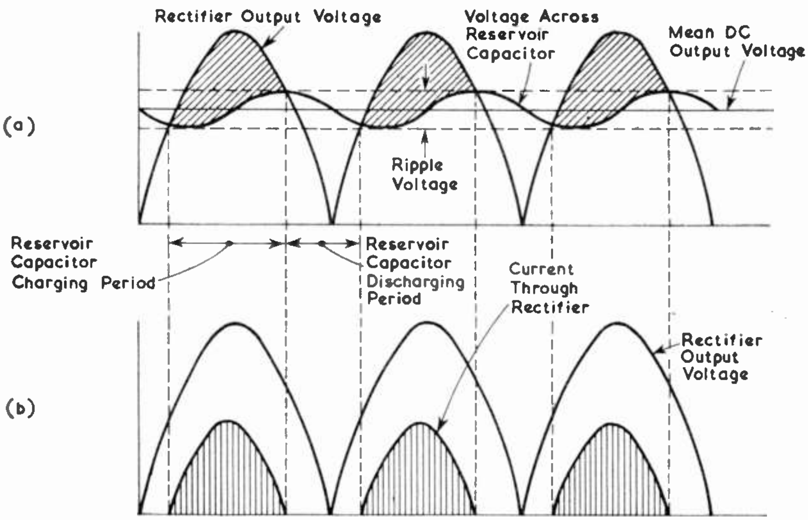


Fig. 17.10. Curves illustrating the output voltage and current waveforms from a full-wave rectifier with capacitor input filter. The shaded portions in (a) represent periods during which the rectifier input voltage exceeds the voltage across the reservoir capacitor causing charging current to flow into it from the rectifier.

are characterized by a relatively high d.c. output voltage for a given a.c. input voltage to the rectifier, but the regulation is poor and the rectifier is called upon to handle higher peak voltages and currents than when choke input is used. A choke input filter provides much better regulation when properly designed but the d.c. output voltage is lower. The resistance-capacitance filter is a variant of the capacitor input filter where resistors replace the inductors. The use of this type of filter is normally confined to low constant current applications because of the relatively high voltage drop across the series resistors and consequent poor regulation. Typical filter circuit configurations are shown in **Fig. 17.9.**

The Capacitor Input Filter

The simplest form of capacitor input filter is illustrated in **Fig. 17.9a** and consists of a single capacitor C1 placed in parallel with the output of the rectifier and the load resistance.

During the periods in which the rectifier is conducting C1 charges up to nearly the peak value of the a.c. voltage applied to the rectifier, and if no load current were drawn the voltage on C1 would remain constant. When a load is placed across the capacitor, it discharges into the load and the voltage across it falls until the arrival of the next pulse of current from the rectifier which recharges the capacitor to its original voltage. This process is repetitive and the resulting waveform of the output voltage is as shown in **Fig. 17.10(a)** for a full-wave rectifier. Thus the capacitor C1 acts as a reservoir of charge which is depleted by the load and replenished by the rectifier during the period when the rectifier output voltage exceeds the capacitor voltage **Fig. 17.10(b)**. C1 is therefore known as the *reservoir capacitor*.

The current which flows into C1 from the rectifier is very large and limited only by the internal resistance of the rectifier and the total effective resistance of the transformer windings. In order not to exceed the maximum surge current rating of the rectifier it may be necessary in practical circuits to increase the effective resistance of the transformer by inserting additional resistance between the transformer and the anode or anodes of the rectifier. The total source resistance as seen by the reservoir capacitor will determine how closely the capacitor voltage will approach the peak value of the a.c. voltage during the charging period, i.e. the lower the source resistance the more closely the two voltages will approach each other. When a load is connected the capacitor voltage falls between pulses of charging current the amount by which it falls being determined by the load resistance.

A decrease in the value of the capacitor or load resistance causes a greater fall in voltage. The amplitude of the variation in capacitor voltage or *ripple voltage* which varies about a mean value corresponding to the d.c. output voltage, is therefore determined by the load resistance, and to a lesser extent by the source resistance as well as the value of the capacitor. A graph showing approximate values of ripple voltage expressed as a percentage of the d.c. output voltage plotted against load resistance is shown in **Fig. 17.11** for three typical values of reservoir capacitor.

A reservoir capacitor of 16 μF represents the maximum value which should normally be used with a full wave rectifier and 32 μF for a half-wave circuit. Increasing these values will produce relatively small improvements in voltage regulation and ripple but are undesirable because of the increased surge current through the rectifier and also because of increased ripple current through the reservoir capacitor. The amount of ripple current flowing is of great importance when an electrolytic capacitor is used as a reservoir, and

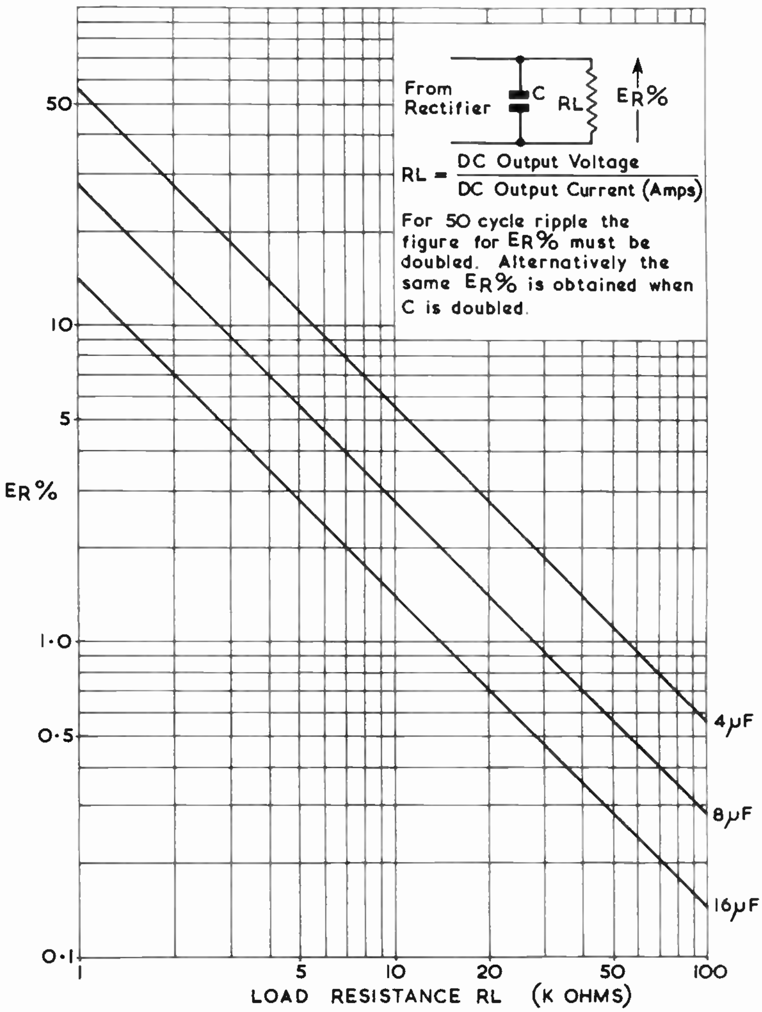


Fig. 17.11 Curves showing 100 c/s ripple component as a percentage of the d.c. output voltage across a reservoir capacitor.

too high a value of ripple current will lead to overheating and breakdown of the capacitor. A typical value of maximum permissible ripple current is 260 mA for a 16 μF 450 volt working electrolytic of plain foil construction and 122 mA for an etched foil type. Provided that a d.c. load current of not more than 120 mA is taken when a plain foil type is used and 70 mA in the case of etched foil types, there is little danger of exceeding the maximum ripple currents stated.

The d.c. working voltage of the reservoir capacitor whether electrolytic or paper should be at least equal to the peak value of the a.c. applied to the rectifier, i.e. $\sqrt{2}$ times the transformer h.t. secondary voltage in half-wave and bridge circuits and $\sqrt{2}$ times the half secondary voltage in the biphase circuit. The peak inverse voltage which the rectifier must withstand approaches $2\sqrt{2}$ times the a.c. voltage at light loads in the half-wave and biphase circuits and $\sqrt{2}$ times the a.c. voltage in the bridge circuit. Table 17.1 summarizes the voltage and current relationships existing in the circuits of Fig. 17.1.

Choke Input Filters

The choke input filter is used when good regulation of the output voltage is important as for example in the h.t. supply for a class AB2 modulator stage, where the maximum load current may be 4–5 times the minimum. The output voltage obtained is somewhat lower than when a capacitor input filter is used and allowance must be made for this when selecting a suitable h.t. transformer.

Full-wave rectifier circuits are normally used with choke input filters as the output voltage would be inconveniently low if the half-wave circuit is used. Neglecting the voltage drop in the choke and rectifier, the mean d.c. output voltage of a choke input filter fed by a full-wave rectifier is 0·9 times the applied a.c. voltage. The choke input filter is illustrated in Fig. 17.9(c) and consists of an iron cored inductor L1 in series with the load and a capacitor C1 across the load. The action of the inductor L1 is to attempt to maintain a constant flow of current through the rectifier throughout the a.c. cycle. If the inductance of L1 were infinite the current would be constant, but in practice this cannot be achieved, and some variation of current does occur and appears as a ripple, but provided the peak-to-peak amplitude of the variation does not exceed the d.c. load current satisfactory constant current operation will be obtained. The minimum value of inductance required to do this is called the *critical inductance*, L_C. Assuming the source resistance as seen by the rectifier is small compared with the load resistance R_L (d.c. output voltage/d.c. output current), then L_C in Henrys is given by the following equation

$L_C = R_L/940$ for a full wave, single phase, 50 c/s supply.

It will be seen from the equation for critical inductance that the value of L_C becomes lower as the load current is increased. This permits the use of a swinging choke for L1 with considerable economy in cost over that of an ordinary "constant inductance" choke. The inductance of a swinging choke is high at low currents falling off to perhaps 20-25 per cent of its low current value as the current flowing is increased to its maximum rating. This characteristic permits the use of a smaller and less expensive component, a typical specification of which would be 25 Henries at 50 mA, falling to 5 Henries at 250 mA.

If the minimum inductance is insufficient the filter will

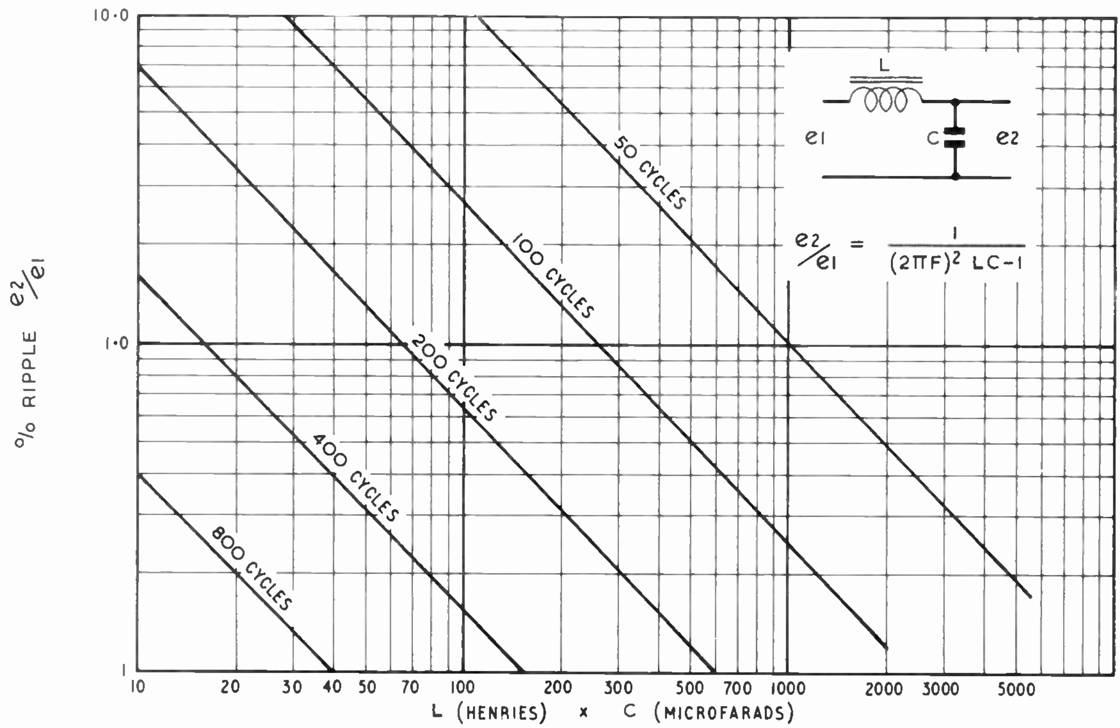


Fig. 17.12. Relationship between percentage ripple and product of LC.

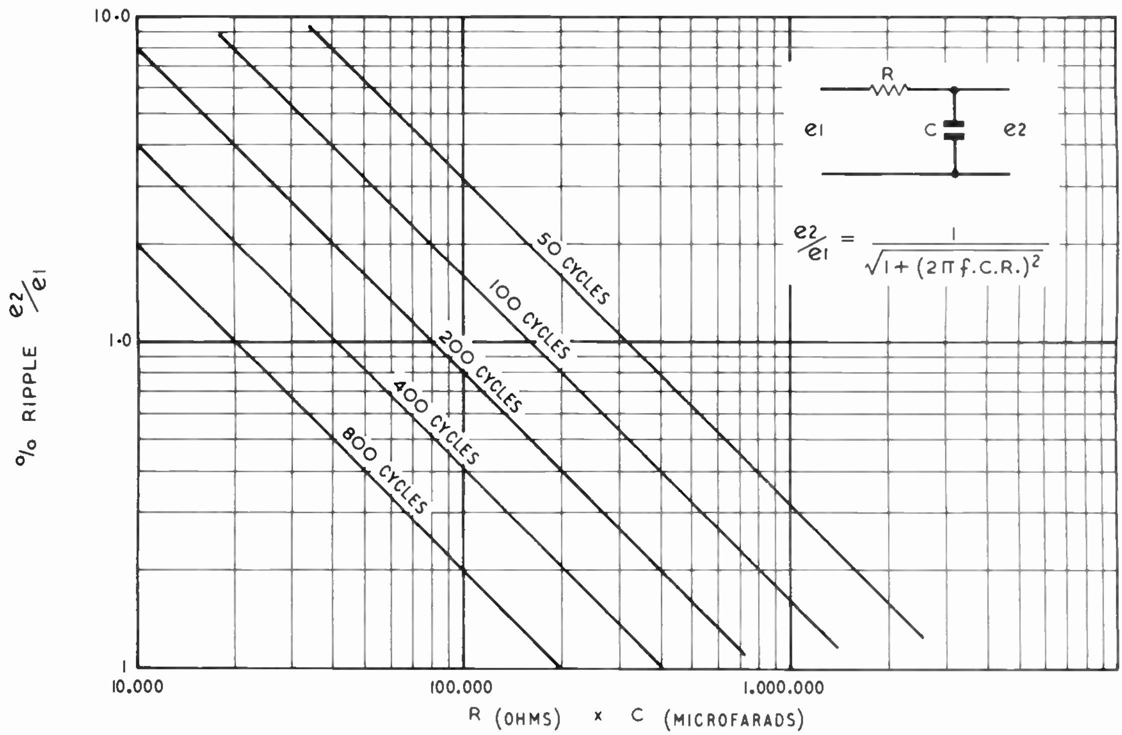


Fig. 17.13. Relationship of percentage ripple and product of RC.

function as a capacitor input filter with its attendant rise in output voltage at light loads. At light loads, i.e. high values of R_L, the value of L_c required for proper operation would be inconveniently high, and it is usual to connect an artificial load or bleeder resistance across the output of the supply to ensure that sufficient current is drawn at all times to enable a reasonable value of L_c to be used. If the value of L_c has already been determined by available components, then the approximate value of bleeder current required (i.e. maximum value of R_L permissible) is given by the following expression:

$$\textit{Bleeder current in milliamperes} = \frac{\textit{D.C. output voltage}}{L_{MAX}\ (\textit{in Henrys})}$$

where L_{MAX} = maximum inductance of the swinging choke.

Further attenuation of the ripple component is achieved by the parallel capacitor C_1 and the overall reduction of ripple by a choke input filter is relatively high compared with the capacitor input filter of Fig. 17.9(a). A graph showing the relation between percentage ripple and the product LC in Henrys and microfarads is shown in **Fig. 17.12.** Values of L and C which lead to conditions approaching series resonance at the ripple frequency should be avoided as the ripple voltage would build up to high values, which apart from defeating the object of the filter, may cause excessive rectifier peak currents and unduly high peak inverse voltages. The product of L and C producing resonance is 2·53 assuming 100 c/s ripple and 10·12 assuming 50 c/s ripple, and the product of C in microfarads and L in Henrys (maximum current value) should be at least twice these figures.

The simple capacitor input and choke input filters described will not reduce the ripple voltage to a low enough value for use in power units supplying the low level stages of transmitters and receivers and further *LC* filter sections will be required as shown in Fig. 17.9 (b) and (c).

As a guide to the number of filter sections needed the following permissible ripple percentages may be quoted for various types of load:

C.w. transmitters	5 per cent
Telephony transmitters	1 per cent
Class B modulators	0·25 per cent
V.F.O.'s, speech amplifiers and receivers, s.s.b. generators	0·01 per cent

For most purposes one additional section will be adequate after a choke input filter, and one or two sections, depending upon the required ripple attenuation, after a simple capacitor input filter. The capacitors used in the additional sections should have a working voltage at least equal to, and preferably greater than, the peak a.c. voltage. The chokes should be constant inductance types rated to provide the required inductance at the maximum current taken by the load.

Resistance Capacitance Filters

The substitution of resistors for inductors in the filter sections following the reservoir capacitor provides an economical and space saving method of ripple attenuation when the regulation of the supply is not important and the relatively high voltage drop across the series resistors can be tolerated. The ripple attenuation x (where x = output/input ripple) plotted against the values of RC from 10^2 to 10^6 is shown in **Fig. 17.13** where RC is the product of series resistance in ohms and parallel capacitance in microfarads. The same requirements regarding the capacitor working voltage apply as when LC sections are used. It is also necessary to ensure that the series resistors are of adequate power handling capacity to avoid overheating. The required value of dissipation in watts may be computed from the usual formula $W = I^2R$ where I is in amperes and R in ohms, and the voltage drop from the formula $V = IR$ using the same units. To reduce the voltage drop, the highest practicable values of C should be used. This will also be advantageous in providing a low impedance path to audio frequencies, so reducing the possibility of low frequency instability when this smoothing system is used in conjunction with multistage audio frequency amplifiers.

SELECTION OF A SUITABLE TRANSFORMER

The transformer secondary voltage required for a given d.c. output voltage is best obtained from the manufacturers' characteristic curves for the rectifier to be used. Due allowance should be made for the voltage drop in the filter chokes, and also for the fall in transformer output voltage when current is drawn. A fall in transformer output voltage of from 5 to 10 per cent of the open circuit voltage may be expected at full load. In most instances a commercially manufactured transformer will be employed in amateur equipment and it will be necessary to specify the maximum d.c. load current and type of rectifier circuit to be used when ordering an h.t. transformer.

The source resistance presented to a rectifier is determined by the winding resistance of the h.t. transformer plus any additional resistance inserted to bring the value up to the minimum required by the operating conditions of the rectifier. The effective transformer winding resistance is given by the following equation:

Effective winding resistance = $N^2Rp + Rs$ ohms

where N = transformer turns ratio primary to secondary and Rp and Rs are the d.c. resistances of the primary and secondary windings. When the transformer feeds a biphase rectifier, the turns ratio from primary to half h.t. secondary should be taken and also the resistance of half the total h.t. secondary winding.

Heater windings should be adequately rated for the total heater current to be drawn. Any deficiencies in this respect will give rise to excessive voltage drop in the transformer windings causing overheating and insufficient heater voltage to the valves themselves. When a heater winding is used to supply valves in a circuit susceptible to hum pickup it may be desirable to use a centre tapped heater winding, the centre tap being earthed and the heater wiring carried out in twisted pair. Transformers having heater windings should have an adequate number of tappings on the primary to enable the heater voltage to be adjusted to its correct value with changes in the supply voltage. The mains voltage is normally held to within 10 per cent of its declared value in U.K. and therefore taps at 200, 220 and 240 volts are desirable.

Valves taking a heavy heater current are usually supplied from a separate heater transformer mounted near the valve-holder to avoid the necessity of running excessively heavy heater leads.

The provision of an earthed electrostatic shield between the primary and other windings of mains transformers is recommended. Apart from its basic function of reducing the interwinding capacitance, the shield provides an added factor of safety in the event of breakdown between high voltage windings and the primary of the transformer. Reduction of the interwinding capacitance results in less mains-borne noise being coupled into the equipment via the power supply.

POTENTIAL DIVIDERS AND SERIES DROPPING RESISTORS

Voltages of lower value than that provided by the power supply may be obtained by the use of a potential divider which is a form of potentiometer where the tapping points on the resistance are fixed or adjustable instead of being continuously variable. As many tapping points as desired may be used to give various voltages, but each additional point from which current is taken will affect the voltage at each of the other tapping points. The extent to which the output voltage varies depends upon the current taken from each point and the resistance between the tapping points. The lower the total divider resistance, the better will be the voltage regulation at the tapping points.

By the application of Ohm's Law, it is a relatively simple matter to calculate the resistances of each section of the potential divider network, provided the voltage and current taken at each tapping point is known. **Fig. 17.14** shows a typical arrangement with two tapping points at B and A, providing voltages V1 and V2 at currents I1 and I2. The resistor R3 acts as a bleeder and determines the minimum current I3 flowing through the divider network when the loads at A and B are removed. A suitable value for I3 would be 10 per cent of the total load, this being a suitable compromise between the degree of voltage regulation obtained and the wastage of current through the divider chain. Having fixed a value for I_3 the procedure is as follows:

Calculate $R3 = V1/I_3$
Calculate $R2 = (V2 - V1)/(I_1 + I_3)$
Calculate $R1 = (V - V2)/(I_1 + I_2 + I_3)$

where V = supply voltage and all currents are in amperes.

The wattage rating of the resistors may be calculated from the usual formulae, $W = VI = V^2/R = I^2R$, V, I, and R being in volts, amperes and ohms respectively.

In order to utilize the current in I_3, which is otherwise wasted, it may be convenient to insert a suitable relay in series with R1, adjusting the value of the latter appropriately. The relay will operate when the supply voltage is applied and its contacts used to operate other circuits as required. The insulation of the relay should be adequate for the functions its contacts are required to perform.

Where only one reduced voltage is required at constant current, a simple series resistance may be used the value of which can be calculated as follows:

$$\textit{Required series resistance in ohms} = \frac{\textit{Voltage dropped across resistor}}{\textit{Current taken by load in amperes}}$$

In many instances it will be found necessary to decouple the load side of the resistor with a suitable capacitor to avoid instability due to the high source impedance presented by the series resistor. Decoupling may also be required at the tapping points of the voltage divider. A capacitor of at least 2 μF will be required to decouple audio frequencies and 0·1 μF for radio frequencies.

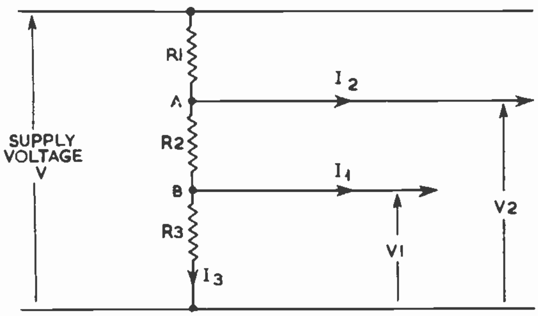


Fig. 17.14. Potential divider circuit.

VOLTAGE STABILIZATION

Certain types of circuit require a source of stabilized voltage for their correct operation. Examples of such circuits are local oscillators and *Q* multipliers in receivers and variable frequency oscillators in transmitters. A stabilized grid bias supply is also desirable for the correct operation of class AB2 and class B amplifiers used in modulators and s.s.b. transmitters. Changes in frequency or operating conditions due to variations in mains voltage and load variations may be eliminated or greatly reduced by the use of voltage stabilization circuits.

A moderate degree of stabilization is obtained by the use of one or more cold cathode stabilizer valves or Zener diodes (described in Chapter 2—*Valves* and Chapter 3—*Semiconductors* respectively). These devices will reduce voltage variations to within 1 per cent to 3 per cent of the nominal value of stabilized voltage, which is fixed by the device selected and cannot be varied. Cold cathode stabilizers are available for operating voltages between 50 and 150 volts, Zener diodes from about 3·3 volts to 150 volts. Current ratings of cold cathode stabilizers are limited to about 50 mA maximum; Zener diodes may be rated up to several amperes when used with a suitable heat sink.

A much closer degree of stabilization is obtainable from a hard valve or transistor stabilizer circuit, where cold cathode tubes or Zener diodes are used as voltage reference sources only. A very low output impedance, typically less than 0·1 ohm may be obtained by suitable circuit design and the output voltage is variable manually over a wide range from zero upwards if desired. These advantages are gained at the expense of relatively complex circuitry as shown by the example in the practical circuit section.

Cold Cathode Stabilizer Circuits

The basic circuit of a cold cathode stabilizer is very simple, consisting of a resistor in series with the tube, the two being shunted across the unstabilized supply. The stabilized voltage is taken from the junction of the series resistor and anode of the tube, thus the circuit resembles a potential divider. The value of the series resistor is determined by the characteristics of the tube and the stabilized supply voltage. The characteristics are such that it will maintain a nominally constant voltage between its anode and cathode over a specified range of current through the tube typically 5 to 40 mA. The value of the series resistor must theoretically be selected to maintain the current within the specified minimum and maximum values under all conditions of operation. In practice it is found that any one value of resistor may not satisfy all the circuit conditions and a compromise is made, the value of series resistor being chosen, so that the current through the tube does not exceed the specified maximum when no current is being drawn by the load circuit (**Fig. 17.15 (a)**). The use of voltage stabilizers having a separate striking electrode will assist in ensuring that the tube strikes when the load circuit is drawing current. The striking electrode is connected to the unstabilized supply via a 270K ohms resistor.

The value of R_1 is given by the following expression:

$$R_1 = \frac{V_{i\ max} - V_s}{I_{s\ max}}$$

where $V_{i\ max}$ is the maximum value of the unstabilized supply voltage, V_s is the nominal stabilized voltage and I_s is the maximum permissible stabilizer current in amperes. The maximum variation of load current I_L should not be greater than ($I_{s\ max} - I_{s\ min}$), and should preferably be rather less to allow for variations in the unstabilized voltage.

The value of the unstabilized voltage must be at least 30–40 per cent greater than the stabilized voltage to ensure that the tube will strike when first switched on. Two or more tubes of similar current rating may be connected in series with a common limiting resistor to produce higher values of stabilized voltage as shown in **Fig. 17.15 (b)**. The unstabilized supply voltage should be at least equal to the sum of the *striking* voltages of each tube in the chain to ensure that they will conduct when the supply is first switched on. The shunt resistors shown in Fig, 17.15 (b) will assist in initial striking and should not be omitted. Current may be drawn from intermediate tubes if desired but the total load should not exceed ($I_{s\ max} - I_{s\ min}$).

The connection of voltage stabilizer tubes in parallel to increase the permissible load current is not recommended,

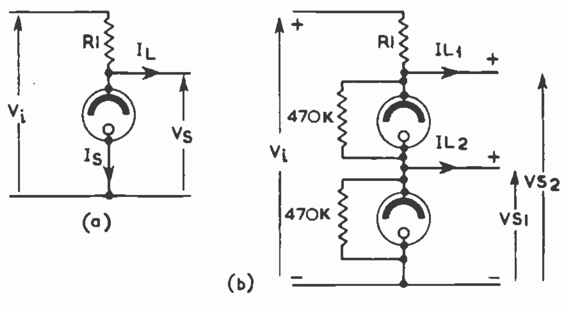


Fig. 17.15. Gas-filled voltage stabilizer circuits. (a) Simple voltage stabilizer arrangement. (b) Two or more stabilizer tubes in series to provide increased values of regulated voltage.

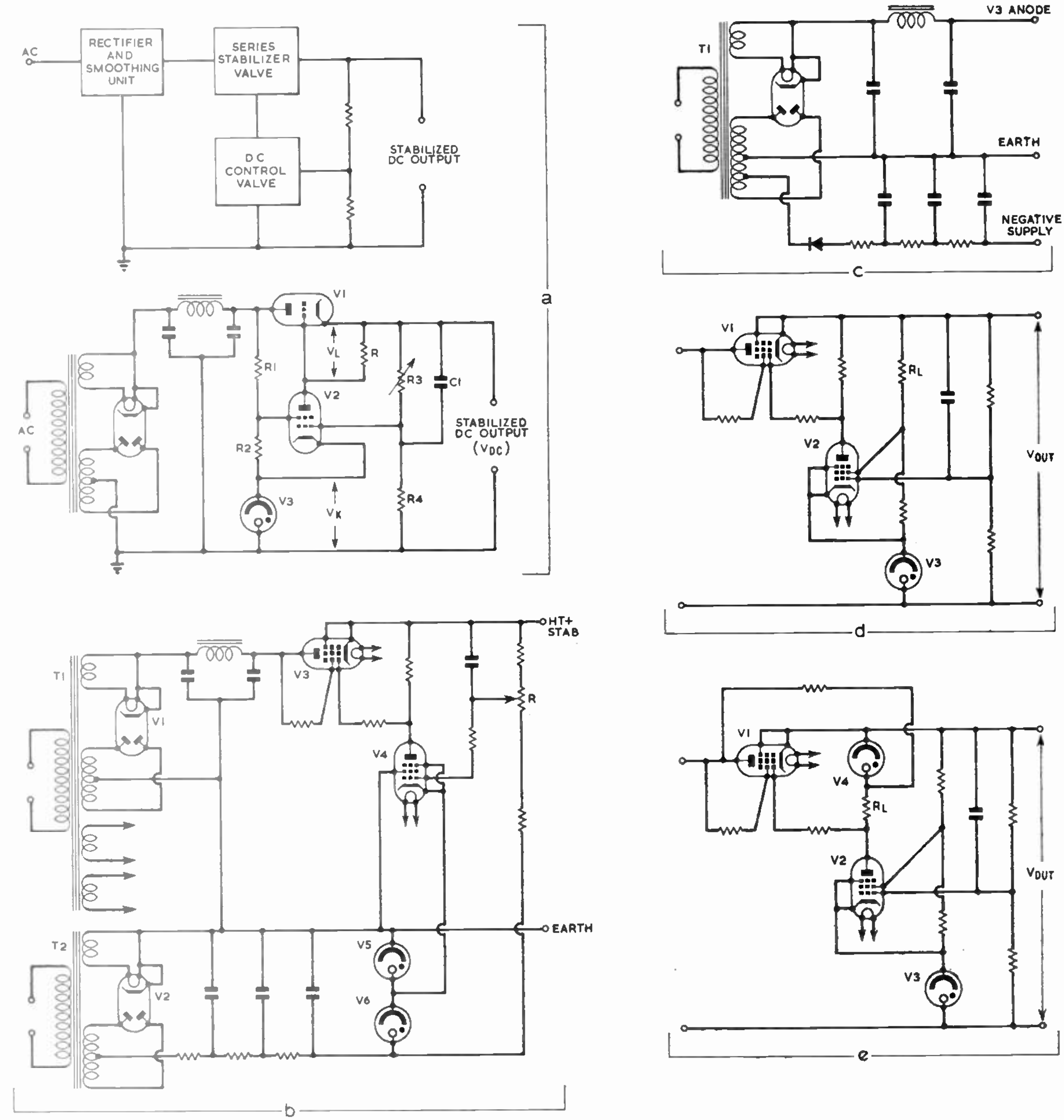


Fig. 17.16 (a). Typical circuits of series stabilized power units.

as one will strike, lowering the available striking voltage on the other tubes which will remain inoperative.

Voltage stabilizer tubes are polarised, i.e. they have a preferred direction of conduction and their stabilizing properties will be impaired if the rated maximum inverse voltage is exceeded.

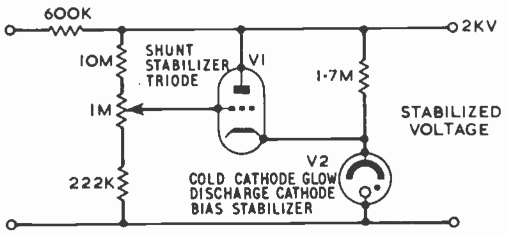


Fig. 17.17. Typical shunt stabilizer circuit.

Stabilized High Voltage Supplies

Another type of cold cathode stabilizer for high voltage, high impedance supplies makes use of the corona discharge in a gas and is used in much the same way as the glow discharge type. This operates at quite low currents with voltages of 350–500V and more. The maximum current is strictly limited and must not be exceeded since the discharge would change from the corona type to the normal glow type with a corresponding change to a much lower voltage, and the valve would be permanently damaged.

Typical circuits for these tubes are similar to those for the glow discharge type, and a suitable series resistor is used to set the current through the tube.

Corona stabilizers are suitable for fixed high voltages and any variable supply must have a suitable potential divider across the output to set the required voltage. However, a variable low impedance supply can be obtained by use of a high voltage shunt triode. A suitable circuit is shown in **Fig. 17.17.**

Valves for this type of application should be of suitable construction for high voltage use as under some conditions most of the voltage will appear across the valve. Special types are available as are valves rated for use at the high voltages developed in anode-modulated r.f. amplifier service.

Zener Diode Stabilizer Circuits

The Zener diode stabilizes by virtue of avalanche breakdown when a sufficiently high reverse voltage is applied and a series resistance is required to limit the current through the diode. A negative stabilized supply may then be taken from the ANODE of the diode. A similar marking convention applies as with other types of semiconductor diodes i.e. the cathode is indicated by a +ve sign or a red marking. In the larger types of Zener diode one terminal is usually a threaded stud to enable the device to be attached directly to a heat sink.

The minimum current through the diode should be sufficient to bring the operating point into the region of low dynamic resistance. This is particularly important when a diode stabilizing at voltages less than about 6 volts is to be used as the slope of the voltage/current characteristic curve is quite gradual at values of diode current less than 10–20 mA.

Like any semiconductor device the Zener diode is temperature conscious in respect of breakdown voltage and device dissipation. Diodes with a breakdown voltage below 5 volts exhibit a small negative temperature coefficient of the order of 1–2 mV/°C., and a similar value of positive temperature coefficient is found in diodes with a higher breakdown voltage. The small wire-ended diodes currently available have a device dissipation of the order of 0·5 to 3 watts at 25°C and stud-mounted diodes, 5–100 watts.

The value of series resistor to be used is calculated in a similar manner to that used with a cold cathode tube, the relevant formula in this case being $Rs = \dfrac{V_{IN} - V_Z}{I_Z + I_L}$

where Rs —series resistor in ohms
V_{IN}—unstabilized supply voltage
V_Z—stabilized voltage
I_Z—diode current in amperes
I_L—load current in amperes.

The greatest degree of stabilization for a given device is obtained when Rs is as large as possible, but to ensure that the required load current is available despite variations in supply voltage and diode voltage/current characteristics the value of Rs should also meet the criterion—

$$Rs\ (\text{max}) \leqslant \frac{V_{IN}\ (\text{min}) - V_Z\ (\text{max})}{I_Z\ (\text{min}) + I_L\ (\text{max})}$$

in amperes, volts and ohms as before. The device dissipation is then calculated ($W = I_Z V_Z$) under no load conditions and a suitably rated diode selected.

Series Stabilizer Circuits

There are many applications where a power supply is required which will give a steady d.c. output voltage which is not affected by any of the following:

(a) Mains fluctuations
(b) Load current variations
(c) Hum or noise.

Stabilization is normally defined as the change of output voltage with respect to the output voltage cut at a stated load, when the input is varied due to a mains fluctuation. This is stated as a percentage:

$$\frac{\delta\ V_{out}}{V_{out}} \times 100 \text{ per cent}$$

Regulation is defined as the change of output voltage with respect to the no load output voltage when the load current is varied.

This is stated as a percentage:

$$\frac{\delta\ V_{out}}{V_{out}} \times 100 \text{ per cent}$$

over the range from no load to full load current.

Fig. 17.16(a) shows a typical arrangement. The rectified and smoothed d.c. is fed to the anode of the series valve V1, and the output is taken from its cathode. Any change in output voltage is fed to the input of a d.c. coupled control amplifier V2, and its output is used to alter the impedance of the series valve, by changing the grid bias. The relationship of the amplifier input and output voltages is such that any drop in stabilizer output voltage is compensated by a reduction in impedance of the series valve. This in turn reduces the voltage drop across the valve and returns the output voltage almost to its original value. A rise in output voltage is compensated in a similar manner.

The amplifier input is taken from the potentiometer network R3 and R4 in order to keep the grid voltage slightly negative with respect to the cathode potential which is fixed by the reference tube V3.

If it is desired to adjust the output voltage of this circuit the resistor R3 should be made variable.

If a wide range of output is required it is generally advisable to provide several taps on the mains transformer. By this means it is possible to keep the voltage drop across the series valve to a reasonable value.

If a wide control of output voltage is required in one range only, it may be necessary to use several series valves in parallel, in order to keep the current in each valve low so that the anode dissipation rating is not exceeded at low output voltages. A number of valves may be connected in parallel if the required current is higher than the maximum obtainable from a single series valve. In order to minimize the effect of variations in characteristics it is necessary to insert small equalizing resistors, of the order of 50-330Ω, in series with each anode. This ensures that a valve taking a higher current is not over-dissipated. It may also be an advantage if a small resistor (of the order of 5-10Ω) is inserted in series with each cathode.

To obtain a more efficient elimination of ripple and noise, a capacitor C1 is connected between the d.c. output line and the control valve grid. This effectively short circuits the top half of the potential divider R3 and R4 for a.c. signals and enables the whole of the ripple or noise to be applied to the grid of the control amplifier valve in order to obtain a degree of cancellation.

The Requirements of a Series Stabilizer Valve

The three main requirements of a valve to be used in the series circuit of a voltage stabilizer are as follows:

(a) Low r_a

This is necessary as the voltage output is the difference between V_{in} and V_{a-k}. Thus, the lower the r_a and hence the voltage drop of the series valve, the lower will be the input voltage required for a given output.

(b) High g_m

A high g_m is required in order to give a high degree of control.

(c) High p_a

A relatively high anode dissipation rating is desirable, in order that a high output current can be drawn from the valve without over-dissipation. This is particularly necessary if the output of the stabilizer is to be adjustable. At low output voltages the anode to cathode voltage will be large and hence the anode dissipation high.

Requirements of a Control Amplifier Valve

A necessary requirement for good stabilization is that the control valve should have high gain. Therefore a pentode amplifier is normally used. An alternative is the use of a double triode in a cascode circuit.

A fixed cathode potential is essential in order that any change of grid voltage may give maximum change of grid-cathode potential and thus give maximum control. For this reason it is not possible to use a cathode resistor, as the cathode potential will then vary with that of the grid. A reference tube is generally used to hold the cathode at a constant potential.

In order to have a large signal at the control grid it is necessary to have a large d.c. voltage at the grid. This necessitates a high cathode potential, and this is normally of the order of 85-150V. The limit is reached when the anode-cathode voltage is too low to give adequate gain.

Where a wide range of output voltage is required, one effective method is to have available a negative supply as shown in **Fig. 17.16(b)**. By this method it is possible to vary the output from zero to maximum—the limit being determined by the series valve or valves used (V3) and the input voltage available.

The screen of the control valve (V4) is connected to the earth potential, the cathode to a constant negative voltage determined by V5 and V6 and the control grid to a potentiometer network connected between the positive stabilized and negative lines. In this way V_{g2-k} and V_k are held constant and V_{g1-k} is negative. R may then be varied until the stabilized output is finally reduced to zero, but V_{a-k} is sufficient to allow the valve to operate.

The negative voltage may be obtained by the method shown in Fig. 17.16(b), by means of a separate winding, or separate mains transformer and rectifier. Another method is shown in **Fig. 17.16(c)**, where a rectifier is connected in the reverse direction across one section of T1. Where a limited range of output or a fixed output is required the negative supply is not necessary, and the basic circuit as in Fig. 17.16(a) may be used.

When a negative supply is not used, it is common practice to derive the screen voltage of the control valve from the unstabilized supply. Normally an increase in input voltage leads to an increase in output voltage which increases the anode current of the control valve, thus increasing the bias on the series valve. The resulting change in the impedance of the series valve reduces the output voltage, but if at the same time V_{g2} of the control valve increases, its anode current will increase still further and the mains variation will be over-compensated. If this condition occurs, an improvement in stabilization can be obtained if R2 in Fig. 17.16(a) is replaced by a reference tube, keeping V_{g2} constant, or if the screen supply is taken from the stabilized voltage if this is not variable. This latter arrangement is shown in **Fig. 17.16(d)**.

R_L should be as large as possible for maximum gain. A limit is reached when the anode-cathode voltage of V2 becomes too small. One method of keeping R_L as large as possible is by the use of a reference tube in the anode circuit of the control amplifier (V4 in **Fig. 17.16(e)**). In this manner, V_a can be high even when V_{out} is low and R_L is large. When the output voltage covers a wide range, care must be taken to choose a stabilizer tube (V4) that is capable of handling the range of current.

Factors which Affect the Design of Stabilizer Circuits

(a) The unstabilized voltage V_{in} will vary as the load current is varied, due to the internal resistance of the rectifier unit. Therefore the regulation of the circuit prior to the series valve must be known and allowed for. A linear regulation of 17 per cent may be assumed.

(b) The input voltage V_{in} will also vary with the variation of mains voltage, which can fluctuate as much as ± 7 per cent. In practice it is found that ± 7 per cent is exceptional and that $+4$ per cent and -7 per cent are more realistic. It is assumed that the valve characteristics will not be affected by a change of $+4$ per cent or -7 per cent of heater voltage.

(c) To obtain the *maximum* stabilized output for any given input, the potential drop across the series valve will be necessarily low. For low voltage drop, V_{g-k} of the series valve is low but in order to avoid grid current flowing, a minimum value of -1V is advised.

(d) To obtain the *minimum* stabilized output voltage for any given input, the potential drop across the series valve will be high, thus the anode dissipation will be high. This must be kept within the valve rating, and may necessitate the use of several valves in parallel.

(e) The total current taken from a series valve may be increased if the input and output voltage is reduced. Maximum anode dissipation and maximum cathode current ratings must not be exceeded.

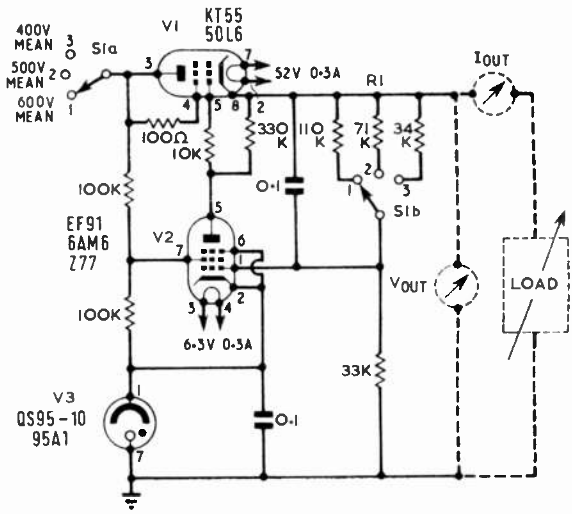


Fig. 17.18. Stabilizer to follow suitable power supply as indicated in text. S1b should be connected to the junction of V2 grid and the 33k ohm resistor via a 0 40 k ohm variable resistance RV1.

200-500V Stabilizer

The circuit of **Fig. 17.18** is that of a stabilizer which will deliver 100 mA with good regulation over a wide range of input and output voltages. A variation of stabilized output voltage from 200-500 volts is possible if an input of 400-600 volts is available. If a greater output current is required, a smaller output voltage per switched step must be accepted or further series valves placed in parallel with V1.

Should extreme stability be required a voltage reference tube QS82/3 or equivalent should be used for V3 in place of the type shown.

Table 17.2 gives operating parameters for the three positions of switch S1 for the circuit shown. The value of $+21$ per cent quoted for regulation allows for a 4 per cent rise in mains voltage and a 17 per cent rise of input due to the power supply unit regulation.

TABLE 17.2

	Position of Switch S1			
	1	2	3	
I_{out}	100 0	100 0	100 0	mA
V_{in} (mean)	600 700	500 600	400 500	V
V_{in} (+4% mains)	624 728	520 624	416 520	V
V_{in} (−7% mains)	560 660	465 560	372 465	V
Variation of V_{out}	390–500	295–400	195–310	V
R1	110	71	34	kΩ
RV1	0–40	0–40	0 40	kΩ
Stabilization between 0–100 mA	10	10	10	mV
Regulation between +21% and −7% of V_{in}	1	1	1	V

GRID BIAS SUPPLIES

Operating and protective grid bias for class C and modulator stages is conveniently obtained from a small power supply shunted with a heavy bleeder resistor to minimize the variation of output voltage when grid current flows. The standing current through the bleeder resistor should be at least five times, and preferably ten times, the total grid current flowing. One or more variable tapping points on the bleeder resistor will enable the bias voltage to be adjusted to any desired value—**Fig. 17.19(a).**

A gas filled stabilizer and limiting resistor may be substituted for the bleeder resistor to improve the regulation of the bias supply—see **Fig. 17.19(b).** The series resistor should be adjusted so that the stabilizer draws its rated minimum current when no grid current is present. The flow of grid current will increase the current through the stabilizer and the total grid current of all stages connected to the supply should be limited to a value not greater than the maximum permissible stabilizer current. When used in conjunction with class C amplifiers the output voltage of the supply is adjusted to provide bias beyond the point of cut-off, the difference between operating bias and cut-off bias being provided by the flow of grid current through the appropriate value of grid leak.

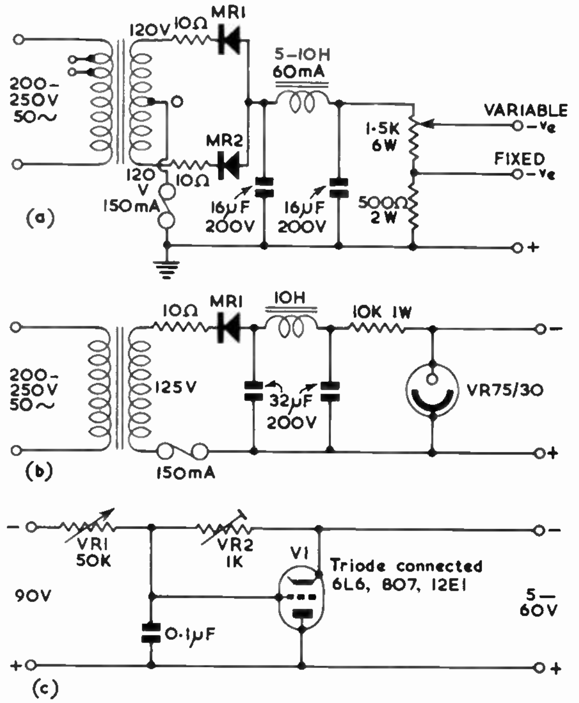


Fig. 17.19. Typical bias supply circuits. (a) Supply providing a fixed bias voltage of —30 volts and a variable bias of —30 to —120 volts. Grid current should not exceed 10-15mA with this circuit if the regulation is important. MR_1, MR_2, may be two type RM1, OA210 or equivalent. (b) 75 volt grid bias supply with VR75/30 stabilizer to improve regulation. Grid current of up to 30-35mA may be drawn. MR_1 may be type RM1, OA210 or equivalent. (c) Simple shunt stabilizing circuit providing variable regulated bias voltage. Grid current of up to 10-15mA may be drawn. The output voltage is held within 5 volts over the range 30-60 volts.

Fig. 17.19(c) shows a simple shunt stabilizer circuit which may be used to hold the grid bias voltage constant within a few volts with quite large variations of grid current. The circuit functions as a cathode follower and has a low output impedance of the order of 100 ohms. The output voltage is set by VR_1 and VR_2 is used to set the minimum grid bias on V_1. A parasitic suppressor may be necessary at the grid of V_1 as the circuit may oscillate at v.h.f. with low values of grid bias on V_1.

Two simple circuits for obtaining low values of bias voltage from the heater line are shown in **Fig. 17.20(a)** and **(b)**. The 6AL5, EB91 or other miniature double diodes may be used if desired. The circuit of Fig. 20(a) is useful in receivers to supply bias for gain control in r.f., i.f. and a.f. stages and the voltage doubling circuit of Fig. 17.20(b) for a.f. output stages which do not draw grid current.

CONSTRUCTION OF POWER SUPPLIES

The layout and construction of power supplies is by no means critical but there are certain design features which if adhered to will assist materially in producing a reliable and safe unit.

The conventional four-sided chassis with or without a front panel has the merit of being readily made up or purchased, and in conjunction with through-chassis mounting components produces a layout where the greater part of the high voltage and mains wiring may be underneath the chassis. A perforated metal cover overall will ensure that exposed high voltage terminals are out of immediate reach.

Input and output connections are preferably made with shielded plugs and sockets, male chassis connectors being used for mains input and female chassis connectors for output. The rear side member of the chassis is convenient for mounting these components. Three pole mains input connectors should always be used when the power supply is built as a separate unit, the third pole of the connector being taken to chassis so that the metalwork is earthed via the three core mains cable. When the power unit chassis forms part of a rack or cabinet of equipment, each chassis should be bonded together independently of any other connections and earthed, either to the mains earth as above or to a separate properly constructed earth connection. A water pipe must not be used because it is now common practice to use underground plastic pipe.

Double-pole mains switches should be used and wired in next to the mains input so that the unit may be isolated completely when changing fuses and making circuit adjustments. When the power unit chassis is to be enclosed in a cabinet or rack, adequate ventilation of heat producing components is essential. Valve rectifiers should be mounted towards the rear edge of the chassis to allow free circulation of air. Semiconductor rectifiers with fins must always be mounted with the fins vertical and preferably over a rectangular hole in the chassis. The more recently introduced

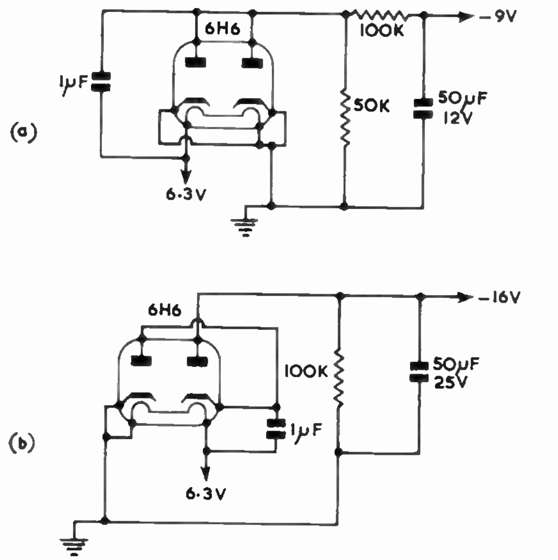


Fig. 17.20. Low voltage bias supplies deriving their input voltages from the heater line in the equipment. These circuits are not suitable for use when appreciable grid current is to be drawn.

wire ended silicon rectifiers are conveniently wired to tag strips and interconnections made between the tags to produce bridge, biphase or other circuits as required.

Paper capacitors are preferable to electrolytic types where reliability is of primary importance but tend to be expensive when high values and working voltages are called for. It is possible to use electrolytic capacitors by connecting two units of similar capacitance and working voltage in series to produce a capacitor of twice the working voltage and half the capacitance of one unit provided that voltage equalizing resistors of 50–100K ohms are connected across each capacitor. The outer can of the capacitor which is at high potential will be at a potential of several hundred volts above chassis and should therefore be adequately insulated.

When multiple unit electrolytic capacitors are used it should be noted that one unit is often specially processed for use as a reservoir capacitor and care should be taken to ensure that the correct unit is used in this position. Plain foil electrolytics are recommended rather than etched foil types as although somewhat larger physically, they provide a greater factor of safety in respect of working voltage and ripple current. It should not be assumed that the outer case of metal cased electrolytics is isolated from the electrodes and when used in grid bias supplies or other circuits where the negative terminal is above chassis potential, the metal case should be insulated by means of p.v.c. or polythene sheet wrapped around it, if this is not already provided by the manufacturer.

A bleeder resistor should always be used across the output of a power supply as it provides a means of discharging the capacitors when the equipment is switched off. The resistor used should be of adequate wattage rating, as an open-circuit bleeder is a hazard to safety. The bleeder may form the series resistor for an h.t. voltmeter. A 5 mA meter movement may be rescaled for this purpose and will provide a useful indication apart from its safety function.

Fuses should not be omitted from power supplies as it is false economy to lose a rectifier or transformer as a result of omitting a fuse. The practical power supply circuits show the points where fuses are most usefully employed. Where large transformers having low resistance primary windings are employed it may be found that the primary fuse blows on switching on due to the high initial surge current; this may be eliminated by the use of a current sensitive resistor (Brimistor or Varistor) in series with the primary winding. The Brimistor or Varistor has a high initial resistance which falls rapidly as it warms up to a value of a few ohms and may be left permanently in circuit. Similarly a Brimistor may be connected in the h.t. winding centre-tap or between the rectifier and the reservoir capacitor to prevent the h.t. fuse blowing at switch-on. Anti-surge fuses are useful in cases where current surges cause trouble with normal type fuses.

Although the layout of components in a power supply unit is not critical electrolytic capacitors and semiconductors should be mounted well away from heat producing elements.

Adequate spacing of all components is necessary if reliability is to be maintained.

Dual Output Power Units

In many applications it is advantageous to obtain high and low voltage h.t. supplies from a single transformer.

Fig. 17.21 shows two general arrangements for this type of operation. The circuit of **Fig. 17.22** is based on that of the stabilizer shown in Fig. 17.16(b).

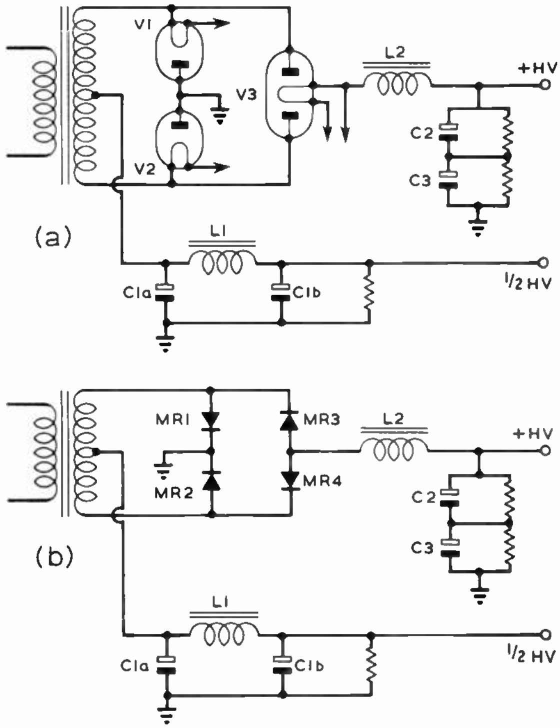


Fig. 17.21. (a) Utilizes both fullwave and bi-phase halfwave rectifiers. A disadvantage of this arrangement is that unless indirectly heated rectifiers are available, three separate filament windings or transformers are required. This however is readily overcome by use of semiconductor rectifiers as shown in (b). Semiconductor rectifiers must be protected against switching surges. Where several diodes are used for each rectifier voltage sharing is necessary across each diode.

POWER UNIT FOR USE WITH TRANSISTOR EQUIPMENT

General Specification:

Output voltage:	0-20V continuously variable in two overlapping ranges.
Output current:	0-1A throughout the voltage range.
Ripple voltage:	Less than 2 mV peak-to-peak at full load.
Overload trip range:	0-1·3A continuously adjustable.
Overload trip switching time:	Less than 30μsec for a short circuit overload.

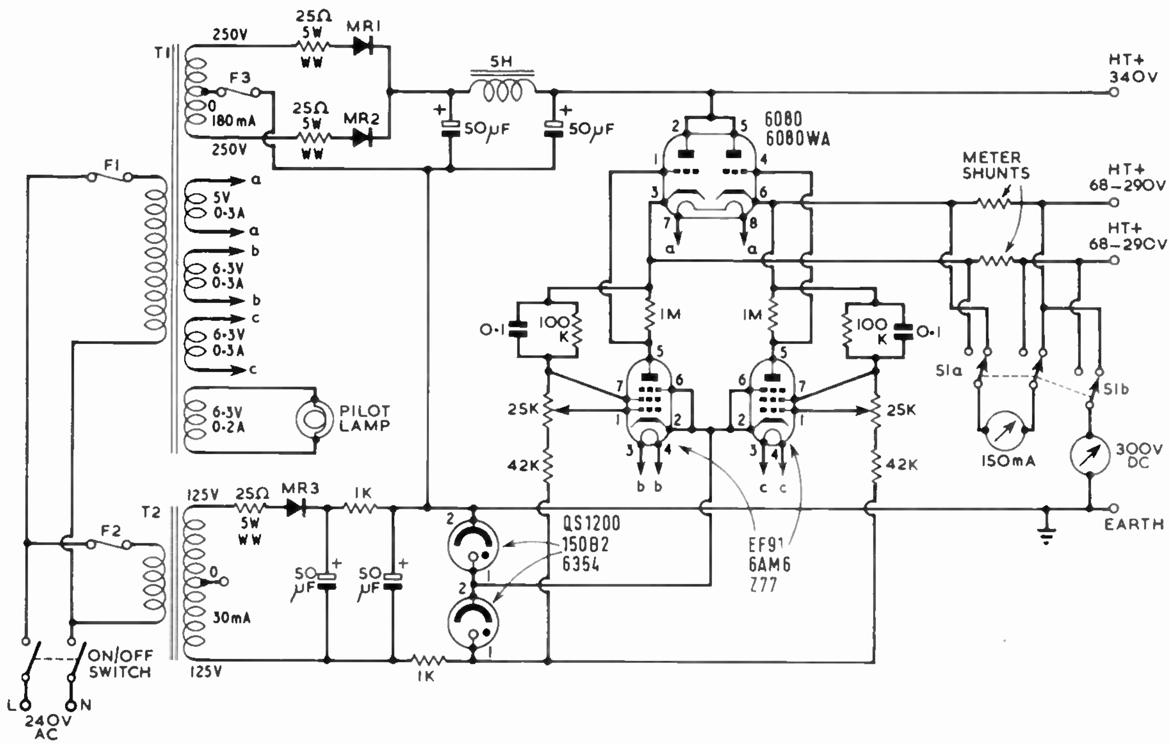


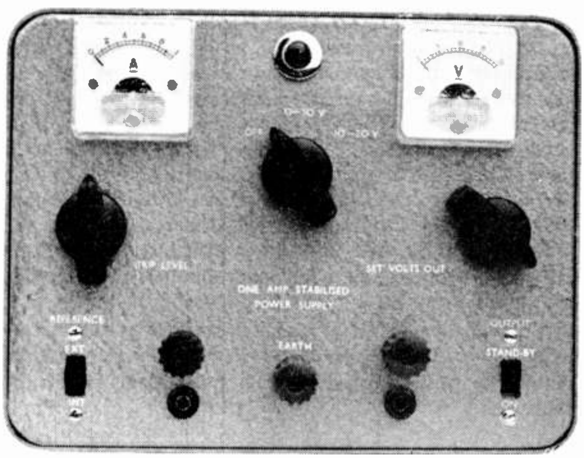
Fig. 17.22. Utility power supply having an output of 340V h.t. at 200 mA (unregulated) or two outputs simultaneously of 68-290V, stabilized. Rectifiers MR1, MR2 and MR3 should be suitable for 250V r.m.s.

Output impedance:	Less than 0·1 ohm under all conditions (0·02 ohm at 0·5A load).
Output change with 5 per cent mains voltage change:	0·6 per cent.
Mains current with unit	
(a) unloaded:	70 mA (10V range). 87 mA (20V range).
(b) fully loaded:	150 mA.
Operating temperature range:	20° C to 45° C ambient.
Weight:	7½ lb.
Suitable chassis size:	7 in. × 5½ in. × 2 in.
Suitable case size:	8 in. × 6 in. × 6 in.

Circuit Principles

Unsmoothed d.c. is produced by a bridge rectifier and is stabilized by a series emitter follower driven by a grounded emitter amplifier in a feedback circuit. Control of the output voltage is achieved by supplying the base of the amplifying stage from an adjustable point on a potential divider between the output terminal and a d.c. reference rail. The use of an auxiliary reference rail permits adjustment to be continued down to zero volts output, without loss of loop gain in the amplifying circuit. This reference rail is produced by use of an overwind on the mains transformer in conjunction with a half wave rectifier and series stabilization by an emitter follower. No feedback is used in this auxiliary supply, the emitter follower being driven by a Zener voltage reference diode.

By the use of an external reference facility, leads and plugs between the supply and a remote piece of apparatus may be included inside the feedback loop, with a considerable reduction in supply impedance.



Front panel view of the transistor power supply.

An overload trip facility is provided, the current sensing element being a one ohm resistor in series with one of the supply rails. The trip circuit comprises a Schmitt trigger circuit and amplifier which effectively short circuits the main feedback amplifier in the event of an overload, thereby reducing the output of the unit to zero.

Circuit Details

A suggested layout is given in **Fig. 17.23** and the complete circuit diagram in **Fig. 17.25.**

The mains transformer is of a readily available type, the taps on the secondary being such that two voltage ranges can be accommodated to minimize power dissipation in the emitter follower.

A d.c. voltage is built up across the main reservoir capacitor C3 by the bridge rectifier D1 to D4. BYZ10 diodes were used here in the prototype but there is no reason why BYZ13 diodes or a similar type should not be used.

The d.c. reference line makes use of the 30V overwind on the transformer, half wave rectification being used to charge the reservoir capacitor C1. The series emitter follower stabilizer is TR1, the base of this being held some 5·5V above the positive output rail by the Zener diode D6. The smoothing capacitor C2 reduces the ripple on the auxiliary line to an acceptable level (about 10 mV).

The main emitter follower stabilizer TR7 is driven by a second emitter follower TR6, R15 being inserted to prevent the output of the unit rising with no load applied, due to collector base leakage in TR7. The potential on the base of TR6 is controlled by the grounded emitter amplifier TR5, d.c. for this amplifier being derived from the reservoir capacitor through R13 and R14. Splitting the d.c. supply path permits a smoothing capacitor C5 to be inserted, this capacitor reducing the ripple in the supply to the small signal amplifying stage. A second important function of C5 is in the standby mode. When S3b is closed, C5 is shorted through a 10 ohms resistor and the base voltage of TR6 is reduced to near zero. Because of the forward base emitter voltages of TR6 and TR7, the output of the unit is exactly zero. Recovery from this condition after opening S3 is regulated by the time constant R13 C5. The slow potential build up caused in this way prevents premature firing of the overload trip circuit due to current surges in the output of the unit.

The emitter of the grounded emitter amplifier is returned to the voltage reference diode used in the auxiliary rail stabilization circuit. The base of the amplifier is connected to a point on a potential divider between the negative output reference and the auxiliary reference rail. Adjustment of RV2 changes the position of the tapping point on the divider and, since the base emitter potential of TR5 remains substantially constant, this results in a change in output voltage. In order to change voltage ranges, a fixed resistor, a little less than the maximum value of RV2, is switched into the divider. C6 improves the loop gain of the amplifier at high frequencies, the value of this loop gain at 1 kc/s being approximately 34dB.

The ammeter is placed inside the feedback loop so that the effect of its impedance is reduced by the amplifier, the error in reading due to the drain in the divider being negligible if a 1A instrument is used.

A small electrolytic capacitor C7 is placed across the output terminals of the unit, this capacitance having most effect at frequencies above those at which the feedback amplifier operates; i.e. above 10 kc/s.

The shunt resistor R2 across the main reservoir capacitor is designed to pass a greater current than is flowing in the auxiliary reference circuit. The effect of increasing load does modify the reference slightly and this may give rise to a negative d.c. output impedance. This negative impedance is not of sufficient magnitude to support oscillation on any practical load.

The Schmitt trigger circuit used in the current overload trip is TR2 and TR3. The emitter supply for this pair is derived from a Zener reference diode D7 in order that the performance should remain unaffected by a change in output range. In the normal state, TR3 is bottomed and TR2 cut off. As the load current increases, the base of TR3 is taken more negative until this transistor begins to cut off. As the collector voltage rises towards the reference rail, TR2 is

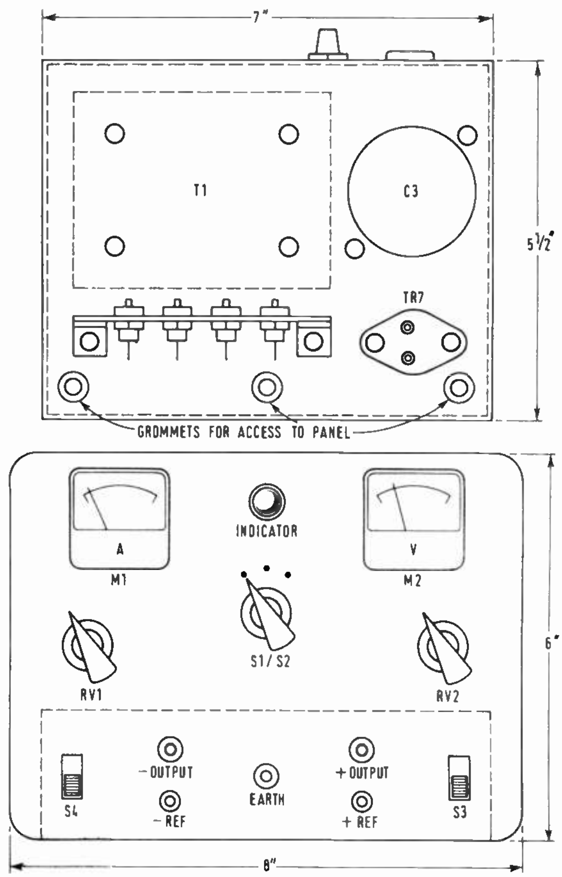


Fig. 17.23. A suitable panel and chassis layout for the power unit.

pulled on by the coupling effect of C4 and R8. Regeneration completes the switching cycle and TR2 becomes fully bottomed, the base of TR4 being pulled negative. TR4 starts to conduct, the collector current being sufficient to pull the base of TR6 to zero with TR5 cut off.

When the output has fallen to zero, the overload current will be zero, but the hysteresis voltage in the trigger circuit is sufficient to hold the unit off. The trigger can only be reset by closing S3a which turns TR2 off after a short time determined by C4 and the magnitude of the hysteresis voltage. This short interval is sufficient to allow C5 to discharge through R21 and S3b, thereby avoiding a pulse of output when selecting the standby mode. The current level at which the trip fires can be adjusted by RV1 in the base circuit of TR3.

If, for sake of simplicity, it was required to construct this unit without the current trip facility, all components associated with TR2, TR3 and TR4 could be omitted without affecting the performance of the unit in any other respect. It would, however, be advisable to put a fuse in the output circuit; a 2A fuse in series with the ammeter would afford sufficient protection to TR7 in most cases.

Construction

Suitable case and chassis sizes for this unit are shown in Fig. 17.23, these components being available as standard items.

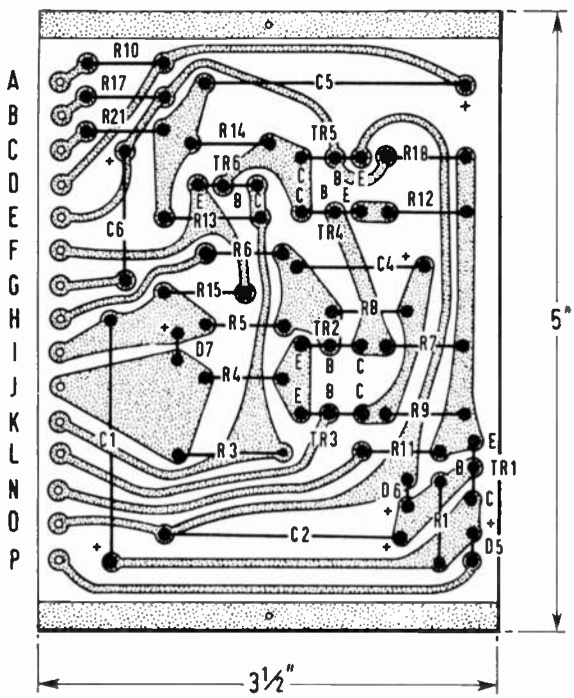


Fig. 17.24. A component layout suitable where printed circuit or pin-board construction is to be used.

The BYZ10 diodes need no conduction cooling and may be mounted on a paxolin strip or some other insulating material. The OC28 emitter follower needs an excellent heat sink if the unit is not to fail under low voltage, high current, conditions. If a chassis similar to that specified is used, the transistor may be bolted to this, a mica or anodized aluminium spacer being used to insulate the transistor from the chassis. If some other chassis form is to be used, it should be ensured that an insulated heat sink with a thermal resistance of less than 3°C/W is provided for the OC28.

The low power sections of this unit lend themselves to printed circuit or pin board construction. All the low power stages can be accommodated on a board 5 in. × 3½ in. which will fit underneath the main chassis. A suitable component layout for this part of the unit is shown in **Fig. 17.24**, although it is in no way critical so far as performance is concerned. Connections between the circuit board and other components are identified by letters A to P on the circuit diagram (Fig. 17.25).

Testing

When the unit has been checked thoroughly for any mechanical faults, it may be tested as follows. Turn RV1 to its maximum position (slider nearest auxiliary rail); RV2 to minimum (shorted); S4 to REFERENCE INTERNAL; S3 to NORMAL and S2 to 0-10V d.c. Switch on the mains supply and check that, on increasing the setting of RV2, an output voltage reading is obtained. If no such output is forthcoming, measuring the voltages across C3 and C1 should give an indication of whether the fault is in the supply circuit or stabilizing circuit. To eliminate the influence of a faulty current trip circuit, it may also be found useful to disconnect the collector connection of TR4.

After a voltage has been obtained on the output terminals, increase RV2 to maximum and check that the output is at, or above, 10V. Now change to the 10V to 20V range and check that RV2 has its expected range.

Return S2 to the 10V range and set the output voltage to 6V d.c. Now reduce the trip level potentiometer slowly to zero. If the output falls to zero, the potentiometer should be very near the low end of its range when the trip fires. Increase RV1 to maximum again and, if in the last test the overload trip was fired, reset the trip by switching S3 momentarily to STANDBY. The output level should be observed to climb slowly back to 6V as C5 charges through R13.

Connect a load of about 10 ohms to the output of the unit and adjust the output voltage until the load current is 0·6A. Now reduce the setting of RV1 until the overload trip fires. RV1 should be near mid range at this time. Reset the trip and RV1 and increase the output to 1A to ensure that the maximum setting of the trip level is sufficient. Any slight discrepancies in the range of RV1 can be removed by adjusting R10 and R11 until the correct range is obtained. Only after the current trip range has been fully checked should any short circuit test be attempted as, if the current trip is faulty, an overload will destroy TR7.

Should any difficulty be experienced in obtaining the expected performance from the unit, the voltages shown in **Table 17.4** may be used as a guide in fault finding. All voltages were measured with respect to the positive output terminal with an Avometer Model 8 on its 25V d.c. range.

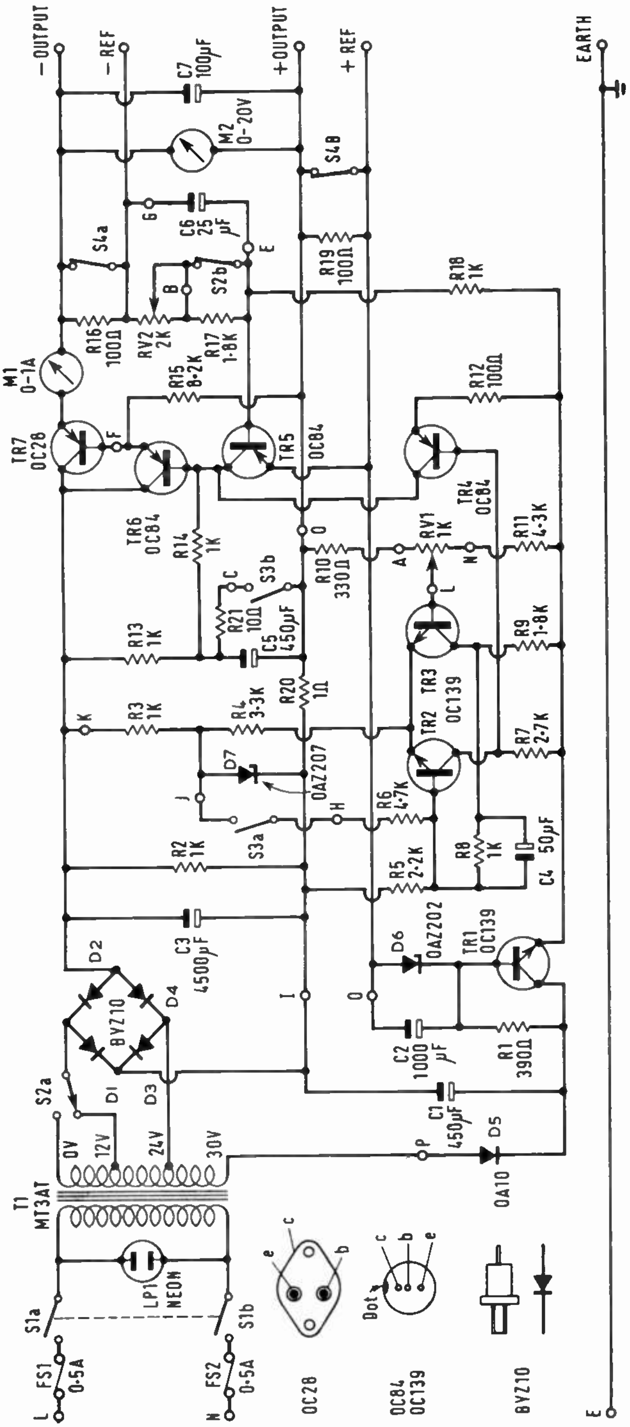


TABLE 17.3

Parts list for the stabilized power supply

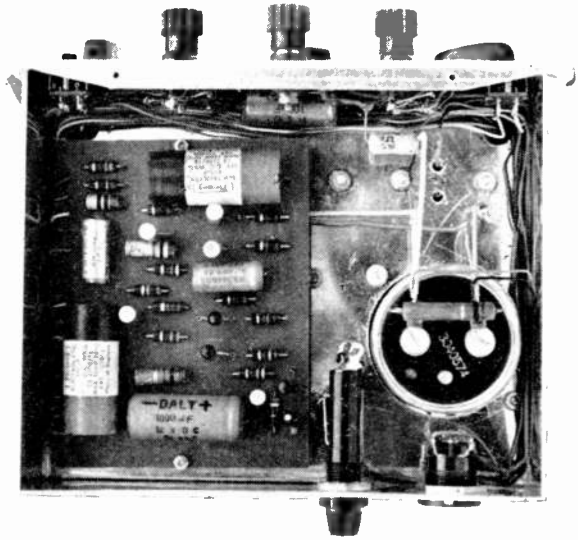
C1,	450μF	25V d.c.
C2,	1000μF	12V d.c.
C3,	4500μF	40V d.c.
C4,	50μF	12V d.c.
C5,	450μF	35V d.c.
C6,	25μF	25V d.c.
C7,	100μF	25V d.c.
F1, F2,	½A anti-surge fuse	
M1,	0-1A d.c.	
M2,	0-20V d.c.	
R1,	390	ohms ½W.
R2,	1K	ohms 1W.
R3,	1K	ohms 1W.
R4,	3·3K	ohms ½W.
R5,	2·2K	ohms ½W.
R6,	4·7K	ohms ½W.
R7,	2·7K	ohms ½W.
R8,	1K	ohms ½W.
R9,	1·8K	ohms ½W.
R10,	330	ohms ½W.
R11,	4·3K	ohms ½W.
R12,	100K	ohms ½W.
R13,	1K	ohms 1W.
R14,	1K	ohms ½W.
R15,	8·2K	ohms ½W.
R16,	100	ohms ½W.
R17,	1·8K	ohms ½W.
R18,	1K	ohms ½W.
R19,	100	ohms ½W.
R20,	1	ohms 3W.
R21,	10	ohms ½W.
RV1,	1K	ohms W.W.
RV2,	2K	ohms W.W.
S1,	2 pole on/off mains switch.	
S2,	2 pole 2 way rotary switch. (May be combined with S1).	
S3, S4,	2 pole 2 way slide switch.	
T1	30V 2A Mains transformer. Douglas type MT3AT.	

All resistors are ± 10% types.

The capacitance and voltage working levels of electrolytic capacitors may be safely increased if supply is difficult.

D1—D4,	BYZ 10
D5,	OA 10
D6,	OAZ 202
D7,	OAZ 207
TR1, TR2, TR3,	OC 139
TR4, TR5, TR6,	OC 84
TR7,	OC 28

Fig. 17.25. The circuit of the 20V 1A stabilized power supply. S2, the voltage range selector, is shown in the 0-10V position. If desired, the switch can be combined with the power switch S1. S3 is in the NORMAL position, and S4, the external reference switch, is in the INTERNAL position.



Underchassis view of the power supply, showing the positions of the printed circuit board and C3.

Operation

Operation of the unit is generally straightforward but a few notes may help to explain the facilities provided to their best advantage.

Current trip. This should generally be set at about twice the anticipated peak load current, though if only short circuit faults are expected, it may be left at its maximum setting.

Voltage Range. At the overlap point between ranges, it is generally better to use the low range than the high, since a great deal of power is wasted in the OC28 if the high range is used near its minimum voltage setting.

Standby Switch. The standby switch is operated momentarily to reset the current trip and may also be used while making adjustments to an experimental circuit under test. When returning to normal. the output will build up slowly to prevent the overload trip firing due to a current surge into a capacitative load. This slow build up may also give enough time to spot any fault in a circuit under test before any damage is done.

External Reference Switch. The external reference facility may be used in situations where a low source impedance from the power supply is vital to prevent oscillation or some other unfortunate condition. This is often the case where the unit being powered is separated from the supply by a long supply lead and, perhaps, also some plugs and sockets which all add to the supply impedance.

In this case, a pair of leads should be brought back to the power supply from the supply rails in the unit under test; these leads being inserted in the appropriate reference sockets on the power supply. If S4 is now opened, the feedback loop will extend to include the supply leads and hence their effective impedance will be reduced by a factor of up to 50 at frequencies over which the feedback amplifier operates.

TABLE 17.4

	e	b	c
TR1	−5·4	+5·6	−9·4
TR2	−0·7	−0·5	+5·3
TR3	+0·7	−0·8	+0·7
TR4	+5·4	+5·3	−10·5
TR5	0	+0·1	−10·6
TR6	−10·3	−10·6	−12·9
TR7	−10·3	−10·3	−12·9

Specimen d.c. voltages present in the power supply while it is supplying 10V at 0·5A. All voltages were measured with an Avometer Model 8 on the 25V d.c. range and are with respect to the positive rail.

Simple Power Units for Transistor Equipments

Figs. 17.26 and 17.27 give the circuits of two easily-constructed power supplies, operating from a.c. mains, which provide low-voltage, regulated outputs suitable for use with transistor equipments.

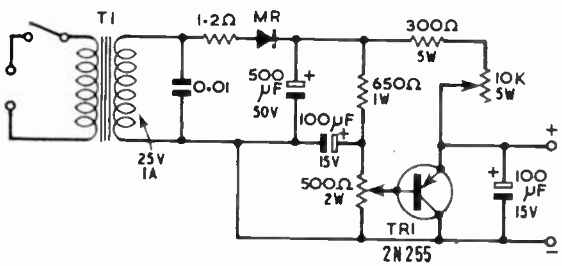


Fig. 17.26. A current-regulated transistor power supply unit which gives an output of up to 15 volts at load currents up to 100 mA. MR silicon rectifier suitable for 25V r.m.s. working.

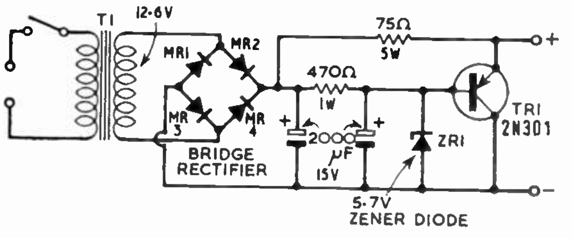


Fig. 17.27. A shunt-regulated transistor supply giving an output of 6 volts at load currents up to 200 mA. MR1-MR4, silicon bridge rectifier suitable for 12·6V r.m.s. working.

CHAPTER 18

INTERFERENCE

IN addition to the carrier frequency and the sidebands necessary to convey intelligence in the form of the Morse code, speech or RTTY signals, a transmitter may radiate unnecessary frequencies due to key-clicks, distortion of the modulation, harmonics of the carrier frequency or parasitic oscillations. Any of these radiated frequencies, necessary or otherwise, may cause interference in various ways:

(*a*) to reception of other amateur transmissions,
(*b*) to broadcast television reception (TVI),
(*c*) to sound broadcast reception (BCI),
(*d*) to protected services,
(*e*) to other services.

Interference to other services is relatively unimportant only because it is rarely encountered, but when such services operate in a shared band the amateur operator, to comply with the terms of his licence, may have to close down or change frequency or temporarily cease transmitting at the request of the other station.

Official reports show that radio amateurs are responsible for only a very small fraction of the total interference to television and broadcast reception. In some cases in which amateurs are involved the interference is due to bad design, maladjustment or careless operation of the transmitter. In other instances interference can be traced to poor installation or inadequate design of the domestic receiver. Occasionally it may be found that a " perfect " transmitter interferes with a " perfect " receiver and in such instances the trouble may be attributed to external cross-modulation effects or the radiation of harmonics produced by contact rectification in metal objects in the immediate vicinity. Lastly, a swamp signal on the fundamental frequency due to close proximity may blanket the domestic installation.

TRANSMITTER CONSIDERATIONS

In general it must be expected that telephony transmissions are more likely to cause interference than telegraphy transmissions and that telephony interference whenever it does occur is likely to be the more troublesome. This does not mean that the problems of preventing a c.w. transmitter from causing unnecessary interference are trivial but that the telephony operator should understand that he is trying to meet a more stringent set of requirements. In this chapter, the special problems relating to telephony and telegraphy will be dealt with first, followed by consideration of the problems common to both, such as the prevention of parasitic oscillation and the elimination of harmonics.

Telephony

In view of the congested state of the radio frequency spectrum the International Amateur Radio Union (Region 1) has recommended that, irrespective of the modulation system in use, the maximum modulation frequency should be 4 kc/s, while the International Telecommunications Union supports 3 kc/s. This upper frequency limit is a reasonable compromise between bandwidth and speech quality. Since most audio frequency amplifiers have a frequency range exceeding the IARU recommendation positive steps should be taken to restrict the bandwidth by incorporating some form of low pass filter. Resistance-capacity networks are worthwhile but much more effective filtering can be obtained by using the inductance-capacity type of filter.

Low pass filters can be installed in the low-level stages of the modulator although if this arrangement is adopted any harmonic distortion occurring in the modulator output stage will be unaffected by the filter and thus the transmitted bandwidth will be greater than desirable. It is therefore strongly recommended that the low pass filter be fitted between the modulator and the modulated radio frequency amplifier.

Even when the frequency response of the modulator has been restricted, the transmitted bandwidth may be much greater than necessary because of bad design or incorrect adjustment of the r.f. section of the transmitter. The major troubles are due to a shifting carrier frequency, nonlinearity (distortion) and over-modulation.

A carrier frequency which changes when the transmitter is amplitude modulated can cause a most annoying form of interference to stations working on adjacent channels. Frequency shift is most commonly due to bad power supply regulation, but may also be caused by the feedback of r.f. energy from the modulated stage into the oscillator. This can occur as a result of poor screening, lack of neutralization, or bad layout, and the effect is most marked on the lower frequency bands where the oscillator and the power amplifier are often operated on the same frequency (i.e. when there are no frequency multiplying stages). In severe cases the carrier may shift several kilocycles from the no-signal frequency and the receiver then has to be tuned to the setting where the carrier happens to be during actual speech transmission thus giving the appearance of asymmetrical tuning. In milder cases the tuning may appear to be normal but any heterodyne interference which may be present cannot be properly rejected at the receiver because the heterodyne note is not constant in frequency.

Appreciable non-linearity is inherent in all efficiency-modulation systems especially when a modulation depth approaching 100 per cent is attempted. Anode modulation, which is relatively distortion-free and does not require critical adjustment, is therefore recommended whenever interference is likely to be a problem.

Where the anode modulation system is used for modulating a tetrode or a pentode, it is essential to modulate the screen as well as the anode, applying only the proportion needed. Failure to do this will result in severe distortion, probably indicated by so-called downward modulation and the consequent radiation of "splatter". Due to the poor regulation of the screen supply it is possible to obtain adequate screen modulation if the screen resistor is fed from the unmodulated supply to the anode. Another common source of trouble is the use of too large a screen bypass capacitance in which case the higher modulation frequencies are prevented from modulating the screen grid. This results in downward modulation of high notes although when a constant tone of low pitch is directed into the microphone the modulation may appear to be quite normal. A somewhat similar effect can be caused when the cathode bias resistor of the modulated r.f. stage is inadequately bypassed for audio frequencies. In this case, however, it is the low frequencies which may be distorted.

Over-modulation is probably the most common fault in telephony operation and the resulting interference can be very severe, often being audible on frequencies more than 100 kc/s away on either side of the carrier frequency. The depth of modulation is checked preferably by using a cathode ray tube monitor but, failing this, the p.a. anode current will serve as a useful guide: during a transmission that is free from splatter it should not vary by more than about 5 per cent. Most cases of over-modulation are due to carelessness, and some form of automatic limiting device such as volume compression or speech clipping is desirable. When speech clipping is used it is essential to install a good low-pass filter after the clipping circuit, preferably between the modulator and the modulated stage. Examples of suitable circuits and filters are given in Chapter 9 (*Modulation Systems*).

Narrow-band frequency modulation is an alternative system which is sometimes used with great success where all attempts to eliminate TVI and BCI appear to fail. The principles and application are also described in Chapter 9.

Telegraphy

Although it may appear that interference problems arising from the use of amplitude modulation could be altogether avoided by reverting to c.w. telegraphy, it should not be forgotten that keying is in itself a form of modulation and many of the difficulties encountered in the use of telephony may still have to be faced. In the conventional method of keying a transmitter the carrier is modulated 100 per cent, and the rate at which the carrier level rises and falls during the keying determines the character of the keying sidebands and therefore the amount of interference likely to be caused to other stations. If the keying arrangement is such as to give a " soft " characteristic the carrier level does not change very rapidly and the sidebands may be negligible. On the other hand a " hard " keying characteristic means that the changes in carrier level are quite abrupt and the resultant sidebands may spread to frequencies far removed from the carrier frequency. **Fig. 18.1** gives an impression of how the interference due to key-clicks depends on the keying characteristics.

Another aspect of interference from a telegraphy transmitter is the variation of carrier frequency generally known as *chirp*, which although being confined within perhaps 200-300 cycles of the nominal carrier frequency, may have disastrous effects on a transmission occupying an adjacent

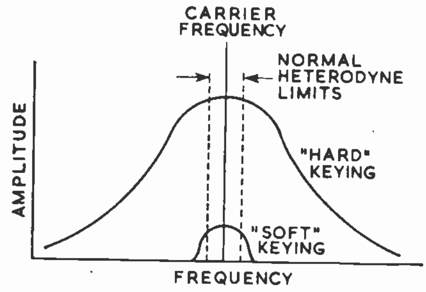


Fig. 18.1. The sidebands produced by key-clicks from a badly adjusted transmitter may extend over a wide frequency range on either side of the carrier. With a " soft " keying characteristic the interference from these sidebands should be negligible.

channel. Chirp is analogous to the frequency shift which may occur in a telephony transmitter and often arises from the same causes. It is unlikely to occur in a well designed transmitter provided that there is a suitable buffer stage between the keyed stage and the v.f.o.

These questions are not examined in detail here as they are better considered in relation to the general subject of keying which is discussed in Chapter 8 (*Keying and Break-in*).

Parasitic Oscillations

Parasitic oscillations in radio transmitters can be responsible for distortion, low efficiency, and instability and can cause interference to other services. Such oscillations can be divided into three groups depending on whether their frequency is

(*a*) Much lower than that of the carrier,
(*b*) Approximately equal to the carrier frequency or
(*c*) Much higher than the carrier frequency.

Low frequency oscillations are usually recognised by the fact that the transmitter radiates a number of frequencies on each side of the carrier at a spacing of some tens or hundreds of kilocycles. This is because the carrier is modulated by the parasitic oscillation and the spurious frequencies are the sidebands spaced at intervals equal to the oscillation frequency. The most common cause of l.f. oscillations is the use of r.f. chokes in the anode and grid circuits of an r.f. amplifier such that, in conjunction with the capacities present, a tuned-anode/tuned-grid type of oscillator circuit is formed. The normal neutralizing circuits are quite ineffective at the parasitic frequency but the trouble may be cured either by tuning the grid choke to a higher frequency than that of the anode choke (in which case the t.a.t.g. circuit cannot oscillate) or, preferably, by avoiding the use of a grid choke altogether. In most tetrode power amplifiers a grid choke is quite unnecessary if relatively high values of grid resistor are used. The extra r.f. loss which may, in some circuits, result from the omission of the grid choke is not usually serious. If, in a telephony transmitter, the trouble cannot be attributed to r.f. chokes the fault is most likely due to oscillation at a supersonic frequency in the modulator.

A parasitic oscillation on or near the carrier frequency is caused by r.f. feedback in one of the r.f. amplifier stages (usually the power amplifier) and is due to lack of neutralization, poor screening or bad layout. The oscillation may not be present while the carrier remains unmodulated and may only occur when the drive is removed from the amplifier (e.g. when the driver stage is keyed). In these circumstances, when the key is up, an unstable carrier may be radiated in another part of or even outside the band in use. Unless the transmitter has been carefully checked for oscillations during its initial tests this state of affairs may persist for some time as it is difficult for a receiving operator to identify and associate an unstable spacer with any particular station. Sometimes these oscillations (together with l.f. and more especially h.f. oscillations) occur only in certain parts of the modulation envelope thus giving rise to disconnected " monkey chatter " on telephony and to violent key-clicks on telegraphy. Once it is determined that oscillations are present the cause is usually obvious. Overrunning and excessive h.t. are other causes of these difficulties, particularly when tetrodes and pentodes are used.

V.h.f. parasitic oscillations are almost always due to an unfortunate choice of layout and they may occur in r.f. amplifiers or in modulator stages. The oscillations, which are usually generated by a form of t.a.t.g. circuit, can be cured by fitting grid-stopper resistors. These may be shunted by v.h.f. chokes to prevent power loss and overheating at the carrier frequency in the case of r.f. stages and to avoid distortion in the case of a.f. stages which are driven into grid current. If it is impracticable to use grid stoppers in v.h.f. transmitters because of the consequent power loss, the layout must be changed with a view to making the parasite-generating circuits such that the anode circuit resonates at a lower frequency than the grid circuit. Sometimes a v.h.f. parasitic can be eliminated by connecting a capacitance of about 30 pF from grid to earth or from anode to earth in the suspected amplifier, although of course capacitances of this order are not permissible in r.f. stages operating on frequencies higher than about 28 Mc/s.

Testing a New Transmitter

An essential procedure in the testing of a new transmitter is the running of the power amplifier without drive and with the maximum possible safe anode current. Under these conditions the anode and grid tuning should be varied on all bands to ensure that no parasitic oscillations are being generated. The presence of parasitics may be indicated by sudden changes in the anode current when the tuning controls are varied, by the existence of grid current or by the lighting of a neon lamp held close to the anode circuit. A transmitter should never be connected to an aerial until it has satisfactorily passed this test. While testing, a suitable load or dummy aerial may be coupled to the output tank circuit.

Harmonics

Perhaps the most common form of interference caused by amateurs is that due to the radiation from the transmitter of harmonics which happen to fall in the wavebands allocated to the television service. Harmonic radiation also causes interference to amateurs working on higher frequency bands and to other services using frequencies which are not in the amateur bands but which are harmonically related to a lower frequency amateur band. The responsibility for curing this type of interference rests entirely with the amateur concerned, and hence the suppression of harmonics is a vitally important factor in the design of a transmitter.

In the majority of amateur transmitters the output stage runs in class C, and thus can be expected to produce strong harmonics which are isolated from the aerial by the coupling circuits. It is, therefore, desirable to drive the grid(s) of the output stage no harder than is necessary to maintain the linearity of a telephony stage. Excess drive not only results in greater harmonic output, but reduces the fundamental output and can damage the valve. Some method of controlling the drive power is most desirable and a convenient way of achieving this is to vary the screen voltage in the driver stage. For c.w. or f.m. telephony the reduction in efficiency when operating the final stage in class AB or B is small, although the harmonic output is much reduced.

It should not be assumed that only the final stage produces harmonics. A driver stage driven hard to produce required harmonics and with an unsuitable tank L/C ratio may produce more trouble than a well designed and lightly driven p.a. which follows it.

Obviously, if no harmonics of the basic oscillator fall in the unwanted region, trouble due to harmonic radiation cannot occur. For example, a v.f.o. or crystal oscillator at 14 Mc/s is unlikely to cause TVI in the Holme Moss area. The mixer v.f.o. principle using a low frequency v.f.o. and high frequency crystals, for example, can be used to produce the output frequency at a low level relatively free from spurious products, followed by linear amplification. The principle of running the basic oscillator at as high a frequency as possible is also valuable in v.h.f. transmitters.

Tank Circuit Design. In any r.f. amplifier the anode current waveform contains harmonics, due to the partial suppression of one half cycle. This effect is most marked in a class C stage but is negligible in class A. In order that only the component of anode current at the desired frequency shall generate appreciable voltage the anode circuit must be tuned, and the Q of the tank circuit determines the suppression of unwanted frequencies. A loaded Q of about 12 is a reasonable compromise between spurious suppression and circuit losses for a p.a. stage, but in the case of a frequency multiplier it may be advantageous to operate at a higher Q since efficiency is not of such prime importance. For detailed design procedure see Chapter 6 (*H.F. Transmitters*).

Even in class A buffer amplifiers, a slug tuned coil with sufficient parallel capacitance to reduce the response outside the desired band is to be preferred to the ubiquitous 2·5 mH choke with its multiple resonances. Better still, a wide band coupler may be used.

It is most important that the tank capacitance is wired to provide an effective bypass from anode to cathode at the harmonic frequency, i.e. with short stout leads to minimize stray inductance throughout the whole of the r.f. path. Inductance between cathode and earth, being common to the anode and grid circuits, can also cause unwanted feedback and thus is doubly undesirable.

Some tank circuit arrangements may prove rather ineffective in reducing the harmonic amplitude in the output unless suitable precautions are taken. A typical example is that of a push-pull output circuit in which the split-stator

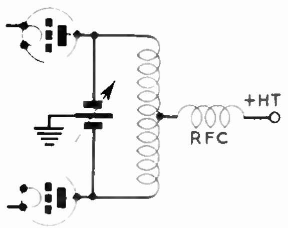


Fig. 18.2. A push-pull anode tank circuit of this form may be responsible for troublesome even-harmonic radiation.

tuning capacitor has its rotor connected to earth and the coil centre-tap connected to the h.t. line through an r.f. choke: see **Fig. 18.2.** This arrangement is normally used to provide a better circuit balance and a low impedance path to earth for harmonic frequencies. The even-harmonic frequencies present at the valve anodes are in phase with each other and may be transferred by unintentional capacitive coupling to the output feeder. Unwanted capacitive couplings can be minimized by making an earth connection to the coupling coil. Transmitters designed to produce a minimum amount of harmonic radiation should have an unbalanced coupling coil feeding into a coaxial output feeder since this arrangement simplifies the screening problem. The coil itself should always be coupled to the " earthy " part of the main tank coil.

In addition, the use of a *Faraday screen* between the tank coil and the output coupling coil is preferable, since this will serve to eliminate the capacitive coupling without affecting the inductive coupling. This is desirable because the higher-order harmonics which are to be suppressed will be more readily passed by any stray capacitive coupling, and this point should be borne in mind when determining the layout. The Faraday screen, shown in its most simple

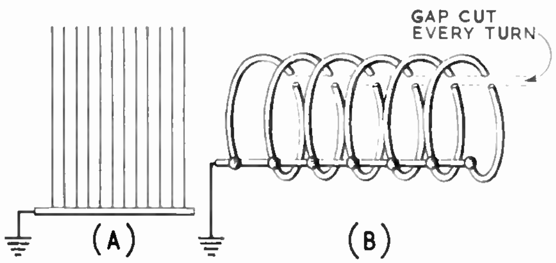


Fig. 18.3. Faraday screens. The wires must be insulated from each other to prevent the formation of closed loops which would act as inductive shields. The flat screen (A) can be used between coils which are arranged end-to-end. The cylindrical screen (B) is intended for use with concentric coils: it can be made by winding a single-layer coil of suitable diameter, connecting all the turns together along one side and cutting a gap along the other side.

form in **Fig. 18.3 (A)**, consists of a number of closely spaced insulated parallel wires joined together at one end and connected to earth (or any other convenient fixed potential). For use between concentric coils it can be given a cylindrical shape, as shown in **Fig. 18.3 (B)**. Where the output coupling coil consists of merely one or two turns the principle of the shielded loop illustrated in **Fig. 18.4** is often used.

Pi-networks. If the p.a. stage is of the single-ended switched-band type the most satisfactory output circuit is that using a pi-network: see **Fig. 18.5.** When used correctly this circuit gives good discrimination against harmonics owing to the frequency-dependent action of the potential divider formed by L1 and C2. The improvement compared with a simple tuned circuit is a factor of 4 for the second harmonic, 9 for the third harmonic and n^2 for the nth harmonic. To obtain the best results from the circuit it should be properly designed to transform the transmitter load impedance (usually less than 100 ohms) to a value suitable for correct operation of the p.a. valve. While the pi-network tank circuit is better than a simple tuned circuit in attenuating frequencies *higher* than that to which it is tuned it is less effective in attenuating *lower* frequencies. A power amplifier using a pi-network circuit should therefore be driven by a buffer amplifier rather than by a frequency multiplier,

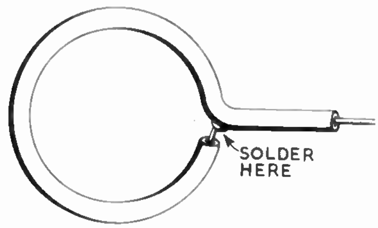


Fig. 18.4. Shielded link coil made from coaxial cable. The outer braiding is left in position but does not form a closed loop. The inner conductor is soldered to the outer braiding at the point shown. The construction is not suitable for coils of more than a few turns.

since in the latter case strong low-frequency signals may be radiated. For this reason the pi-network should be connected to the aerial via a suitable aerial tuning unit. If there is a standing wave on the output coaxial cable it may be found that the insertion of a low pass filter, which is equivalent to an extra length of cable, transforms the load impedance to a value outside the range of the pi tank, and the transmitter will not load correctly. Even a moderate standing wave ratio can lead to trouble, but this can be completely cured by the insertion of an aerial tuning unit.

Harmonic Traps and Filters. In spite of the use of a correctly designed tank circuit and coupling system, it is sometimes found that the harmonic radiation is still sufficient to interfere with local television receivers. In this event, a series-tuned trap adjusted to resonate at the offending harmonic frequency should be connected across the transmitter output socket as shown at L2C4 in Fig. 18.5. This trap circuit should be tuned for minimum output at the harmonic frequency, using an insulated trimming tool inserted through a hole in the transmitter screening. When the harmonic to be rejected lies in the television band the component values will be such as to leave the fundamental operation of the transmitter unaffected while the trimmer is being adjusted.

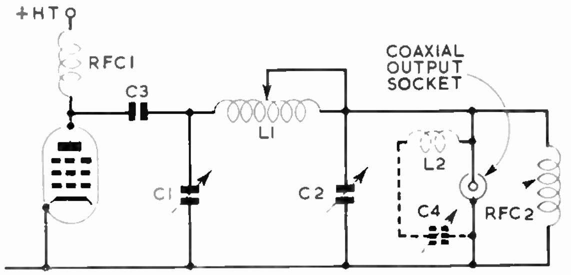


Fig. 18.5. Pi-network output circuit.
C1, 250 pF tuning capacitor; C2, 0·0015 μF loading capacitor; C3, 0·001 μF blocking capacitor; C4, 50 pF trimmer; L1, tapped inductance; L2 to resonate with C4 for reducing a specific harmonic; RFC1 anode choke; RFC2, safety choke. The choke RFC2 serves to prevent the appearance of high voltage from the h.t. supply on the output feeder cable in the event of a failure of the blocking condenser C3. It is not essential if the cable has a low-resistance d.c. path to earth at the far end.

One use of parallel-tuned circuits as harmonic traps in the anode lead of the p.a. valve is not recommended since they may give rise to parasitic oscillations. Tuned traps may need readjustment whenever the transmitter frequency is varied, and for this purpose it is helpful to incorporate a harmonic-indicating device in the transmitter.

Further attenuation of harmonics may be achieved by fitting a low pass filter between the transmitter and aerial coupler. To avoid interference to television reception such a filter would normally pass all frequencies up to 30 Mc/s and have a high attenuation at higher frequencies. Examples of suitable filters are given later in this chapter, and here it is only necessary to stress that if optimum performance is to be obtained the filter must be designed for, and terminated by, the correct load impedance. In general, the proper adjustment of the aerial coupler can only be achieved by using a v.s.w.r. bridge (see Chapter 19—*Measurements*). Harmonic radiation may also be caused by s.w.r. bridges (due to the rectifying action of the diodes) or by a high v.s.w.r.

It is sometimes found that a low pass filter does not reduce the harmonic radiation; on the contrary the interference may be worse. This condition can occur when harmonic radiation takes place directly from the transmitter wiring. Such radiation may be reduced by operating all the frequency-multiplying stages at the lowest possible power level, but the most reliable cure is to screen every part of the transmitter.

Screening

The correct use of screening plays a very important part in the reduction of TVI but unless the principles involved are fully understood the results may be disappointing. The basic principle can be explained with the aid of the simplified arrangement shown in **Fig. 18.6 (A)**. A transmitter *T* is enclosed with its load resistance *R* in a perfect screening box. Because the screening is perfect there will be no detectable radiation outside. For the sake of simplicity, it is assumed that the transmitter derives its power from batteries also contained within the box so that there can be no question of radiation from any power supply lines. The problem is to allow the transmitter to radiate its energy outside the box but only at the proper frequency: any other frequencies such as those which could cause TVI must be prevented from leaving the box. **Fig. 18.6 (B)** shows how this can be achieved. Here the transmitter is connected to an external load by screened cable (with its associated plugs and sockets) and a screened lowpass filter. The load receives r.f. power from the transmitter at the wanted frequency while the filter stops the transmission of any r.f. energy whose frequency may lie within the television bands. It is important to note that the screening must be continuous to ensure that the only way out of the box is through the filter.

Although the arrangement of Fig. 18.6 (B) enables a carrier to be radiated with minimum harmonic content there still remain the practical problems of making connections to a modulator or a keying circuit and also of connecting a mains supply to feed a power unit instead of the internal battery. These problems are really one, i.e. that of taking connections into or out of the box and at the same time preventing r.f. energy from escaping out of the box along the lead. To do this it must be made impossible for any r.f. potential difference to exist between the leads and the outside of the box.

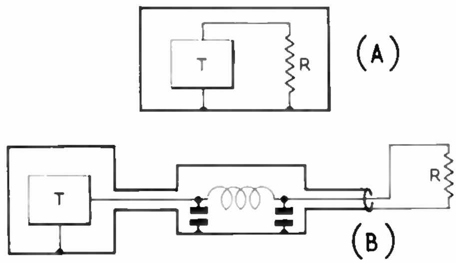


Fig. 18.6. The perfect shielding shown at A prevents any radiation from the transmitter or its load. In the arrangement B a low-pass filter is inserted in the line between the transmitter and its load to eliminate harmonics while permitting the radiation of the fundamental.

Fig. 18.7 shows two convenient methods. The capacitance values are not critical and usually lie between 500pF and 0·001 μF; such values are large enough for adequate by-passing at television frequencies and yet have negligible effect on the supply carried by the lead. The capacitors should preferably be of the feedthrough or mica-disc type because these have exceptionally low series inductance. When other types of capacitors are used the connecting leads must be kept as short as possible: a typical arrangement is shown in **Fig. 18.8.** It will be noted that the portion of the lead *inside* the box is shown as being screened. The use of screening is strongly recommended on all internal leads which are not normally carrying r.f. currents. In such screened leads the insulating material need not be of a low-loss type since any r.f. loss will help to attenuate any r.f. currents which might be present.

The choke shown at (B) in Fig. 18.7 may be made by winding approximately one-quarter wavelength (at the frequency to be attenuated) of suitable wire on a former about ⅜ in. diameter.

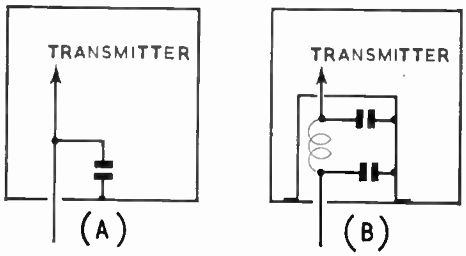


Fig. 18.7. Methods of preventing the escape of r.f. energy along the supply leads feeding a screened transmitter. The pi-filter shown at B is usually more effective than the simple bypass capacitor A

It is well to bear in mind that r.f. chokes often have series resonances and may therefore offer negligible impedance at certain frequencies. These frequencies are not readily predictable but they can be changed by altering the number of turns in the coil or the spacing between the turns. A grid dip oscillator (see Chapter 19—*Measurements*) is useful in detecting such resonances.

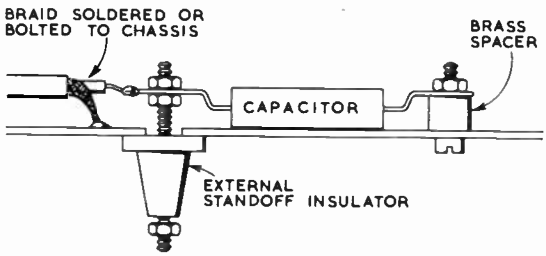


Fig. 18.8. An effective method of using a standard moulded-case mica capacitor for bypassing a power supply lead entering a screened cabinet.

Where the power supply unit is remote from the transmitter chassis some difficulty may be encountered in constructing an effective r.f. filter for the heater supply leads owing to the relatively heavy currents and the consequent large amount of space occupied by the chokes. This may be avoided by mounting the heater transformer directly on the transmitter chassis, as near as possible to the valve sockets, and the r.f. chokes for the heater supply can then be inserted in the primary circuit instead of the secondary.

The effectiveness of lead filters may be checked by coupling sensitive harmonic indicator or a receiver fitted with a signal strength meter to the lead in question. In some cases a simple crystal set may be sufficient.

The screening box itself should preferably be made from sheet brass or aluminium, although when ventilation has to be provided it may be satisfactory to use perforated or expanded metal sheet. Fine-mesh woven wire gauze will give good results but its performance may deteriorate with time owing to the development of high-resistance contacts at the intersections of the wires. There should not be less than ½ in. overlap at every joint, and bolted joints should have bolts not more than 3 in. apart. It is not always essential to screen meter holes, but when necessary the meters can be mounted in screened enclosures (with the leads bypassed to the enclosure).

Screened transmitters built according to the principles described and used in conjunction with a screened low pass filter will give a substantially harmonic-free output irrespective of the transmitter circuit details and of the amount of harmonic energy produced in the various stages. The principles are particularly easy to apply to self-contained equipment such as a " table-top " transmitter.

RECEIVER CONSIDERATIONS

The majority of interference problems occurring in receivers can perhaps be analysed in terms of selectivity of the various tuned circuits, although sometimes the source of trouble can be traced to other parts of the receiver such as the a.f. amplifier or the power supply. In general, they may be grouped under one or more of the following headings:

(*a*) Adjacent-channel selectivity.

(*b*) Spurious responses in superheterodyne receivers.

(*c*) Cross-modulation.

(*d*) Rectification in a.f. stages.

(*e*) Mains-borne interference.

Adjacent-channel Selectivity

The overall selectivity of the receiver determines its ability to reject transmissions occurring on adjacent channels. As the channel spacing varies according to the mode of transmission so likewise does the receiver bandwidth. Typical bandwidths vary from a few megacycles in television receivers to a few hundred cycles in receivers for hand-keyed telegraphy reception. In some types of receiver the bandwidth is variable. Because television receivers have the greatest bandwidth they are the most susceptible to interference from spurious signals radiated by transmitters whose fundamental output is in another part of the frequency spectrum.

Spurious Responses in Superheterodyne Receivers

Interference effects in a superheterodyne receiver can be of several different forms including interference by image signals, the break-through of signals on or near the intermediate frequency, and various beats produced by harmonics of the signal or of the local oscillator.

Images. If the local oscillator in a superheterodyne receiver has a frequency f_0 and if the intermediate frequency is f_i, the receiver will produce an output when the incoming signal has a frequency of either $f_0 + f_i$ or $f_0 - f_i$. If, as in most receivers, the oscillator frequency is designed to be higher than that of the wanted signal, the unwanted or " image " signal will be on a frequency $f_0 + f_i$. The only possible means of preventing interference by an image signal is to provide sufficient selectivity ahead of the mixer stage.

"Half-image" Frequency. It is possible for signals on frequencies of $f_0 + \frac{1}{2}f_i$ and $f_0 - \frac{1}{2}f_i$ to produce at the mixer anode a frequency of $\frac{1}{2}f_i$ together with its harmonics. The second harmonic of this frequency is of course f_i and interference may therefore occur. It should be noted that $f_0 - \frac{1}{2}f_i$ is relatively close to the wanted frequency of $f_0 - f_i$ and that the pre-mixer selectivity which is adequate for image-frequency rejection will not necessarily preclude this form of interference. Fortunately this effect is only likely to be encountered when the interfering signal is exceptionally strong since the output of such harmonics from the mixer is quite low.

Intermediate Frequency Breakthrough. This type of interference is due to the reception of a transmission occurring on or near the intermediate frequency of the receiver. It is usually due to insufficient selectivity ahead of the mixer stage but in rare cases it may result from direct pick-up in the i.f. amplifier wiring. Most communications receivers have adequate pre-mixer selectivity to avoid this breakthrough effect and some broadcast receivers have a built-in i.f. trap to reject stations working near the intermediate frequency (usually 465 kc/s). Many television receivers have no i.f. traps and are rather susceptible to i.f. breakthrough.

Harmonic Beats. Interference may be caused by any signal whose fundamental or harmonics can beat with the receiver's local r.f. oscillator or with any of its harmonics and thereby produce the intermediate frequency. The effect can be minimized by reducing the harmonic output of the local oscillator, but in the case of interference from an amateur transmitter it is usually easier to prevent the interfering signals from reaching the receiver input by installing a suitable filter in the aerial feeder at the input of the receiver.

Cross-modulation

Cross-modulation occurs when two signals are applied to a non-linear device such as a badly adjusted or overloaded valve amplifier and results in each signal being modulated by the other. The effect used to be very marked when sharp cut-off screen grid and pentode valves were in common use as r.f. amplifiers but it has ceased to be important in normal reception since the advent of the variable-μ valve with its gradual change of slope. However, when the interfering signal is very strong cross-modulation will occur even with variable-μ valves and, because the modulation from the interfering signal is impressed on the wanted signal, no amount of selectivity in the stages following the one at which the effect was produced will remove the interference.

Ideally all the selectivity should be achieved ahead of the first valve in the receiver so that cross-modulation cannot occur. As this is impossible it is common practice in communications receivers to incorporate the most selective circuit, usually a crystal filter, immediately after the mixer valve and care is taken that the r.f. stages have only enough gain to obtain a reasonable signal-to-noise ratio. This ensures a low signal level at the mixer grid and hence minimizes cross-modulation.

The effect is only likely to occur when the interfering signal is very strong and it usually shows itself as a speech background to the wanted signal. When the interfering station is using telegraphy the strength of the wanted signal may vary in synchronism with the keying. In both cases no trace of the interfering signal will be heard if the wanted signal ceases.

While not strictly a cross-modulation effect, it is appropriate to mention here a property of most a.m. detectors which allows a strong signal to dominate a weaker one. If the interfering signal is sufficiently strong at the detector stage the wanted signal may completely vanish leaving only the interference. This effect is variously known as *wipe-out*, *blanketing* or *overloading* and usually occurs only in receivers having very poor selectivity.

Rectification in A.F. Stages

Appreciable r.f. energy from a strong local transmitter can be picked up directly by the a.f. stages of a receiver; rectification may then occur owing to non-linearity in the amplifier characteristics and interference will result. The interference may be in the form of speech or, in the case of telegraphy interference, fluctuating output. The effect is not restricted to radio receivers and it may be observed in public address equipment, record players, tape recorders, electronic organs and deaf aids. It is sometimes experienced in ordinary telephones and when this occurs the Post Office or telephone company will undertake to suppress it.

The r.f. energy may enter the equipment from the aerial or along the leads or microphone cables or may even be picked up by the internal wiring. The trouble can usually be cured by fitting a small bypass capacitor (100-500 pF) between the grid of the offending stage and earth. Often a grid stopper will suffice, the resistor in conjunction with the input capacity of the valve then forming an r.f. attenuator. The component values are normally found by trial-and-error.

It has been found that such troubles in a.f. amplifiers may be cured by connecting bypass capacitors of 0·01 μF from the heater leads to the chassis. Also it is sometimes worth while to connect an r.f. bypass capacitor across a cathode bias resistor which, although it may already be bypassed for a.f. currents by a large capacitor, is not effectively bypassed for r.f. currents owing to the appreciable self-inductance of the a.f. shunt capacitor. A suitable value would be about 0·001 μF.

High fidelity (hi-fi) equipment is also susceptible to r.f. pick-up on the loudspeaker leads, since negative feedback to the earlier stages is taken from the output transformer secondary. Bypassing should be done with care, since it is possible to cause supersonic oscillations in the amplifier.

If 1·8 Mc/s and 3·5 Mc/s transmissions are picked up by the video stages of a television receiver interference to the picture will be caused without any necessity for rectification.

The problem may be difficult to solve because installing r.f. bypass capacitors will completely upset the performance of the receiver. It is fortunate that this effect is rarely encountered since the only satisfactory cure is to improve the screening in the receiver.

Direct pick-up on the c.r.t. grid and the use of unscreened leads from the vision output to the c.r.t. may also be the cause of interference.

Mains-borne Interference

Even when the transmitter is fitted with a mains filter a nearby receiver may suffer from mains-borne interference owing to r.f. energy being induced in the house wiring from the transmitting aerial especially if an indoor or a low outside aerial is used at v.h.f. A cure can usually be effected by fitting capacitors (of the order of 0·005 μF) from each mains lead to the receiver chassis although a more elaborate filter might be necessary in stubborn cases. In general the receiver should have a good short earth connection but the effect of removing the earth lead should not be overlooked since in some instances this may be found to reduce the interference.

Mains-borne interference is seldom very troublesome but a.c./d.c. receivers are more susceptible than purely a.c. types. If the latter have a mains transformer with a screened primary winding there should be little or no r.f. energy entering the receiver circuits: even when the screen is omitted the capacity of the transformer winding to earth is often sufficient to prevent interference which may be carried by the mains. In a.c./d.c. receivers it is important that the chassis should be connected to the neutral line.

Amateur Station Receivers

An unskilled operator may have some difficulty in deciding whether interference is due to inadequate selectivity or to a "broad" transmission. If the receiver can be detuned some 10 kc/s from a strong telephony signal without greatly impairing the quality of the speech, the selectivity can be considered poor: if the speech is received as "splatter," the transmission is at fault.

Image interference is usually no problem when using the more elaborate type of communications receiver having two r.f. stages although it is frequently troublesome at 14 Mc/s and higher in the simpler receivers. The interference can be reduced by connecting a parallel *LC* circuit, tuned to reject the image signal, in the aerial lead close to the aerial terminal of the receiver. Image rejection can be improved in the design stage by choosing a relatively high value (e.g. 1·6 Mc/s) for the intermediate frequency since this increases the frequency difference between the signal and image frequencies. The adjacent-channel selectivity of a normal i.f. amplifier operating at 1·6 Mc/s or above is not very good and it has become popular to use a double-superheterodyne circuit with a lower second i.f. with which it is comparatively easy to obtain a high degree of selectivity. In such an arrangement, of course, the selectivity is concentrated at the back end of the receiver thus leaving the front-end wide open for cross-modulation to occur. The design of double superhet receivers incorporating crystal or mechanical filters and local oscillators of sufficient stability for the reception of s.s.b. signals is described in Chapter 4—*H.F. Receivers.*

A further improvement can be achieved by using the transmitting aerial for reception rather than the nondescript piece of wire that is so often relied upon, since most transmitting aerials are resonant.

Severe cross-modulation may appear as a result of "improving" the front-end of a receiver by changing the first r.f. stage from the original variable-μ valve to a sharp cut-off type in order to increase the signal-to-noise ratio. While this may improve the performance on the higher frequency bands it will be at the expense of cross-modulation trouble on the lower frequency bands where local stations radiate extremely powerful ground waves.

Communications receivers, in common with other short-wave receivers, can cause TVI by the radiation of harmonics of the local oscillator. Such harmonics can be detected by coupling a sensitive indicating absorption wavemeter to the oscillator coil or wiring. For example, interference to television on Channel 1 may be caused by an early model HRO receiver due to third-harmonic radiation when operating on the 14 Mc/s band. This interference can be eliminated by bypassing the heater pins of the oscillator valve to earth and connecting a series-tuned circuit between its cathode and earth as shown at L1C1 in **Fig. 18.9.** The trimmer *C*1 is adjusted for minimum third-harmonic output.

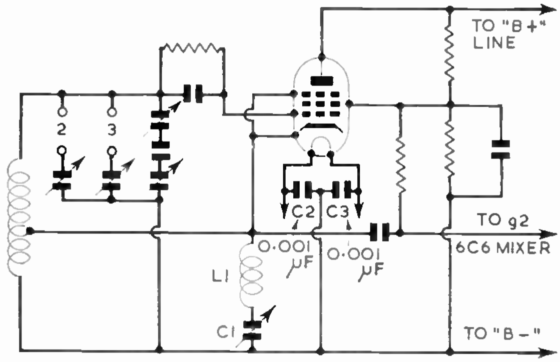


Fig. 18.9. Modification of the oscillator circuit in an early model HRO receiver to suppress the third harmonic when operating in the 14 Mc/s band. C1, 30 pF air trimmer; C2, C3, 0·001 μF mica (connected directly to valve socket); L1, 20 turns No. 22 s.w.g. enamelled copper wire close-wound on $\frac{7}{16}$ in. diameter former, winding length $\frac{13}{16}$ in.

Broadcast Receivers

Modern broadcast receivers are of the superheterodyne type and rarely suffer interference from amateur transmitters on the score of poor adjacent-channel selectivity. On the other hand, because they usually have only one tuned circuit at the signal frequency they are rather susceptible to interference due to spurious responses. Receivers having an intermediate frequency of 465 kc/s have an image response in the 1·8-2·0 Mc/s amateur band when tuned to frequencies between 870 and 1,070 kc/s. As amateur stations operating on this band radiate extremely strong local ground waves it is hardly surprising that interference can be caused not only by image response but also by harmonic beats and cross-modulation.

The interference can nearly always be cured by installing a suitable filter between the aerial and the receiver. A single r.f. choke in series with the aerial and mounted close to the aerial terminal may prove successful in mild cases. A tuned wave-trap is often satisfactory but it is not recommended for general use because it will require re-tuning whenever the transmitter frequency is changed appreciably. The most satisfactory device is a conventional low pass filter.

Filters for use with broadcast receivers are commonly designed for a terminating impedance of 400 ohms, and an example is given later in this chapter. The filter should preferably be screened and it should be fitted as close as possible to the aerial and earth terminals of the receiver to avoid pick-up of the unwanted signal after the filter.

If filtering the input to the receiver proves unsuccessful, the interference is due to r.f. pick-up either by the mains or by the a.f. stages.

Telegraphy interference from coast stations may be experienced in broadcast receivers having no i.f. trap. Widespread interference has also been traced to v.h.f. parasitic oscillations generated in the a.f. stages of broadcast receivers or in gramophone amplifiers. Neither of these effects really concerns the amateur but it is possible that initially he may be blamed for them. It is in cases like these that the value of a well-kept log book becomes unquestionable.

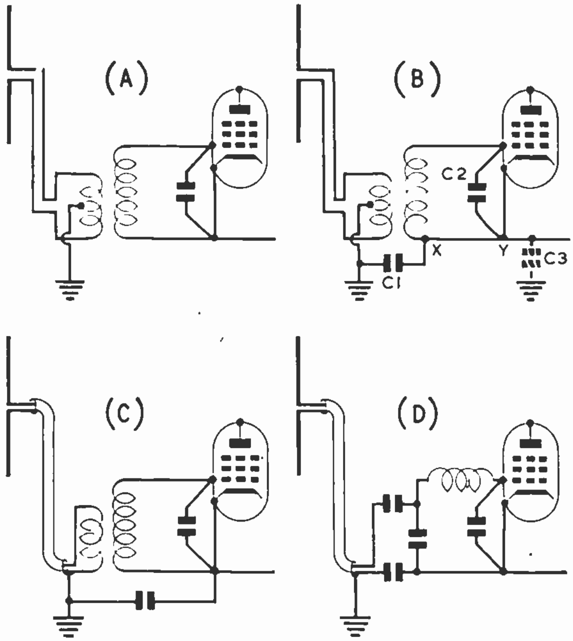


Fig. 18.10. Aerial coupling arrangements in television receivers.

Television Receivers

After harmonic radiation, perhaps the most common form of interference to television receivers by local amateur transmitters is cross-modulation or blanketing due to the presence of a strong unintentional signal at the input terminals. Where a high outdoor television aerial is installed it behaves as an excellent Marconi type of vertical receiving aerial for signals on the lower frequency amateur bands. If the aerial system were balanced, as shown in **Fig. 18.10 (A)**, any pick-up on the feeder from a local transmitter would be the same in each wire and would be self-cancelling at the centre-tapped input coil of the receiver. Thus in spite of heavy interfering currents flowing in the feeder and earth lead reception would be free from all interference except for the slight amount picked up by the dipole itself.

The receiver chassis is usually connected either directly or, in the case of a.c./d.c. sets, through a capacitor to the earth terminal, and this can give rise to interference even with a perfectly balanced aerial system as a result of earth currents flowing in the chassis or the wiring. An example of such a possibility is illustrated in **Fig. 18.10 (B)**: here C1 is the capacitor connecting the chassis to earth, C2 is the tuning capacitor (often merely the valve input capacity) and C3 is the capacity of the chassis and mains lead to earth. Some of the interfering current induced in the aerial will flow directly to earth and some will flow to earth through C1 and C3, the proportions depending on the relative impedances of the two paths to earth. If C1 is connected as shown to the point *X* an interfering current flows in the input circuit between the points *X* and *Y*, and since there may be appreciable reactance between *X* and *Y* an interfering voltage can reach the grid of the valve. The interference may be reduced if C1 is connected to the point *Y* instead of *X*. Chassis currents can also cause trouble in other ways. The situation is greatly eased if C1 is connected to earth through an entirely separate path such as the earth wire of a three-core mains lead. In this case the only interfering current in the chassis will be that flowing through the very small capacitance between the aerial and the grid coil and this could be eliminated by introducing a Faraday screen.

The balanced aerial system shown in Fig. 18.10 (A) is not often used, the unbalanced system shown in **Fig. 18.10 (C)** being far more common. In this arrangement most of the interfering current flows in the outer sheathing of the coaxial cable, and provided that the earthing system is reasonably well designed the selectivity of the first tuned circuit is often sufficient to reject the interference present on the inner conductor.

Fig. 18.10 (D) shows a particularly bad arrangement in which a pi-network is used to match the feeder to the input impedance of the valve. As explained earlier, this circuit gives little protection against signals on a lower frequency than that to which it is tuned. Owing to the absence of inductive coupling, it is impossible to keep the interfering currents out of the receiver by the use of separate earth leads. If the interference is very severe it may be necessary to rebuild the input circuit in more conventional form.

In the majority of cases interfering signals from amateur transmitters can be prevented from reaching the receiver by installing a suitable wave-trap or filter in the coaxial feeder close to the receiver. Wave-traps are useful for exploratory work when trying to identify the interfering frequency, but for a permanent installation a screened high pass filter should be used. Examples of suitable filters are given later in this chapter.

It must be appreciated that the filter will only attenuate interfering signals present on the inner conductor of the coaxial cable. If the effect of the filter is not as good as was expected the residual interference is probably caused by currents on the outer conductor affecting the receiver in a manner similar to that indicated in Fig. 18.10 (B). In such circumstances the interference may be appreciably reduced by ensuring that the current induced in the outer conductor can reach earth without passing through the receiver

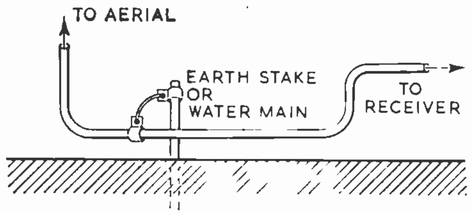


Fig. 18.11. Bypassing interference currents on a TV coaxial feeder cable to earth before they enter the receiver.

wiring, for example by running the cable past an earth stake or the point at which the water main enters the house and there earthing the outer conductor. The cable should then run along the ground until it reaches the receiver, as indicated in **Fig. 18.11.** *Caution*: The earthed outer conductor must not come into direct contact with the receiver chassis.

In some early receivers the intermediate frequency adopted was near the 14 Mc/s amateur band. A frequency of this order was chosen in order to reduce internal beats or " birdies " without any consideration being given to the possibility of interference from amateur transmitters. This unwise choice has lead to many cases of i.f. break-through, the majority of which could easily have been cured by fitting i.f. traps or high pass filters. When i.f. interference cannot be cured by external filtering the trouble is probably due to pick-up in the wiring and this may necessitate major changes in the receiver. In general the amateur is advised not to undertake such work. A satisfactory solution in a case of this type can only be assured by the close collaboration between the amateur, the complainant, the GPO and representatives of the receiver manufacturer. Fortunately such severe cases are rare and i.f. break-through is now of less importance as a result of the general adoption of a high intermediate frequency and the inclusion of i.f. filters.

Most of the interference to television is picked up by the aerial but occasionally it is brought in along the mains. Mains-borne interference can be identified by removing the aerial from the receiver, turning up the contrast control and then noting whether or not there are signs of interference to sound or vision. If interference is present, it is either being picked up in the set wiring or, far more likely, mains-borne. If the latter, it can often be cured by fitting r.f. chokes in the mains leads close to the receiver chassis. These chokes must of course be capable of carrying the receiver mains current (large transmitting-type chokes are generally suitable) and they should be mounted in a well-earthed screening box. The earthing is important not only for the effective suppression of the interference but also in the interests of safety. Raising the impedance of the mains leads may cure interference which is not mains-borne by reducing the stray " aerial " currents in the receiver wiring as described above.

Due to the 3·5 Mc/s spacing between sound and vision carriers in the 405 line system, some receivers are susceptible to a form of interference in which 3·5 Mc/s energy beats with the vision carrier to cause sound interference and vice versa. Sometimes these receivers can be recognised by their radiating the sound on 3·5 Mc/s! The cure is a high pass filter, but as the isolating capacitors in series with the feeder may present appreciable reactance at 3.5 Mc/s, it may be necessary to mount the filter on the set side of these. *Caution*: One pole of the mains is connected to the chassis.

CONTACT RECTIFICATION

Any substantial lengths or areas of metal which make partial contact with each other will, by virtue of the existence of oxides and other substances associated with tarnishing, behave like an aerial system having a detector somewhere along its length. If two signals are present simultaneously in this system, cross-modulation may occur and the wanted signal, modulated by the unwanted signal, may re-radiate from the metalwork and be picked up by a local receiving aerial. Even if no cross-modulation takes place, harmonics of the signals present may be generated as a consequence of the non-linear behaviour of the partial contact and these may again be radiated and cause interference. In the event of cross-modulation no amount of selectivity at the receiver can reduce the interference, and in the event of harmonic generation nothing can be done at the transmitter to cure the trouble.

Common causes of the effect are rusty joints in gutters, drain pipes, gas pipes, electrical conduit, etc. Unless the effect is only slight, the source may often be located by the use of a battery-operated portable receiver having two r.f. stages tuned to the offending harmonic frequency and having a tuned loop aerial. The transmitter should be modulated and operated at full power while the receiver is moved round the neighbourhood exploring for points of origin as indicated by maximum receiver output. The tuned loop aerial will be found to be directional enough to locate faulty joints in hidden conductors in walls and under floors.

Even when the source has been identified it may be difficult to effect a cure since the faulty joint may be inaccessible or may be on the property of a third party. If the joint is accessible, the trouble can easily be eliminated by improving the electrical contact, but if it is not little can be done unless the receiving aerial can be orientated so as to minimize the interference.

CURING TVI

This section is intended to be a practical guide to an amateur confronted with a complaint of interference. The emphasis is therefore on logical procedures, rather than on repeating the particular application of topics dealt with elsewhere in the chapter. For example, the bald statement "Apply wavetraps " implies " For details see the section on wavetraps and filters."

When it is realized that in an area of moderate field strength (1mV/metre) a television receiver is working on a received signal of about 0·1 micro watt, it is not surprising that a high degree of suppression is necessary to prevent interference. As it has been found possible, however, to reduce the harmonic output of transmitters to 0·0001 micro watt from a 100W output transmitter, i.e. 120db attenuation, the problem should not be regarded as insoluble.

It will be helpful to have a complete circuit diagram of the transmitter, including control and mains wiring, so that no potential radiators are neglected, and to keep a record of all tests, even those which do not work, since this will avoid doing the same test again later.

Before TVI can be cured the nature of the trouble must be correctly diagnosed. The chart shown in **Fig. 18.12** sets out a logical series of tests which will enable this to be done expeditiously. Once the cause is known the cure should follow the lines already described.

Although the cure may be simple from a technical point of view its accomplishment may present numerous difficulties. A great deal of tact is called for on the part of the amateur and it is important to remember that the complainant is seldom technically minded and, having spent a considerable sum of money on his receiving equipment, he feels, rightly or wrongly, that he is entitled to interference-free reception. The amateur should never become embroiled in an argument on " rights "; if the complainant appears unreasonable he should be referred to the GPO and the amateur himself should also contact the local Radio Branch of the GPO stating the facts of the case as completely as possible. The local GPO engineers are usually very co-operative and possess test equipment not always available to the amateur. Nevertheless they are often extremely busy and their assistance should not be requested unless it is absolutely necessary.

It is not to be expected that each amateur should have full facilities for dealing with all TVI problems, and here the value of co-operative effort becomes apparent. All well-organized radio groups and clubs should have a small nucleus of members who can form a TVI committee. The committee as a whole should be reasonably well skilled in the diagnosis of TVI and should have available, as club property, test equipment such as high pass and low pass filters, wave-traps, coaxial couplers and adaptors. If an

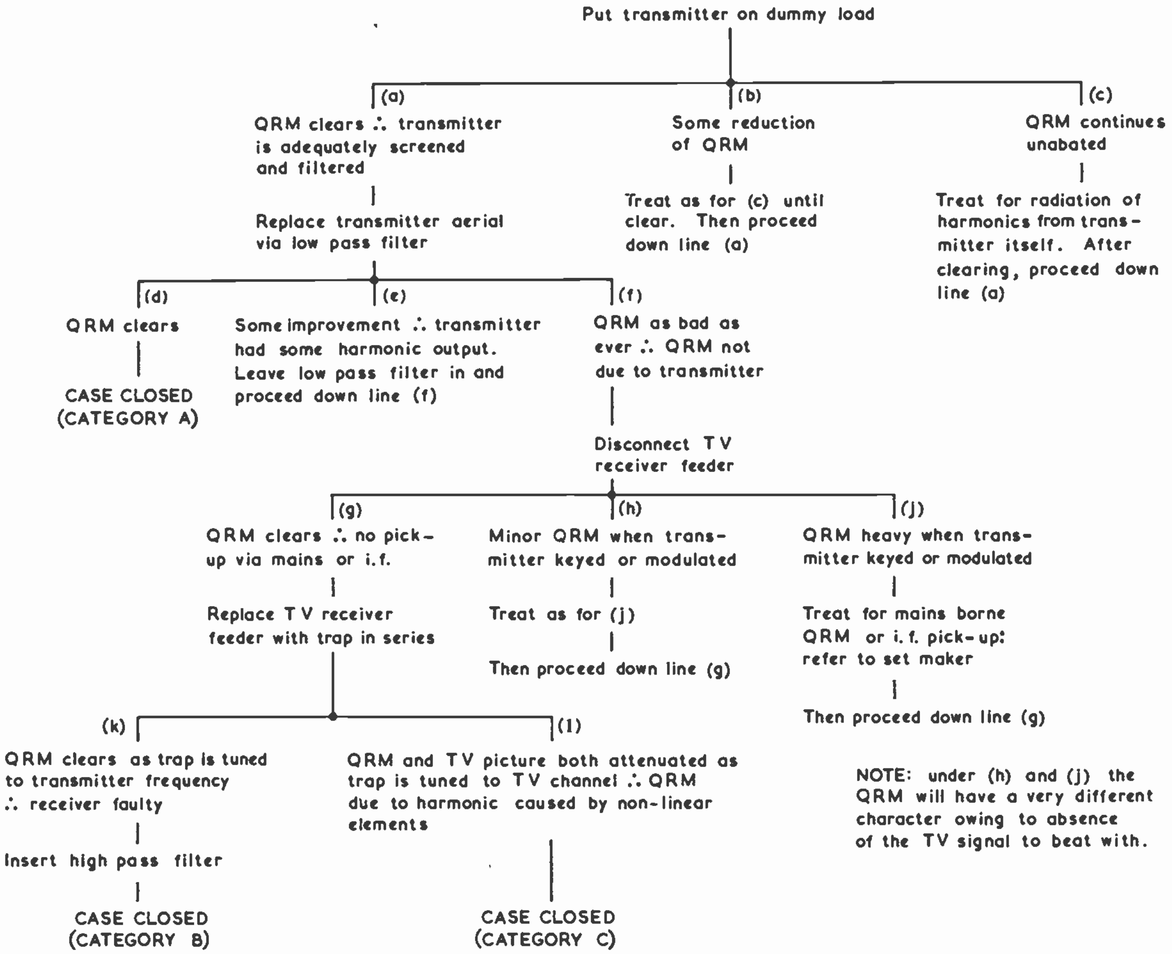


Fig. 18.12. Chart devised by G2IG for the rapid diagnosis of television interference.

old t.r.f. television receiver can be acquired or borrowed it will be found invaluable in checking interference to BBC transmissions as convincing demonstrations can be given in the complainant's house when the trouble is due to spurious responses in a superheterodyne receiver. The services and facilities of the TVI committee should preferably be available to *all* local amateurs (whether club members or not) because a badly handled case of TVI directs public opinion against Amateur Radio as a whole.

There are three categories of TVI in which the amateur may be involved:

(*a*) Harmonic, parasitic or spurious radiation from the trensmitter and/or its aerial system.

(*b*) Response by the television receiver to signals outside its proper passband.

(*c*) Cross-modulation or the generation of harmonics caused by contact-rectification in non-linear elements in the vicinity of the transmitter which re-radiate and enter the television receiver in the same manner as if they were radiated from the transmitting aerial.

Cases in Category (*a*) must obviously be treated at the transmitter and the amateur cannot escape responsibility. Those in Category (*b*) can be cured only at the receiver, and in general the GPO is sympathetic towards the principle that the amateur is not to blame. In Category (*c*) neither the transmitting amateur nor the receiver owner is to blame except insofar as either of them may have somewhere about his property metalwork which, owing to corrosion or the manner of its assembly, is causing trouble. A corroded receiving aerial of course comes into this category and the owner has the cure within his own province. In some cases it may be desirable to resite the receiving and transmitting aerials, bearing in mind the " second comer " policy adopted by the GPO.

Category (a). Causes which must be dealt with at the Amateur Transmitting Station

The system to be adopted in this case is as follows:

1. Connect the transmitter to a dummy load. Operate the transmitter in all other respects in the same manner as that used when interference is known to be caused.

Possible Results

(i) Interference no longer caused.
(ii) No change in interference.
(iii) Appreciable reduction of interference.

If the results are as in (i) then it is clear that all the trouble is brought about by the signal radiated from the transmitting aerial. It may, therefore, be due to harmonic radiation, to receiver defects in category (b) or to effects in category (c).

If the results are as in (ii) there is strong evidence of harmonic radiation from the early or final stages of the transmitter and well-known methods of cure, such as screening and filtering of leads, should be applied. It is unlikely that the receiver is to blame or that non-linear elements are involved since there should be no swamping signal, as would be the case if the transmitting aerial, instead of the dummy load, were in use.

If the results are as in (iii) there is every likelihood of a combination of harmonic radiation from the transmitter itself as in (ii) plus further interference falling into Categories (a), (b) and (c). The procedure, therefore, is to work on the transmitter screening and filtering, etc., until interference is eliminated on dummy load.

2. When all interference on dummy load has been cured, the following test should be carried out. Reconnect the aerial to the transmitter through a low-pass filter of good or known performance.

Possible Results

(i) Interference no longer caused.
(ii) No change in interference.
(iii) Appreciable reduction of interference.

If the results are as in (i) this is the end of this particular branch of investigation and the case is closed. However, if the results are as in (ii) there is strong evidence that the transmitter was blameless even without the low pass filter and that the case falls into either Category (b) or Category (c) or both.

If the results are as in (iii) the transmitting station with the low pass filter in circuit is probably now blameless and the remaining interference is due to causes in Categories (b) or (c) or both. It is, of course, necessary to make sure the low-pass filter is really effective before these assumptions can be true.

At this stage of the investigation the transmitting station and, therefore, Category (a) have been eliminated and only Categories (b) and (c) remain.

Category (b). Causes which must be dealt with at the Television Receiver

3. The system to be adopted in this case is as follows:

Disconnect the aerial from the television receiver and turn up the brilliance control until the raster is just visible. Modulate the transmitter by speech or keying and check whether interference persists.

Possible Results

(i) No interference visible.
(ii) Significant interference still present.

If the results are as in (i) then the interference is coming in via the aerial and the frequency of the interfering signal should be checked. This is best done by means of a tuned trap or traps which will cover the fundamental and appropriate harmonic frequencies of the amateur signal (see Section 4 following).

If the results are as in (ii), then at least some interference is entering the receiver via the mains connection or is being picked up on the i.f. wiring in the receiver. Apart from putting r.f. chokes in the mains lead and trying elementary screening around obviously vulnerable i.f. circuitry there is not much that can be done by anyone but the set manufacturer.

4. Reverting to Section 3 (i)—the case where on removal of the receiver aerial no trace of interference is to be seen when the transmitter is keyed—the following tests should be carried out.

Insert a parallel tuned circuit, resonant at the transmitter output frequency, in series with the inner conductor of the receiver co-axial feeder. For 14 Mc/s the tuned circuit should preferably cover at least a 3 : 1 frequency band so that at one sweep of the tuning capacitor both transmitter fundamental and third harmonic can be rejected. For lower frequency bands the tuned circuit need only resonate at the transmitter output frequency but a second tuned circuit should be available to cover the television band.

With the transmitter keyed or modulated, and the television transmission on the air (preferably showing a test card), rotate the trap capacitor in the vicinity of the known resonance point for the transmitter frequency as determined with a grid dip meter.

(i) If a substantial reduction in interference is observed, then the trouble is either swamping (cross modulation) or i.f. break-through or image response. Which it is can usually be deduced from a knowledge of the receiver circuit but it is of academic interest only since the receiver is at fault anyway.

When it is found that a trap resonant at the transmitter output frequency is effective in reducing interference, a properly designed high pass filter of known performance should be inserted in the receiver feeder. Any remaining interference is probably due to causes in Category (c).

(ii) If no appreciable reduction is observed on tuning the trap to the transmitter output frequency, the evidence is that the receiver is not at fault. Retune the trap—or insert a second trap—to the television channel. Clearly, if the trap is operating effectively, it will seriously attenuate the picture.

If the interference is due to a transmitter emission (such as a harmonic or spurious signal) or to a Category (c) source, then the trap will attenuate the interference to the same extent as the picture. In earlier tests it has already been established that there is no transmitter output in the television band. Therefore, we have the case of a harmonic-free transmitter and a faultless receiver, yet harmonics are being received.

From this it may be deduced that the cause is in Category (c) and sheer dogged searching or inspired deduction are needed to find it and attempt a cure.

Category (c). Harmonics caused by Non-linear Elements

The problem of harmonic generation by contact rectification has already been referred to.

The commonest causes are rusty joints in domestic plumbing such as gutters, drain pipes, gas pipes and electrical wiring conduit. Indeed, the phenomenon has been called for many years the " drain pipe effect " or " rusty bolt effect "—the latter, particularly in sea-going installations where an earth bolt has rusted, giving rise to the conditions described. More often than not the efficiency of the rectifier in the corroded joint is very poor and the proportion of harmonic re-radiated to the amount of the fundamental re-radiated is very low, but it must be realized that a field strength of many volts per metre at the fundamental is common in the immediate vicinity of the transmitting station, and a re-radiated harmonic field of 1/1,000,000 compared with the fundamental may be sufficient to cause TVI. Occasionally, however, the nature and condition of a rusty joint may be such as to rectify quite efficiently, with the result that any modulation of the transmitter may become audible at the joint! In one case, for example, a gutter pipe 20 ft. high with a loose-fitting joint about 5 ft. from the ground was found to be emitting an audible tone when the transmitter was being modulated for test purposes. On disturbing the joint by vigorously shaking the pipe, the sound output vanished but there was still a varying degree of harmonic radiation (as detected on a harmonic indicator) as the pipe was moved about.

Some of the most obscure causes, which are at the same time most difficult to cure, are rusty conduit pipes embedded in the plaster of walls. The only hope of tracing these is by means of a sensitive harmonic indicator, preferably in the form of a portable two r.f. stage battery-operated receiver working at the harmonic frequency and having a tuned loop aerial or flexible probe terminating in a loop. The transmitter should be modulated and operated at full power while the portable receiver is taken around the neighbourhood exploring for the points of origin and maximum harmonic indication. The tuned loop aerial will be found quite directional enough to pin-point even hidden conductors in walls and under floors.

After the source has been located it may be an altogether more difficult problem to eliminate the generation of harmonics. One of the most disheartening things about this particular trouble is that houses immediately either side may also contain rusty connections which in most cases cannot normally be dealt with.

Further Aids to Diagnosis

One of the commonest forms of TVI is the diagonal " cross hatch " pattern formed on the picture. By observing and measuring the horizontal spacing of the light and dark bars **Fig. 18.13** it is possible to deduce the interfering frequency For example, suppose the horizontal pitch of the pattern so formed is 0·5 in. on a display 20 in. wide, then there would obviously be 40 complete cycles of the interference "beat" (or heterodyne) occurring in the 80 microseconds of active line (forward scan) duration of the television picture. If 40 cycles take 80 microseconds, then 1 cycle takes 2 micro-

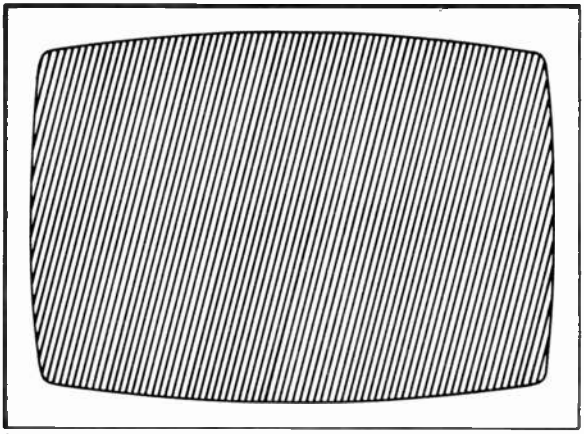


Fig. 18.13. The approximate frequency of c.w. interference to television can be found from the " herringbone " pattern on the screen of the affected receiver due to the beat with the vision carrier. Since the horizontal forward scan period is 80 or 50 microseconds for 405 or 625 lines respectively, the number of dark stripes counted as shown, divided by 80 or 50 gives the beat frequency in Mc/s. It is also possible to distinguish between i.f. and signal frequency interference since for i.f. interference the pattern will change if the fine tuning is varied.

seconds and the frequency is 0·5 Mc/s. Similarly, a heterodyne of 2 Mc/s would be represented by a horizontal pitch of one quarter of 0·5 in. i.e. $\frac{1}{8}$ in.

If the transmitter is on a frequency of, say, 14·333 Mc/s its third harmonic will be exactly 43 Mc/s, and this harmonic will beat with a vision carrier on 45 Mc/s to produce a heterodyne of 2 Mc/s. Thus, if the interference is due to the third harmonic, an $\frac{1}{8}$ in. horizontal pitch pattern will be produced on a 20 in. display. Changing the transmitter frequency to exactly 14 Mc/s will produce a 3 Mc/s heterodyne and the pitch should reduce in width to two-thirds of the previous measurement.

Any pattern having a pitch detectably larger than $\frac{1}{8}$ in. on a 20 in., display (in the case of 14 Mc/s and a 45 Mc/s vision carrier, for example) is indicative of a heterodyne at a frequency lower than 2 Mc/s. Such should be impossible if the trouble is really third harmonic since the transmitter would have to operate outside the high frequency end of the 14 Mc/s band to produce any heterodyne appreciably lower than 2 Mc/s.

On the other hand, if the trouble is due to i.f. breakthrough or image response in the receiver, heterodynes of this order can be caused. Furthermore, due to " inversion " produced by the mixing process in the receiver, it is possible to increase the pattern pitch instead of reducing it when the transmitter is changed from 14·333 Mc/s to 14 Mc/s.

CURING BCI

Generally, BCI is easier to deal with than TVI but again it is quicker to do tests in a logical series, as set out in the chart in **Fig. 18.14.**

The GPO deals with BCI complaints in much the same way as TVI, and it is worth noting that complaints are only entertained which relate to local BBC stations. In the interests of " keeping the peace " it may, however, be advisable to consider listeners to other authorized stations.

F.M. Reception

Interference to Band II reception is seldom met, partly due to the " capture effect " of frequency modulation, which automatically rejects weak interference, and also because a filter adequate for TVI is also adequate for this purpose. If harmonic interference is met, it should be treated as for TVI.

However, as the standard i.f. of 10·7 Mc/s is in harmonic relationship with the 3·5 Mc/s band, interference can be caused by radiation at this frequency. If an aerial tuning unit does not eliminate the trouble, the generation of this frequency should be minimized by reducing the input to multiplier stages and employing a tuned interstage coupling (or a wideband coupler with rapid fall-off outside the desired bandwidth). A high pass filter at the receiver may effect a cure, but if 10·7 Mc/s is being radiated, this should be suppressed at the transmitter, since other receivers will be similarly affected. Clearly it is possible for a receiver overloaded by harmonic free 3·5 Mc/s r.f. energy to generate interference at 10·7 Mc/s. (This can be enhanced by a resonant feeder.) A high pass filter is the cure here.

The assistance of a nearby amateur listening around 10·7 Mc/s is helpful, but the normal station receiver can be used away from the transmitter, with a 3·5 Mc/s wavetrap to prevent overload.

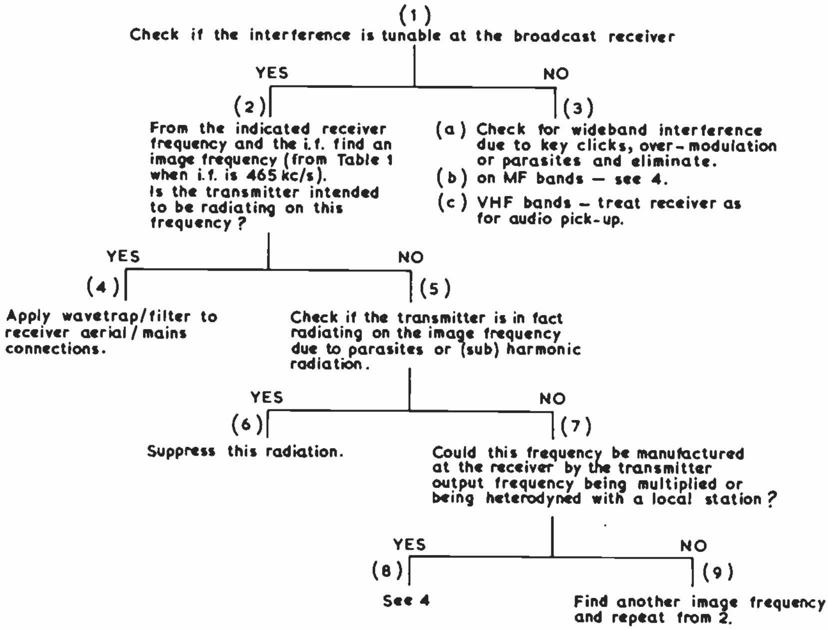


Fig. 18.14. Chart for the diagnosis of BCI.

Spurious Responses of Broadcast Receiver

Against the frequency of each BBC station shown in Table 18.1 are listed those spurious response frequencies which lie in or near amateur bands or at sub-harmonics when the receiver in use has an i.f. of 465 kc/s. Allowance must be made for the variation in i.f. over the normal range 440-470 kc/s. Receiver local oscillator harmonics up to the third have been considered. Frequencies of higher responses can be calculated from the formula

$$f = n(f\,sig + i.f.) \pm i.f.$$

where n = order of harmonic

$f\,sig$ = broadcast station frequency.

TABLE 18.1

Station		Possible Spurious Responses	
Light	200 kc/s	1795 kc/s	
Third	647 kc/s	1759 kc/s	3801 kc/s
North	692 kc/s	1849 kc/s	
Scottish	809 kc/s	1739 kc/s	
Welsh	881 kc/s	1811 kc/s	3573 kc/s
London	908 kc/s	1838 kc/s	3654 kc/s
South and West	1052 kc/s	1982 kc/s	3499 kc/s
Midland	1088 kc/s	2018 kc/s	3571 kc/s
North	1151 kc/s	3697 kc/s	
Light	1214 kc/s	3823 kc/s	
Northern Ireland	1340 kc/s	None	
South and West	1457 kc/s	None	
Third	1546 kc/s	3557 kc/s	

WAVE TRAPS AND FILTERS

A wave trap is a resonant circuit so connected that signals at its resonant frequency are severely attenuated. It may be either a parallel tuned circuit placed in series with the main circuit, or a series tuned circuit shunting it at the specified frequency. Both systems will largely divert current at the unwanted frequency. Cases of spot frequency interference can be effectively dealt with by the use of wave traps. When a band of frequencies must be rejected, the more sophisticated wave filter must be used.

An example of the use of a parallel-tuned circuit is the i.f. trap in a broadcast receiver. Here the voltage from the aerial is applied to the trap in series with the receiver input circuit. Since the impedance of the tuned trap circuit is very much higher than that of the receiver input circuit the interfering voltage is much reduced.

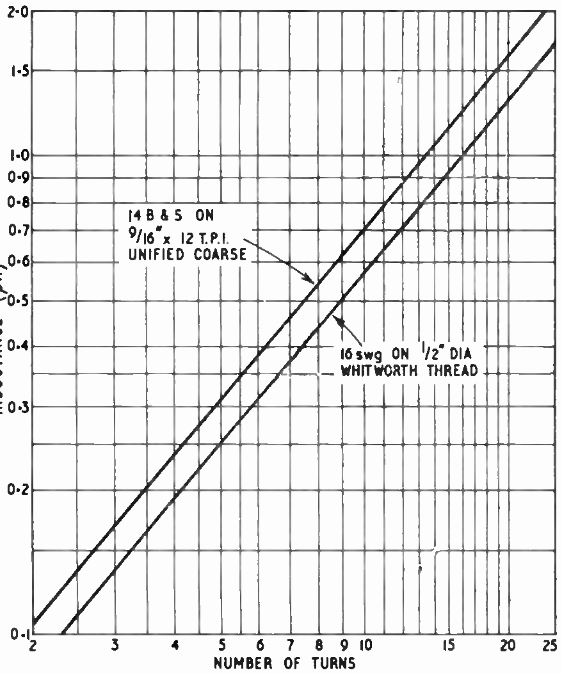


Fig. 18.15. Inductance of air spaced coils wound on bolts used as mandrels.

A parallel-tuned wave trap can be a powerful tool in identifying interfering frequencies. A shielded trap having a set of plug-in coils can be inserted into the coaxial feeder of a television receiver and tuned to the point at which the interference is reduced, thus indicating the offending frequency. Such a trap may be calibrated either by means of a grid dip oscillator or by connecting it in series with the input of a calibrated communications receiver, the latter method having the advantage that the calibration can be carried out with the shielding in place. A tuning capacitor of not less than 150 pF is recommended in order that a reasonable frequency range can be covered.

A simple and cheap but effective filter can be made from a coaxial cable stub cut to an electrical quarter wavelength at the unwanted frequency and connected in parallel with the main coaxial cable. The remote end of the stub, which may be coiled up for convenience, should be open circuited. The length will be that of a free space quarter wavelength multiplied by the velocity factor of the particular cable in use, typically 0·67 for solid polythene insulation and 0·85 for the foam type. Alternatively, a short circuited stub may be used to pass a desired frequency and its odd multiples. This technique, which can take the form of a balun, is particularly suitable for 28 Mc/s where the capacitance of an open circuit stub cut to 43 Mc/s would be excessive. It can also be used as a bandpass filter on a television aerial feeder, provided a separate feeder is used for each channel.

Adjustment simply consists in cutting the stub a little longer than the theoretical length, and then pruning by about ½ in. steps until the minimum signal is obtained, taking care not to short circuit the stub when cutting the cable. (If necessary, overshoot and then cut a new stub to the correct length.) This procedure may also be quicker for a short circuited stub; rather than short circuiting each time, prune for a minimum, and then short circuit.

Wave trap components are usually of the variable type and therefore the correct setting can be obtained by trial and error whereas filter components (other than those used in " brute-force " filters) are fixed at values determined by the design formulae. The values should be within 5 per cent of the calculated values, and to this end capacitors may be purchased with the correct value or, if a suitable measuring bridge is available, they can be made up of series and parallel combinations.

The graph in Fig. 18.15 enables air cored coils to be wound to a reasonable degree of accuracy by using a bolt as a mandrel. The coils can be squeezed or stretched to tune m-derived sections to their exact rejection frequencies. This is more important than the exact L and C values and can conveniently be done with a grid dip oscillator. To preserve the Q and inductance, coils should always be mounted at least one diameter from any metalwork.

It is important that a filter is correctly terminated. A 150 watts transmitter (400 watts p.e.p. output) will produce 180 volts r.m.s. in a matched 80 ohm feeder, but in the event of mis-match the maximum voltage is 180 times the s.w.r., so a very high voltage can be produced which may damage the filter. A coupling unit is therefore necessary, preferably incorporating some form of reflectometer. Mis-termination may also introduce a variable loss across the passband, due to reflections, which can seriously distort television reception, and will require a transmitter to be re-tuned.

In broadcast receivers the input impedance is usually unknown, but filters for these receivers are based on an impedance of 400 ohms, and a 400 ohm resistor is often included in the filter. As a slight loss in the passband (which would not be permissible for television) is quite acceptable, the effect of mis-match is not too serious.

Filters for r.f. use should always be enclosed in screening boxes with suitable screened connections. Failure to ensure good earthing of the outer screening of the coaxial cable via the plug and socket to the box can result in interference currents flowing on the outside of the cable and bypassing the filter. It is also desirable to divide the box into separate screened compartments, so that no coupling exists between the various sections although if a low amount of attenuation only is necessary, say 50db or less, it may be sufficient to orientate the coils with their axes at right angles and well spaced i.e. several coil diameters.

Connecting leads, particularly on capacitors, should be as short and stout as possible to avoid adding inductance. A rule of thumb is that $\frac{1}{2}$ in. of 22 s.w.g. wire is equivalent to 0·01 μH; a 200 pF capacitor with $\frac{1}{2}$ in. leads and no internal inductance would be series resonant at 80 Mc/s, and would cease to be an effective bypass above this frequency. Copper or brass tape is much superior in this respect; if the newer types of capacitor with tape leads are not available it is possible to solder strips of copper foil along the existing round wires.

Theoretically it is possible to design a filter, build it and put it into service, no alignment being necessary. In practice this is only possible where a very high degree of performance is not required, e.g. a low pass filter to pass up to 15 Mc/s and stop 40 Mc/s and above. When a more difficult requirement has to be met—for example, the passband required to extend to 30 Mc/s and the stopband to start at 35 Mc/s—component tolerances and stray capacitance and inductance necessitate some adjustment at least to m derived sections. If a signal generator or grid dip oscillator is not available, a small electron coupled crystal oscillator with its anode circuit tuned to the appropriate harmonics may be used as a signal source, and a simple diode voltmeter can be made up as shown Fig. **18.16** to act as the detector. This permits the resonant frequencies of the m derived sections to be adjusted to give maximum attenuation at the design frequencies, and allows an estimate to be made of the attenuation by replacing the filter by a coaxial adaptor and noting the increase in output. A quicker if less accurate procedure is to insert the filter in the aerial of a receiver tuned to the unwanted frequency, and adjust for minimum response, not forgetting that the receiver a.g.c. will tend to counteract the drop in signal strength.

When testing a low pass filter for use with a transmitter it should be inserted into the coaxial cable feeding a communications receiver. Within the passband of the filter there should be little or no effect on the strength of incoming signals. To test the effectiveness in the stopband the filter may be inserted into the feeder of the television receiver: it should be found that the television signals are greatly attenuated or even totally suppressed. With a high pass filter for use with a television receiver there should be no reduction in signal strength when so used, but when the

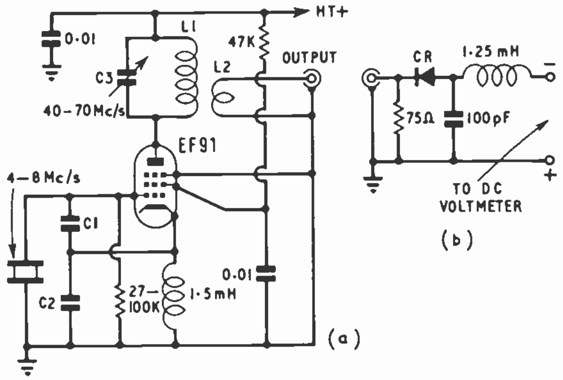


Fig. 18.16 (a). Filter test oscillator using crystals in the range 4-8 Mc/s. Increasing the value of the grid leak may give more harmcnic output. C1, 30 pF (4-6 Mc/s crystals) or 15 pF (4-8 Mc/s crystals); C2, 100 pF; C3, 50 pF; L1, 3 turns 16 s.w.g. enam. on $\frac{3}{4}$ in. diam. former; L2, 1 turn link winding adjacent to L1. (b) Simple d.c. detector.

filter is inserted into the feeder of the communications receiver all signals should be greatly attenuated. The reduction in S meter reading with a given signal when the filter is inserted may be taken as a rough measure of the filter attenuation.

High Pass Filters

It is difficult to design a high pass filter to fit the needs of all amateurs without undue complexity, since operating habits and television receiver susceptibilities vary. For this reason a series of 75 ohm " building blocks " (Fig. **18.17**) have been designed so that each amateur can select those sections which will give him the characteristic required when assembled according to the following instructions:

(*a*) From the bands used and the television receiver i.f. or image frequencies, decide which frequencies need maximum attenuation, some attenuation, and those which can be neglected.

(*b*) For each band requiring maximum attenuation pick the corresponding m derived section (for 28 Mc/s, for example, the matching half sections, which must be used, will contribute high attenuation, but another full section may be added if extreme attenuation is needed).

(c) For all bands, add up the attenuation (Table **18.2**) given by the chosen sections, plus that of the end half sections, subtract 6db for losses and consider whether this is adequate. If not, the addition of one or two constant *k* sections will probably correct matters, and may enable one of the other sections to be discarded.

(d) Draw out the circuit diagram of the whole filter placing the sections chosen between the matching half sections and replacing pairs of capacitors in series by the resultant.

(e) Design the coils with the aid of Fig. 18.15.

(f) Build it.

(g) Temporarily short circuit the series tuned sections and resonate with a grid dip oscillator.

It is quite possible to construct a small filter of this type in a cylindrical tin with all the coils soldered to the lid. Using this technique, it is possible to build a filter in less than an hour, with a measured attenuation of over 30db in the 28, 21 and 14 Mc/s band, and less than 0·5db above 40 Mc/s.

Low Pass Filter

The low pass filter shown in **Fig. 18.18** is a well proven design which gives a minimum of 45db attenuation of all harmonics above 40 Mc/s and is suitable for use in most areas of reasonable television field strength. In fringe areas it may be advisable to add extra constant *k* sections, i.e. insert extra L3 C3 combinations between C2 and L3. The end sections are designed to resonate at 42 Mc/s, and C1 and C4 may be made variable for this purpose. In order to maintain a high attenuation across Channel 1 it is desirable to align one at 42 Mc/s and the other at 43 Mc/s.

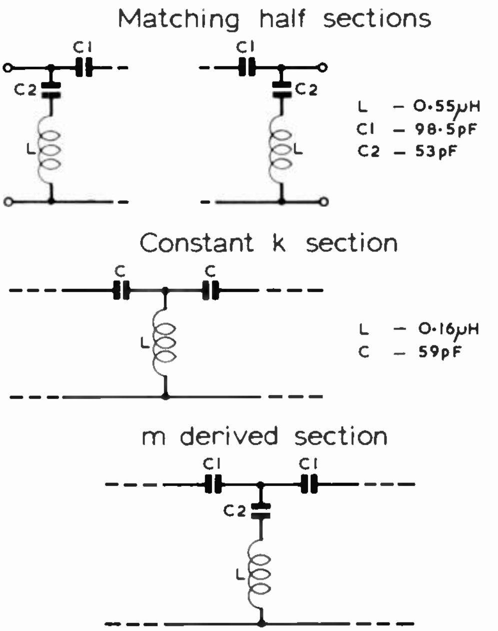
The working voltages of the capacitors should be at least 750 volts to provide a reasonable factor of safety during the initial adjustment of the aerial tuning unit, but in many cases a lower voltage rating may be found practicable. As previously described, the a.t.u. must be properly adjusted if optimum performance is to be obtained from the filter. When correctly terminated a filter of this type has an attenuation characteristic which should give adequate harmonic suppression in all locations.

TABLE 18.2

The attenuation in db of different sections

Band	Pair of matching half sections	28 Mc/s section	21 Mc/s section	14 Mc/s section	7 Mc/s section	Constant K section
28 Mc/s	28 *	28 *	15	12	10	10
21 Mc/s	17.5	17.5	40	23	19	19
14 Mc/s	13.5	13.5	23	40	27	27
7 Mc/s	12	12	19	30	40	40
3.5 Mc/s	12	12	19	28	40	40
1.8 Mc/s	12	12	19	28	40	40

* Due to the width of the 28-29·7 Mc/s band, the attenuation falls to this figure at the edges, if sections are peaked at the band centre. It may be desirable therefore to stagger tune the end sections (or an additional section) to 28·5 and 29·5 Mc/s. The attenuation at the series resonant frequency in all cases will exceed 60db, but the figures of 28 or 40db represent the minimum attenuation at any point in the band if the circuit is resonated at mid-band.



	28 Mc/s	21 Mc/s	14 Mc/s	7 Mc/s
L	0·28μH	0·21μH	0·18μH	0·17μH
C1	98·5pF	74pF	64pF	60pF
C2	107pF	300pF	700pF	3000pF

Fig. 18.17. High pass filter sections for use in 75 ohm cable.

V.H.F. FILTERS

At frequencies higher than 30 Mc/s a low pass filter is often inadequate to deal with interference, since harmonics of the fundamental drive frequency above and below the output frequency can lead to trouble. A bandpass filter is therefore necessary.

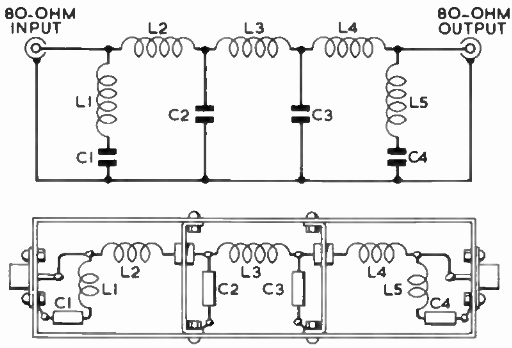


Fig. 18.18. Circuit and layout of a four-section low-pass filter suitable for use with any transmitter on all bands 1·8–30 Mc/s. It is designed for insertion in a 75 ohm coaxial feeder. C1, C4, 36 pF mica, 750 V d.c. working (5% tolerance). C2, C3, 120 pF mica, 750 V d.c. working (5% tolerance). L1, L5, 0·36 μH: 7 turns, winding length 1 in. L2, L4, 0·59 μH: 10 turns, winding length 1 in. L3, 0·73 μH: 12 turns, winding length 1$\frac{3}{16}$ in. All coils are of 16 s.w.g. copper wire, ½ in. internal diameter self-supporting, with a connecting lead 1 in. long at each end.

Conventional Filters

Bandpass filters can be designed with coils and capacitors but the calculated values at impedance levels around 75 ohms are very awkward in practice, and considerable cut-and-try is necessary to get good results. A simpler technique is to cascade a low pass and a high pass filter. If the respective cut off frequencies are suitably staggered about the working frequency a bandpass filter is produced. Both filters must be *m* derived to improve the matching but even so the filter may introduce some mis-match into the coaxial line. Nevertheless the resulting loss can be held to a low level by careful matching.

Coaxial Line Filters

The coaxial line technique allows conventional filters to be made for the 70 and 145 Mc/s bands, but does need care in construction and would be difficult to apply at 435 Mc/s. Fortunately, at v.h.f., the coaxial line becomes a usable filter element. From the equation for the efficiency of a tuned circuit

$$n = 1 - \frac{Q_L}{Q_C}$$

it is known that for an efficiency of 90 per cent, i.e. 0·5db loss, Q_L must be 12 or less with a Q_U of 120 typical of h.f. band coils. However, a 3 in. copper coax line has a theoretical Q_U of 3,000, so that 0·5db loss would give a working Q_L of 300! Q_L is adjusted quite simply by varying the size of the input and output coupling loops.

Unfortunately, a coaxial line is also resonant at odd or even harmonics depending on how it is connected, so that half the harmonics will not be affected. However, by adding capacitative loading (see Chapter 7—*V.H.F. Transmitters*) which reduces the length of line required and also allows the resonant frequency to be adjusted, the harmonic relationship of the "overtones" of the line is broken up. For example, in the quarter wave 145 Mc/s line described below, the three-quarter wave resonance, which would be at 435 Mc/s for an unloaded line, occurs at 550 Mc/s. If it is desired to eliminate these "overtones," this can be done by making two filters of unequal length, so the "overtones" do not coincide, and coupling them in cascade with an electrical quarter-wave of coaxial cable. Bearing in mind that here the aerial presents a considerable mis-match, the "overtones" are unlikely to give trouble provided they do not coincide with a harmonic.

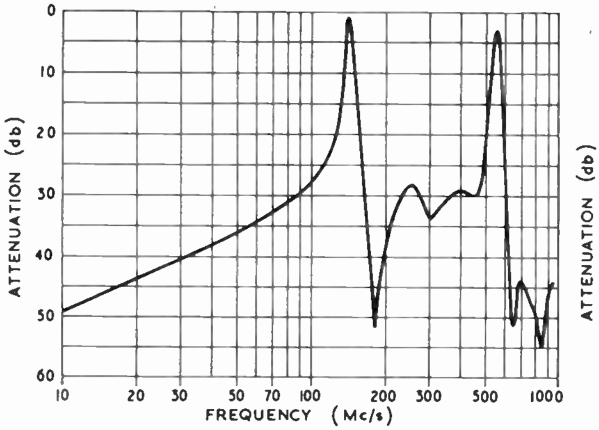


Fig. 18.20. Response curve of the 145 Mc/s co-axial filter of Fig. 18.19.

Although square section line is shown in **Fig. 18.19** for ease of construction, circular section is equally suitable, or strip line which can be conveniently bent round inside a large chassis in order to compress a 70 Mc/s line to a smaller size. Silver plated copper is the ideal material but even tinplate can give a useful performance, although the Q_U is naturally lower.

The performance of a filter built to the design of Fig. 18.19 is shown in **Fig. 18.20.**

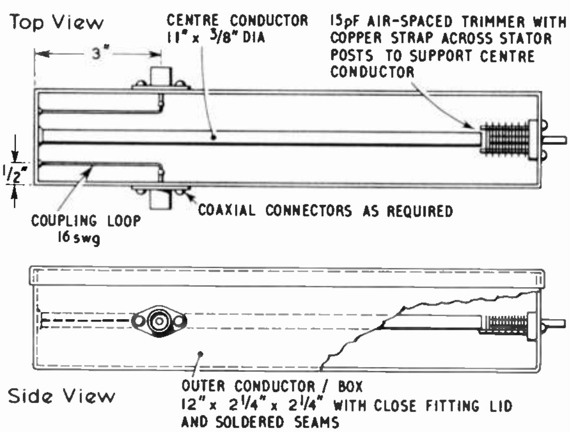


Fig. 18.19. Construction of a 145 Mc/s co-axial filter.

Filter for Broadcast Receiver

A filter circuit suitable for protecting a broadcast receiver from transmitters working on the lower frequency bands is shown in **Fig. 18.21.** The theoretically correct capacitance values are given but in practice good results will be obtained by choosing the nearest standard values. If the inductances L1, L2, and L3 are provided with adjustable iron cores they can be set so that the resonant frequencies of L1C1 and L3C5 are each 1·9 Mc/s and the resonant frequency of L2C3 is 3·65 Mc/s. Since few broadcast receivers have a coaxial input socket this particular filter is provided with terminal connections; it cannot be overstressed that the leads between the filter and the receiver must be as short as possible.

Mains Filters

The usual type of r.f. mains filter is not designed to work into a specific load resistance and the component values are not critical. A typical arrangement for protection from H.F. transmitters is shown in **Fig. 18.22.** For use with a receiver the coils may be made by winding 100-200 turns of 18 s.w.g. enamelled wire on a ½ in. diameter former while the capaci-

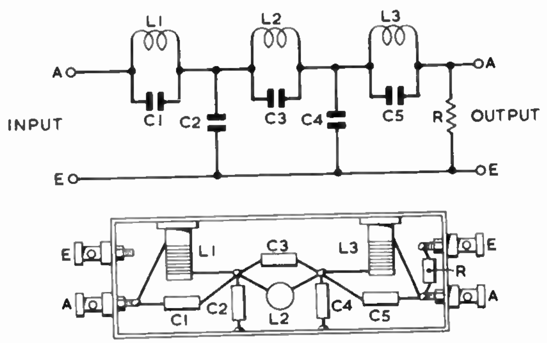


Fig. 18.21. Circuit and layout of a low-pass filter suitable for use with a medium-wave broadcast receiver.

	Calculated values	Suggested nominal values
C1, C5	327 pF	330 pF
C2, C4	357 pF	360 pF
C3	26·2 pF	27 pF

L1, L3, 21·45 μH: 50 turns 32 s.w.g. enamelled copper wire on Aladdin former type F804 with dust-iron core.
L2, 71·7 μH: 90 turns 38 s.w.g. enamelled copper wire on Aladdin former type F804 with dust-iron core.
R, 400 ohm, ¼ watt (10% tolerance).

tors may have a capacitance of 500 pF to 0·005 μF: they should have a mica dielectric and for a 240 volt supply their voltage rating should be at least 750 volts. If the performance is not entirely satisfactory an improvement may be obtained by removing the earth connections from capacitors: this is expecially worth trying when the earth lead is a long one. In some cases the chokes alone will suffice.

If the receiver has a three-core mains cable it may be necessary to fit a choke in the earth wire and rely on a separate earth for the aerial system.

Commercial suppressors are available for installing in mains leads and these are often very effective. A suppressor suitable for protection on medium and short waves should be chosen where the problem is to reduce TVI from amateur transmissions. Filters should always be installed as close to the receiver as possible to minimize the risk of pick-up on the leads following the filter.

Attenuators

Although not strictly a filter, an attenuator can often have a similar effect on interference due to receiver non-linearity such as cross-modulation and overloading. Sometimes, when the trouble is due to a local signal on an adjacent frequency, a filter is not suitable and an attenuator is the only cure external to the receiver. In the case of a television receiver where a high pass filter would be desirable an

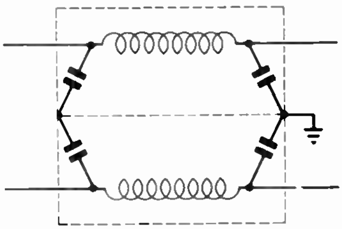


Fig. 18.22. A typical r.f. filter for insertion in mains supply leads. See text for details.

attenuator can be quickly and cheaply made up to effect a temporary cure.

Details of convenient attenuators for 80 ohm lines are given in Fig. 18.23. For other impedances the standard formulae should be used and the values rounded off to the nearest 5 per cent preferred values.

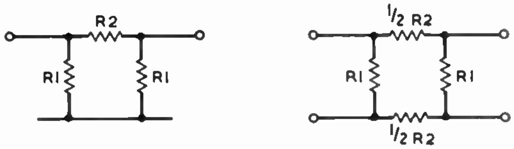


Fig. 18.23. Simple attenuators.

	R1	R2	½R2
10db	150	120	56
15db	120	220	100
20db	100	390	200

The Effect of Mistermination on Filter Performance

The performance of the conventional (Zobel) filter is calculated on the assumption that it is terminated in a highly theoretical Z_o. If the termination is the more practical R_o, i.e. the low frequency value of Z_o, the performance is not greatly modified. For example if the filter is provided with m-0·6 end half sections, the main effect is a reduction of up to 10db in the stop band loss.

Conventionally, a filter is designed with an R_o of 75 or 52 ohms, and it is most important that it is correctly terminated at the operating frequency in order to prevent damage to the filter. In general this correct termination has no predictable effect on the termination at other frequencies, i.e. the actual attenuation of a harmonic will vary from the predicted figure for R_o. In particular if the length of feeder exceeds about 50 ft. the termination seen by the filter can vary widely over a TV channel, so that TVI checking is desirable at several not equally spaced frequencies throughout each band.

Obviously, it is important to find the magnitude of this deterioration in performance. It is theoretically possible to find a filter termination for which the filter acts as a matching section, i.e. all the attenuation is lost! This can only happen away from an " infinite frequency," since a parallel trap in series, or a series trap in parallel is always a fairly good attenuator, and the probability of the worst case termination being produced by a practical aerial system over any range of frequency is remote. Very little work has been done on the problem but Reference [1] suggests a reduction of 14 to 18db which agrees reasonably with the 10db reduction due to changing from Z_o to R_o. Thus the reasonable practical procedure seems to be

(i) Design for 10–20db excess attenuation, using m derived sections for all sensitive frequencies.
(ii) Minimise the feeder length.
(iii) Make checks at several randomly spaced frequencies in each band.
(iv) If the filter performance seems poor, try lengthening or shortening the feeder slightly.

Reference

[1] " High Frequency 250 kW Broadcasting Transmitter," Morcom & Bowers, *Proc. I.E.E.*, February 1966.

CHAPTER 19

MEASUREMENTS

CORRECT operation of amateur radio equipment involves measurements to ensure optimum performance, to comply with the terms of the amateur transmitting licence and to avoid interference to other users. The purpose of these measurements is to give the operator information regarding the condition under which his equipment is functioning. Basically, they are concerned with voltage, current and frequency. For example, in even the simplest transmitter it is necessary to know the grid drive to the various stages (current measurements), the input power to the p.a. (current and voltage measurements) and the frequency of the radiated signal.

If apparatus is to be designed and constructed to give reliable and optimum performance, a considerable range of measuring equipment may be desirable. While it is possible to make do with only one or two simple items of test gear, this chapter describes equipment for a.c., d.c., and r.f. voltage, current and power measurements, wavemeters and frequency standards and a number of more sophisticated items for both transmitter and receiver measurements.

D.C. MEASUREMENTS

The basis of most instruments for the measurement of voltage, current and resistance is the *moving coil meter* in which a coil of wire, generally wound on a rectangular former, is mounted on pivots in the field of a permanent magnet (**Fig. 19.1**). The coil experiences a torque proportional to (i) the current flowing through it and (ii) the strength of the field of the permanent magnet. Current is fed to the coil through two hairsprings mounted near to each end of the spindle. These springs also serve to return the pointer to the zero position (on the left hand side of its travel in standard meters) when the current ceases to flow. Provision for adjusting the position of the pointer is made by a *zero adjuster* accessible from the front of the instrument.

Since the movement of the coil and its associated pointer is proportional to the field of the magnet and that of the current being measured, the scale is linear. A minor disadvantage is that the instrument can only be used on d.c., but it can be adapted to measure a.c. with a suitable rectifier.

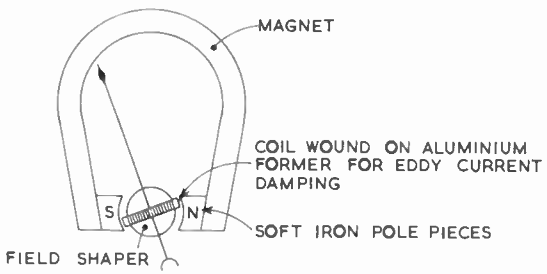


Fig. 19.1. Construction of the moving coil meter.

It is usual to damp the coil system (i.e. prevent it swinging freely after a change of current), a common method being to wind the coil on an aluminium former which then acts as a short circuited single turn coil in which the eddy currents serve to oppose the movement. The degree of eddy current damping is also dependent on the external resistance across the terminals of the moving coil and is greatest when the resistance is low. It is a wise precaution to protect sensitive instruments not in use by short circuiting the terminals.

When a moving coil meter is used for measuring current it is called an *ammeter*, *milliammeter* or *microammeter* depending on its full scale deflection (f.s.d.). A *d.c. voltmeter* is a milliammeter or microammeter equipped with a voltage dropping (series) resistor.

Milliammeters

Milliammeters and microammeters are commonly manufactured with basic full scale deflections of 0–50 μA 0–100 μA, 0–500 μA, 0–1 mA, 0–5 mA and 0–10 mA. For higher current ranges, a *shunt* resistor is connected across the meter

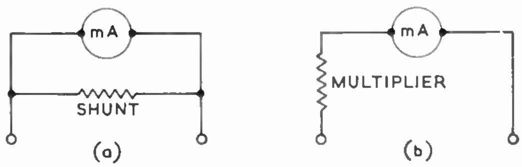


Fig. 19.2. Extending the range of the m.c. meter. (a) To read higher current with a parallel shunt. (b) To measure voltage with a series resistor or multiplier.

(**Fig. 19.2(a)**). The value of the shunt may be obtained from the formula

$$R_s = \frac{R_m}{n-1}$$

where R_s is the resistance of the shunt, R_m is the resistance of the meter and n is the scale multiplying factor.

For example, if a milliammeter of 10 ohms resistance and a f.s.d. of 1 mA is to be used to measure 100 mA, a shunt must be provided to carry the excess current, that is, 100–1 milliamperes (= 99 mA). Thus the required resistance of the shunt is

$$R_s = \frac{10}{100-1} = \frac{10}{99} = 0{\cdot}101 \text{ ohm.}$$

When moving coil meters are used in circuits where high voltages are present (i.e., in p.a. anode circuits) care must be taken to avoid accidental electric shock from the zero adjuster which may be live. If the instrument is of the flush mounting type the front of the meter may be covered with a piece of clear plastic about $\frac{1}{16}$ in. thick.

Voltmeters

A milliammeter may be used to read d.c. voltages by connecting a resistor, termed a *multiplier*, in series with it (**Fig. 19.2(b)**). The value of the multiplier depends on the full scale deflection of the meter and may be calculated from Ohm's Law.

For low voltage ranges, the value of the multiplier can be obtained from

$$R_S = R_M\left(\frac{V}{V_M} - 1\right)$$

where R_S is the resistance of the multiplier, R_M the resistance of the meter, V is the required voltage and V_M the voltage across the meter (this can be determined by applying Ohm's Law to the resistance of the meter and current flowing through it). In practice, however, the resistance of the meter can be ignored and the formula simplified to

$$R_S = \frac{1000\,V}{I}$$

where V is the desired voltage range and I is the f.s.d. of the meter in milliamps.

For example, a 0–5 mA meter is to be used as the basis of a voltmeter to read 100 volts. Then

$$R_S = \frac{1000 \times 100}{5} = 20{,}000 \text{ ohms}$$

It is usual to describe the sensitivity of a voltmeter in ohms per volt; in the example considered above, the meter reading 100 volts for a full scale deflection of 5 mA would be said to have a sensitivity of 200 ohms per volt. In practice, such a sensitivity is too low for accurate measurements in radio and electronic equipment due to the current drawn and a more sensitive meter would therefore be chosen as the basis of the instrument. The lowest sensitivity which can be considered satisfactory for amateur purposes is 1000 ohms per volt (requiring a 0–1 mA meter) but sensitivities of 5000, 10,000, 20,000 and 100,000 ohms per volt are common for accurate work.

Whenever a voltage measurement is made on a circuit of appreciable resistance, the current taken by the voltmeter should be considered to ensure that the operating conditions are not significantly altered by connecting the voltmeter.

The accuracy of a voltmeter depends largely on the accuracy of the multipliers. Precision resistors are the most suitable but rather costly and may be replaced in home-built equipment by 1 per cent close tolerance carbon resistors of adequate power rating. Alternatively, a supplier may be persuaded to select suitable values from normal stock.

It should be noted that even high stability resistors may change in value by several percent with time; if an instrument is required to maintain high accuracy for an extended period only wire wound resistors of suitable low temperature coefficient material should be used.

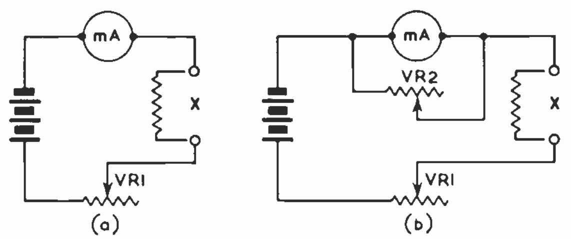


Fig. 19.3(a). Measurement of resistance with a simple ohmmeter. In (b) VR2 is to adjust the sensitivity of the meter to compensate for a drop in battery voltage. In both circuits VR1 is for setting the meter to full scale deflection with the terminals X short circuited.

Measurement of Resistance

The measurement of resistance is based on Ohm's Law, two common arrangements being shown in **Fig. 19.3(a)** and **(b)**. It will be seen that each circuit comprises a battery in series with a milliammeter, a resistor VR1 and an unknown resistor X. In practice, the terminals across which X is connected are first short circuited and VR1 adjusted until the meter reads full scale. When the resistor X is connected across the terminals, the meter reading will fall; calibration can therefore be carried out by connecting a number of resistors of known value across the terminals in turn and marking the scale accordingly. Alternatively, a graph relating meter current and resistance can be prepared.

An instrument designed for resistance measurements is termed an *ohmmeter* and may have several ranges.

The resistance ranges in multimeters are based on these simple circuits but are not very accurate below about 5 ohms. Such inaccuracy is partly due to the resistance of the connecting leads.

A circuit arrangement suitable for measuring low resistances and differentiating between them and short circuits is shown in **Fig. 19.4**. The instrument measures resistance up to 5 ohms by comparing the voltage drops across a standard resistor and the unknown resistor when the same current flows through both. By selecting a 5 ohms resistor as the standard (R1) and using a 0–500 μA meter as a 0–0·5 voltmeter, it is possible to use the calibrated scale, each 100 μA (0·1 volt) division representing 1 ohm. The value of R2 should be 1000 ohms less the internal resistance of the meter.

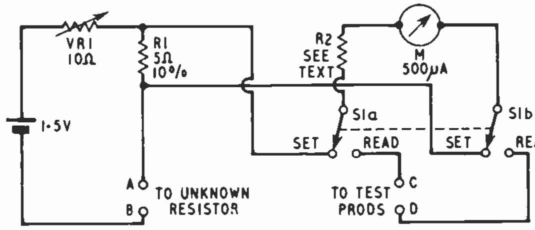


Fig. 19.4. Circuit of the direct reading low resistance ohmmeter. R1 should be a high stability resistor of ½ watt rating. R2 should be 1000 ohms less the internal resistance of the meter. The terminals A and B should be insulated and have large contact areas.

The leads from *A* and *B* terminate in strong crocodile clips and those from *C* and *D* in sharp test prods.

The method of operation is as follows. With the meter switch at SET and the variable resistor VR1 at maximum, the unknown resistor is connected across terminals *A* and *B*. VR1 is then adjusted so that the meter reads full scale.

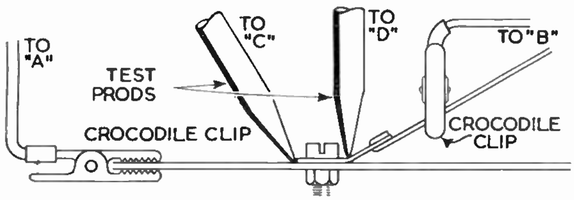


Fig. 19.5. Use of the low resistance ohmmeter of Fig. 19.4.

The switch is next moved to the READ position, the meter then indicating the resistance of the unknown resistor directly.

An example of the use of the instrument is illustrated in **Fig. 19.5** where the resistance between a solder tag and chassis is to be found. Terminal *A* is connected to the chassis and *B* to the wire to the tag at any point; terminals *C* and *D* are connected to the chassis as near to the joint as possible. In this way, the same current flows through the standard resistor R1 as through the joint and the resistance of the connecting leads to *A* and *B* does not affect the reading.

A.C. MEASUREMENTS

The moving coil meter can be adapted to measure a.c. by the addition of a suitable rectifier such as the *copper oxide* type. Such a meter will read a.c. of audio frequencies, indicating the average value, 0·636 of the peak value of a sine wave.

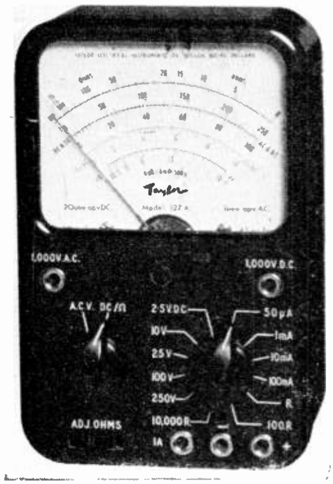
Commercial a.c. instruments of the rectifier type are calibrated in R.M.S. values assuming a sine wave and hence read incorrectly if used on any other wave form.

In conjunction with a *thermo-couple*, a moving coil instrument can be used to read alternating currents of both audio and radio frequencies. The thermo-couple is a junction of two dissimilar metals which, when heated, produces a d.c. voltage. The junction is heated by the current to be measured passing through a " heater " to which it is attached. A disadvantage of the arrangement is that low current readings are rather severely compressed and it is necessary therefore to have several instruments if widely different currents are to be measured. Thermo-couple instruments read true R.M.S. values irrespective of waveform. Unless specially designed, such instruments become less accurate as frequency increases due to the effect of the shunt capacitance.

The Avo Multiminor test instrument which uses a printed circuit.
(Photo by courtesy of Avo Ltd.)

The *hot wire ammeter* is also designed to make use of the heating effect of a current through a wire. In this case, the wire is supported at both ends and kept under tension by a fibre attached to its centre and loaded by a spring. The current through the wire causes heating and expansion to take place so that a spindle, round which the fibre is wound, moves a pointer across a scale. The heating effect is independent of the type of current and both a.c. and d.c. may therefore be measured. Despite its simplicity, the instrument is now rarely used as it is rather inaccurate.

The *moving iron meter* is useful for measurements on a.c. circuits (40–60 c/s) where the current taken is unimportant. In these instruments fixed iron and moving iron elements are mounted in the magnetic field at the centre of a coil through which the current to be measured is passing. Under these conditions, a mutual repulsion exists between the two parts,



The Taylor Model 127A multimeter with a sensitivity of 20,000ohms per volt. The instrument has 20 ranges and measures resistance up to 20 Megohms.
(Photo by courtesy of Taylor Electrical Instruments Ltd.)

the moving iron being deflected, thus indicating on a scale the current measured.

The deflection obtained is approximately proportional to the square of the current or voltage measured (i.e., it is a square law instrument). The scale divisions are thus close together at the low end of the scale and opened out at the higher. This means that accurate readings cannot be taken at the low end but the open scale at the top is sometimes useful. For example, a moving iron meter with a full scale deflection of 250 volts would be excellent for reading small changes of voltage on a nominally 240 volt mains supply but would be useless for measuring voltages below about 50V.

Multirange Meters

A number of shunts and multipliers selected by a switch can be used in association with a single basic meter to form a multirange instrument (often known as a *multimeter*) measuring current and voltage. Meters of the type illustrated on this page, adapted to measure resistance as well as a.c. and d.c. voltage and current are available from a number of manufacturers at prices dependent largely on the size of the meter scale and the sensitivity of the movement. Kits are also available.

A multirange meter can be constructed in two units, the first containing the 0-1 mA meter movement with switches

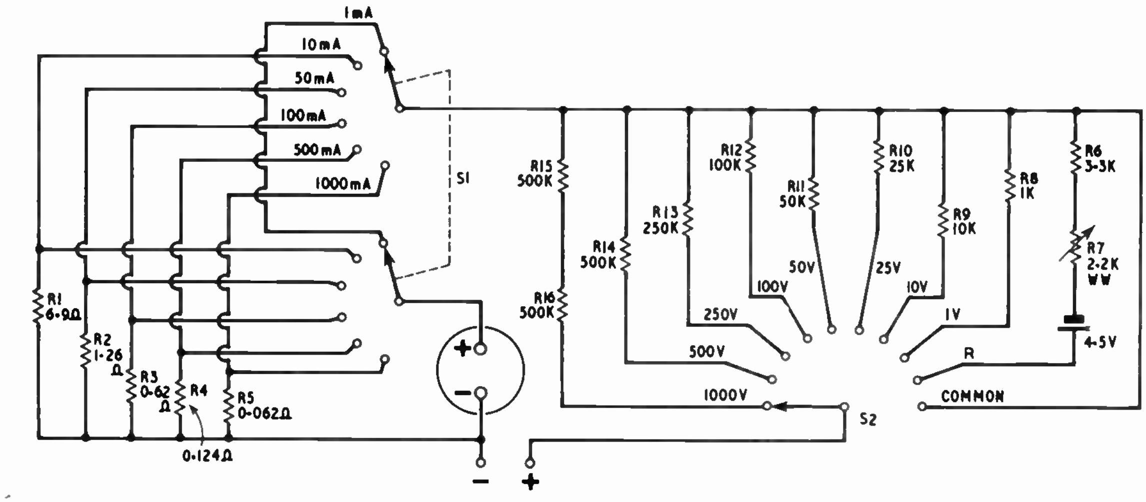


Fig. 19.6. Basic multimeter unit reading current to 1A d.c., voltages to 1000 V d.c. and resistance to 20,000 Ω.

to select various shunt and series resistors to give six d.c. current ranges up to 1 amp and eight d.c. voltage ranges up to 1000 volts. An internal battery provides an ohms range readable up to 200,000 ohms which corresponds with the first division of the meter (0·02 mA). A chart relating scale readings to resistance values is housed in the lid of the box.

An add-on unit contains a meter rectifier with associated switched series resistors to give four a.c. voltage ranges up to 1000 volts while additional shunt and series resistors extend the d.c. ranges to 10 amps and 5000 volts. When using the add-on unit the main instrument is set to measure 1 mA full scale and the add-on unit connected to its terminals by the lugs. The two units could, of course, be mounted in a single box if desired.

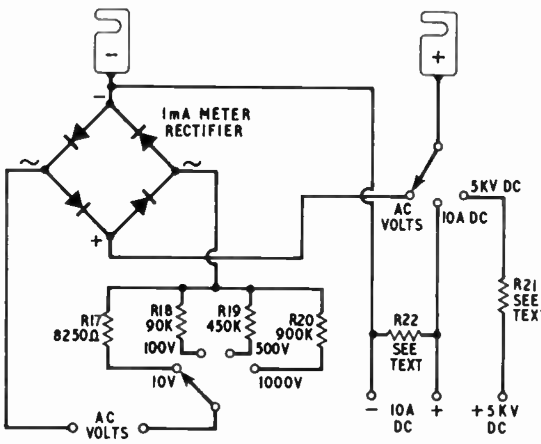


Fig. 19.7. Add on unit to extend the range of measurements provided by the circuit of Fig. 19.6.

The circuit diagrams and component values of the two units are shown in Figs. 19.6 and 19.7. The series resistors are 1 per cent tolerance high stability types while the shunts are made of lengths of Eureka resistance wire. The values of the shunts are correct for a meter of 60 ohms internal resistance but would need modifying for any other meter. In any case, the precise value of each shunt should be adjusted experimentally to give the correct reading against a meter of known accuracy. Care should be taken to allow the shunts to cool to room temperature after soldering in position before any calibration is attempted since their resistance will be modified by temperature.

VALVE VOLTMETERS

Even in its simplest form, the valve voltmeter is of great value in amateur stations, especially those in which experimental work is the main interest. One of its most useful applications is the direct measurement of voltage across the grid resistor of any stage drawing grid current, which makes it unnecessary to provide jacks, meters or their associated decoupling components in the grid return circuit. Due to its high input resistance, the valve voltmeter enables potentials of a fraction of a volt and upwards to be measured with little or no disturbance to the circuit to which it is connected. By the addition of a simple probe, r.f. voltages may be measured as easily as d.c.

In addition to its high input resistance, the valve voltmeter has the advantage that large deflections may be obtained on a relatively insensitive meter movement. It is also much less susceptible to damage by a heavy overload than a sensitive voltmeter of the normal type.

SIMPLE VALVE VOLTMETER

Fig. 19.8 shows the circuit of a simple but reliable instrument. In some aspects the design is not ideal, but it will be found adequate for the demands which most amateurs are likely to make upon it. On the credit side, it will work from almost any receiver or other low-voltage power supply, because it is not unduly sensitive to variations in h.t. voltage.

Layout is not critical, although it is advisable to use a ceramic or p.t.f.e. valveholder because of the high resistance in the grid circuit at the lower voltage ranges. For the same reason, a ceramic switch wafer is most suitable for the range switch S1 which should naturally be mounted on the front panel, as should VR1, the zero-adjusting control. S2 is provided so that the polarity of the micro-ammeter M1 may be reversed, thereby allowing both positive and negative voltages to be read directly. This control is useful rather than essential, and if it is included it should be fitted on the front panel. VR2 is used to adjust the calibration of the instrument. Normally it can be set initially and then ignored, so it may be mounted anywhere on the chassis and should, for convenience, have a slotted spindle.

Adjustment is quite simple. The instrument should be allowed to warm up for 10 or 15 minutes, and then VR1 should be adjusted for zero deflection of the meter. While this is being done, it is desirable to short the tip of the d.c. test probe to the chassis. It will be noticed that unless the prod is earthed, the meter needle will tend to stray a little on the two lower voltage ranges due to hum pick-up. This is a natural effect with most valve voltmeters and need cause no misgivings. (The resistance of many of the circuits to which the instrument is likely to be connected will be low enough to nullify all tendency to error under practical working conditions.) A 1·5 volt dry cell of proven accuracy should then be placed across the input. With the range switch set to the 1·5 volt range and S2 at the correct polarity, VR2 should be adjusted until the meter reads 1·5 volts. As a final check, the polarity of both the battery and S2 should be reversed and the reading noted. If it is unchanged, all is well. If not, the valve is out of balance and should be changed. A very small amount of unbalance may of course

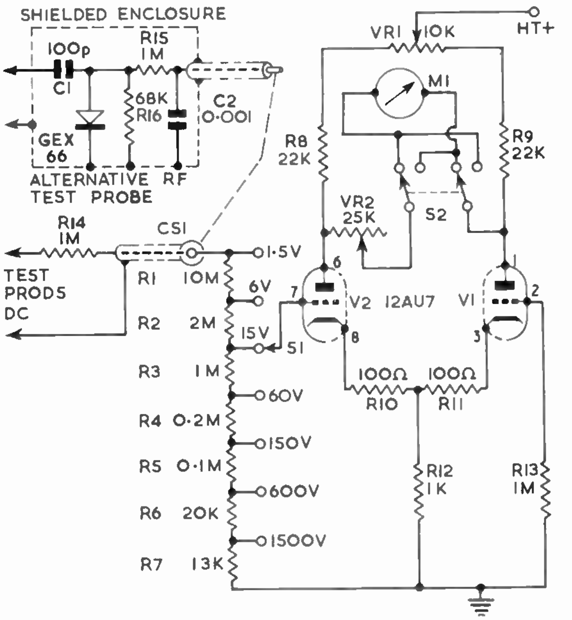
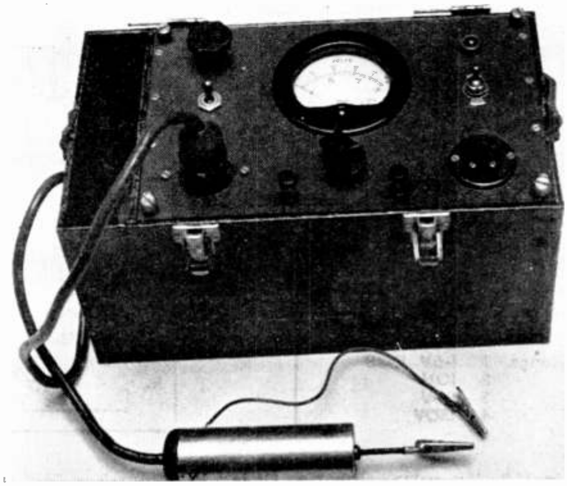


Fig. 19.8. Simplified valve voltmeter. The meter M1 (0-500 μA f.s.d.) should be calibrated 0-1·5 and 0-6 volts. R1-R7 should be 1 per cent tolerance. S1 should have ceramic insulation.



External appearance of the a.c. valve voltmeter.

be tolerated, but that is a matter for each individual constructor to decide for himself. It is wise to age the valve by running it for at least 24 hours or so before carrying out final calibration.

The probes are more or less self-explanatory. Quarter-inch flexible coaxial cable makes good test leads, and although R14 may be a little difficult to attach securely to the inner conductor, it should not be omitted, because it serves to isolate the circuit under measurement from the self-capacitance of the test lead. This enables d.c. to be measured in the presence of r.f. without upsetting the r.f. operation of the circuit. The alternative test probe allows r.f. to be measured in the presence of d.c. The germanium diode should be tested for high reverse resistance, otherwise misleading results may be obtained. C1 should be capable of withstanding the highest d.c. voltage which it is likely to meet. R15 and C2 play no part in the operation of the probe; they are merely a low-pass filter to prevent r.f. leaking back into the valve voltmeter itself.

In operation, a valve voltmeter behaves in the same way as a more conventional instrument. It is, however, desirable to check the zero-adjustment before making a measurement, as this is liable to fluctuate slightly.

A.C. VALVE VOLTMETER

A more sophisticated valve voltmeter suitable for measuring audio and radio frequency voltage up to at least 50 Mc/s and providing useful indications at 150 Mc/s is illustrated above. It is based on a design evolved by S. W. Amos of the BBC.

The circuit (Fig. 19.9) consists of a diode detector producing a d.c. output voltage proportional to the peak value of the applied alternating voltage, followed by a differential d.c. amplifier using a double triode. Considerable d.c. negative feedback is applied to each triode which serves to linearize the input/output characteristic of the amplifier, and also improves the zero stability. Further improvement of zero stability is obtained by applying the output of a second diode to the other input of the differential amplifier. This type of amplifier responds to the difference between the potentials

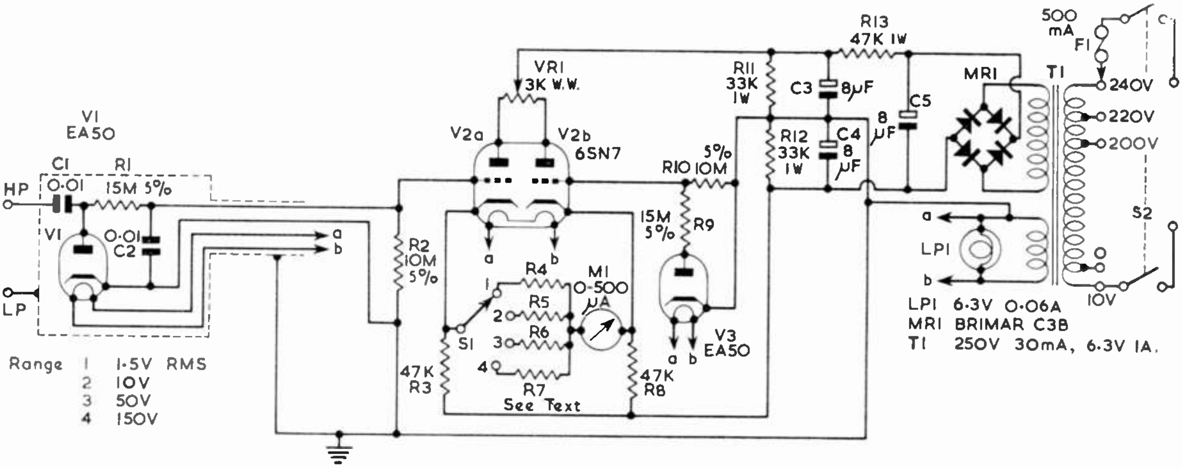


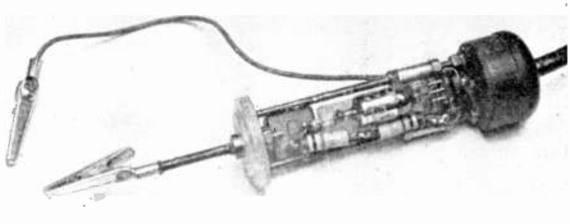
Fig. 19.9. A.c. valve voltmeter. C1 is a mica capacitor, 500V wkg. and C2 mica 350V wkg. R1, R2, R9, R10 should be ½ watt high stability resistors 5 per cent tolerance; R3, R8, should be ½ watt 10 per cent tolerance. The value of R4 is 200 ohms less the resistance of the meter M1. R5, 7K ohms; R6, 39K ohms; R7, 119K ohms; R4, R5, R6, R7 are ½ watt rating, R11, R12, 1 watt 10 per cent tolerance and R13 1 watt 20 per cent tolerance.

applied to its two inputs, thus provided similar diodes are used at V1 and V3, changes in the contact potential of the diodes will not vary the zero setting appreciably.

The diode V1 and its associated components R1, C1, and C2 are built in the form of a shielded probe which is applied to the circuit under test when measurements are to be made. The input resistance of the probe is 3–4 Megohms and the input capacitance approximately 8 pF. The input capacitance would be reduced further by dispensing with the earthed shield but this would render the probe very sensitive to stray pickup, particularly when the instrument is used on its lowest range. An illustration of the probe is shown below. Connection of the probe to the remainder of the circuit is made by a four-way screened lead terminated in a four-pin valve base which fits a standard four-pin valveholder on the panel.

In use the high and low potential terminals of the probe are short-circuited and the meter set to zero by means of the potentiometer VR1 with the range switch S1 at Range 1. The terminals are then separated and the high potential terminal applied to the point at which measurements are to be made, the low potential terminal being connected to chassis or earth. The instrument should be allowed to warm-up for five minutes or so before setting the zero.

Calibration may be carried out at 50 c/s. Provided that the supply is sinusoidal, the reading of a rectifier type voltmeter multiplied by 1·414 will give the peak value of the calibrating voltage. Exact adjustment of the full scale deflection for each range is achieved by variation of the appropriate meter series resistances R4, R5, R6 and R7 by adding an additional resistor in series or parallel with the specified resistor. At least two separate scales on the meter will be required as the 1·5 volt range is non-linear due to curvature of the diode characteristics at low input voltages.



Construction of the probe for use with the a.c. valve voltmeter.

BRIDGE MEASUREMENTS

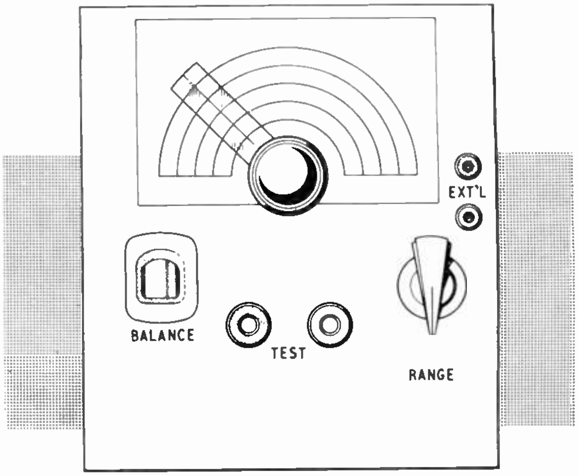
If accurate measurements of resistance and capacitance are required, bridge methods are preferable since greater accuracy and wider range can be obtained.

The circuit of **Fig. 19.10** shows a bridge which will measure resistance between 10 ohms and 10 Megohms and capacitance from 1000 pF to 1000 μF. An unusual feature of this bridge is that electrolytic capacitors can be measured while a polarizing voltage is applied. The leakage current may be monitored at the same time. The bridge is energized from a 6·3 volt 50 c/s source through the capacitor C1; closing the switch S4 reduces the voltage applied to the bridge and assists in locating the balance point.

When measuring capacitance, the arms of the bridge are formed by the balancing control R1, the range resistors R4–9, the series combination R2 C2 and the unknown capacitance X. When measuring resistance, R2 C2 are replaced by R3 and this is interchanged with R1, permitting the same scale calibration to be used for both resistance and capacitance if R3 is correctly chosen. The output of the bridge is amplified by V2 and the signal is passed without rectification to the magic eye indicator which shows balance by minimum shadow area. All the resistors in the bridge should be high stability types and R1 should be a good quality wire-wound component with a linear winding.

When measuring electrolytic capacitors a polarizing voltage may be applied from the built-in power supply through series resistors R16–21 which are selected by S3. The actual polarizing voltage applied to the capacitance will depend on the leakage current but typically can be controlled between 20 volts and 300 volts by S3. The position of this switch marked S/C removes the applied voltage and discharges the capacitor through R22.

The bridge should first be calibrated on capacitance and it is convenient to do this on the lowest range since close tolerance capacitors are readily available in low values. A set of 1 per cent capacitors capable of giving values of 1000 pF, 2000 pF, etc. to 10,000 pF are required since the lowest range measures 0·01 μF full scale. The capacitance calibration will then be correct for all six ranges thus measuring up to 1000 μF. The bridge should be switched to measure resistance by the change-over switch S2 and the range switch set to the highest capacitance range which corresponds to the lowest resistance range. This will measure 100 ohms full scale. A single close-tolerance resistor in the range 10–100 ohms should be connected across the bridge and the dial set to read the known value using the calibration determined for capacitance. The value of the resistor R3 should now be adjusted for balance when all resistance ranges will correspond to the scale markings established for capacitance.



Type of panel layout suitable for resistance/capacitance bridge instruments.

R.F. CAPACITANCE BRIDGE

A bridge operating at 50 c/s is not suitable for measuring small values of capacitance because their reactance is too high at this frequency. The following bridge operates at 1·5 Mc/s and is suitable for measuring capacitances down to 1 pF and up to 1500 pF. The circuit, **Fig. 19.11** is a version of the Wheatstone bridge and is energized from a Colpitts oscillator at about 1·5 Mc/s. The output of the bridge is rectified and applied to a magic eye indicator V2 which is arranged to be fully open when the bridge is balanced.

Calibration

The calibration of the bridge is quite easy, all that is required being a few accurate capacitors. The following capacitors are needed and they should be good quality silver mica with a tolerance of ±1 per cent ±1 pF: 5, 10, 25, 50, 100, 250, 500, 1000 pF. (Suitable types are made by Radiospares Ltd.)

With these standards it is possible to calibrate the bridge throughout its range. Proceed as follows. Start on the lowest range and with no capacitor connected across the terminals, adjust C1 for balance. This point should be marked zero. (This is to balance out the stray capacity. If it is not possible to obtain a balance—the eye must be fully open—solder a small capacitor, about 5 pF, across the terminals until a balance can be obtained.) Next connect the 5 pF standard to the test terminals using the shortest possible leads and adjust C1 for balance, mark scale 5. Repeat the process using the 10 pF standard. Next solder the two together and repeat, marking the Scale 15.

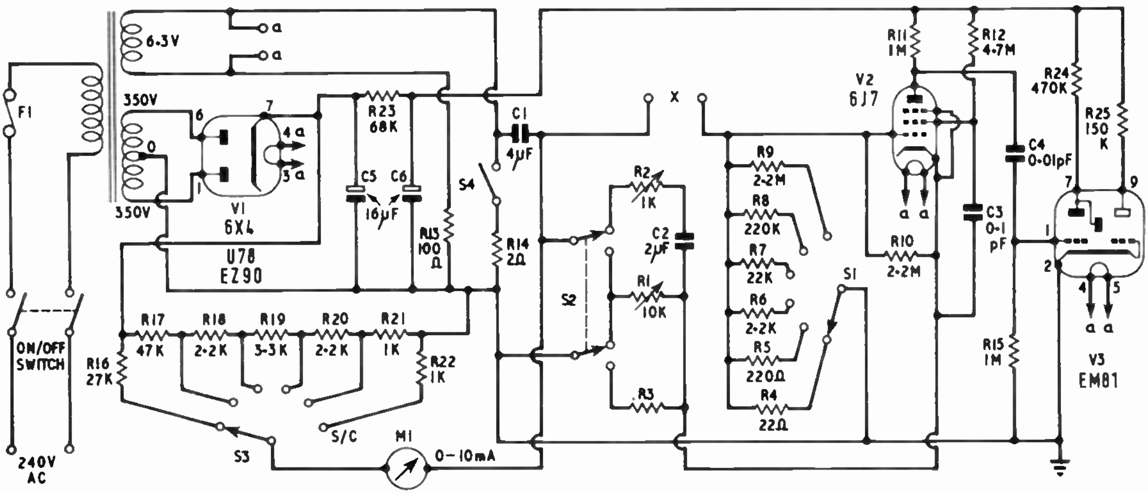


Fig. 19.10. Circuit diagram of the resistance/capacity bridge.

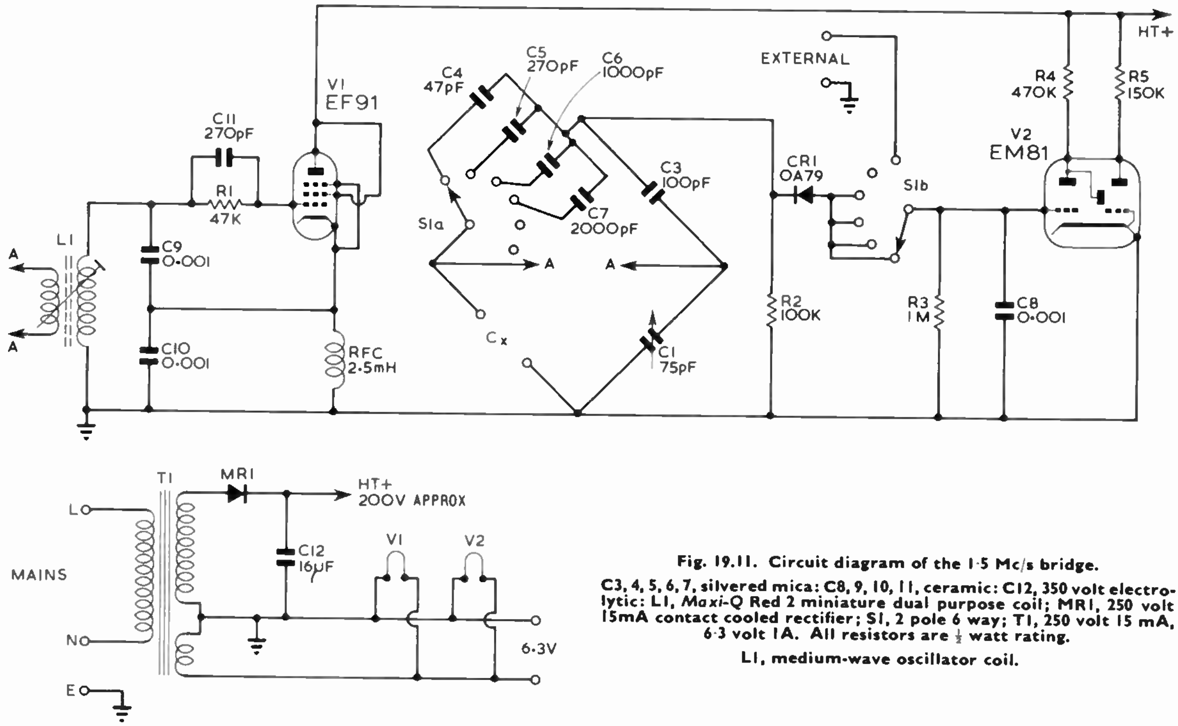


Fig. 19.11. Circuit diagram of the 1·5 Mc/s bridge.

C3, 4, 5, 6, 7, silvered mica: C8, 9, 10, 11, ceramic: C12, 350 volt electrolytic: L1, *Maxi-Q* Red 2 miniature dual purpose coil; MR1, 250 volt 15mA contact cooled rectifier; S1, 2 pole 6 way; T1, 250 volt 15 mA, 6·3 volt 1A. All resistors are ½ watt rating.

L1, medium-wave oscillator coil.

This process should be carried on over the complete range of the instrument, changing ranges where necessary. Each range should overlap the previous one to give complete coverage with no gaps. If it is found that there are gaps, C4, C5, C6 or C7 should be altered to give unbroken coverage from zero to 1500 pF.

If a straight line capacitance type component is used for C1 it will be found that the calibration is linear and any points which have been missed, such as 20 or 200, can easily be interpolated into the scale.

All the components of the bridge should be mounted as rigidly as possible, and once the bridge has been calibrated they should not be moved. The specified 2 pole 6-way switch allows the indicator to be used from an external source and to be switched to zero input during standby periods and for comparison. It is a help if C1 is fitted with a slow motion drive since the null is quite sharp. All the wiring should be run in straight lines to keep stray capacity to a minimum.

RADIO FREQUENCY BRIDGE

The need for an instrument which will measure impedance is felt at some time or other by every experimenting amateur. The instrument normally used is the full r.f. bridge, but commercial r.f. bridges are elaborate and expensive. On the other hand, it is possible to build a simple r.f. bridge which, provided its limitations are appreciated, can be an inexpensive and most useful adjunct in the amateur workshop. In fact, it is essential if experiments with aerials are undertaken.

The instrument described here will measure impedances from 0 to 400 ohms, at frequencies up to 30 Mc/s. It does not measure reactance nor show whether any reactance present is capacitive or inductive, but a good indication of the reactance present can be obtained from the fact that any reactance will mean a higher minimum meter reading.

There are many possible circuits, some using a potentiometer as the variable arm and others variable capacitors, but a typical circuit is shown in Fig. 19.12. The capacitors have to be differential in action, mounted in such a way that as the capacity of one decreases, the capacity of the other increases. The capacitors should be the type which have a spindle protruding at either end, so that they can be connected together by a coupling. To avoid hand capacity effects, the control knob on the outside of the instrument should be connected to the nearest capacitor by a short length of plastic coupling rod. These capacitors form two arms of the bridge, the third arm being the 100 ohm resistor and the fourth the load. Balance of the bridge is indicated by a zero reading on the meter M1.

The construction is simple. The bridge should be totally enclosed in a metal box and the screening indicated in Fig. 19.12 incorporated. All the leads should be kept as short as possible.

On completion, the instrument is calibrated by placing across the load terminals various non-reactive resistors (i.e. not wire wound) of known value. The calibration should preferably be made at a low frequency, where stray capacity effects are at a minimum, but the calibration holds good throughout the frequency range.

In using the instrument, it should be remembered that an exact null will only be obtained on the meter when the instrument is looking into a non-reactive, resistive load. When reactance is present, however, it becomes obvious from the behaviour of the meter: swinging the control knob, it will be noticed that, although there is a minimum reading, a complete null cannot be obtained.

The r.f. input to drive the bridge can be obtained from a grid dip oscillator or other small oscillator of about 1 watt input power. The oscillator is coupled to the bridge by a short length of coaxial cable, terminating in a link coil of about four turns, which is placed on or near the g.d.o. coil. Care should be taken not to overcouple or the g.d.o. may change frequency or even stop oscillating. As the coupling is increased, it will be seen that the meter reading of the bridge increases up to a certain point, after which further increase in coupling causes the meter reading to fall. A little less coupling than that which gives the maximum bridge meter reading is the best to use. The bridge can be used to find aerial impedance and can be used for many other purposes: for example, to find the input impedance of a receiver on a particular frequency.

One useful application of this type of simple bridge is to find the frequency at which a length of transmission line is a quarter or half-wavelength long, electrically. If it is desired to find the frequency at which the transmission line is a quarter-wavelength long, the line is connected to the bridge and the far end of it left open-circuit. The bridge control is set at 0 ohms. The dip oscillator is then adjusted until the *lowest* frequency is found at which the bridge shows a sharp null. This is the frequency at which the piece of transmission line is a quarter-wavelength long. Odd multiples of this frequency can be checked in the same manner. In the same way, the frequency at which a piece of transmission line is a half-wavelength long can also be found. The procedure is the same, except that the far end of the transmission line is shorted instead of being left open. The method in both cases is illustrated in **Fig. 19.13.**

The bridge can also be used to check the characteristic impedance of a transmission line. This is often a worthwhile exercise, since appearances can be misleading. The procedure is as follows:

(a) Find the frequency at which the length of transmission line under test is one quarter-wavelength long. Once this has been done, leave the oscillator on this frequency.

(b) Select a carbon resistor of approximately the same value as the probable characteristic impedance of the transmission line. Substitute this resistor for the transmission line as the bridge load and check its value at the frequency obtained in (a). (This will not necessarily agree with its d.c. value).

(c) Disconnect the resistor from the bridge and re-connect the transmission line. Connect the resistor across the far end of the transmission line.

(d) Measure the impedance now presented by the transmission line at the frequency of sub-paragraph (*a*). Then the characteristic impedance Z_o of the line is given by

$$Z_o = \sqrt{Z_s \times Z_r}$$

where Z_s = impedance presented by the line and Z_r = resistor value.

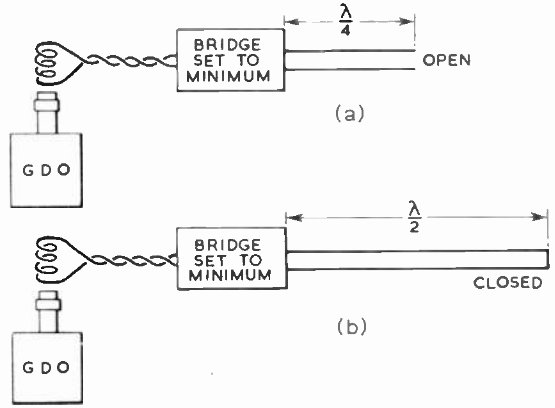


Fig. 19.13. Method of using the r.f. bridge to determine the frequency at which a transmission line exhibits a quarter or half-wave characteristic.

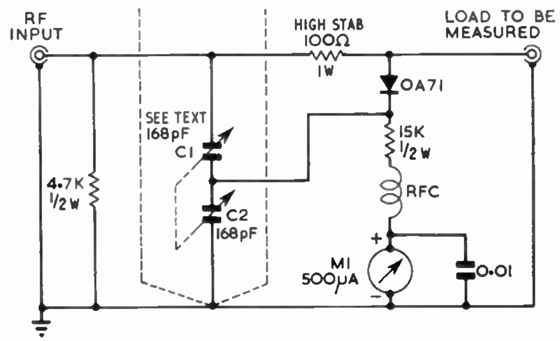


Fig. 19.12. Simple r.f. bridge. C1, C2 form a differential capacitor with maximum value of 168 pF in each section.

FREQUENCY MEASUREMENT

It is important to know the frequency to which a receiver is tuned and the frequency of a signal radiated by a transmitter; in particular, it is necessary to be quite sure that the signal is within the amateur band in which operation is taking place.

For this purpose, a crystal controlled oscillator is almost essential as a reference by which a v.f.o. can be calibrated. Even if the transmitter is directly controlled by a crystal of certified accuracy, it is necessary to ensure that the transmitter is working on the correct harmonic of the crystal and for this purpose an absorption wavemeter is convenient. When constructing new equipment, it is most helpful to be able to check the resonant frequency of the tuned circuits before power is applied; this can be done with a grid dip oscillator.

Standard Frequency Services

Even when a crystal oscillator is used as a calibration source, it is necessary to set it against a frequency standard as the actual frequency obtained from a crystal depends slightly on the circuit conditions. Standard frequency transmissions are provided in the United Kingdom by

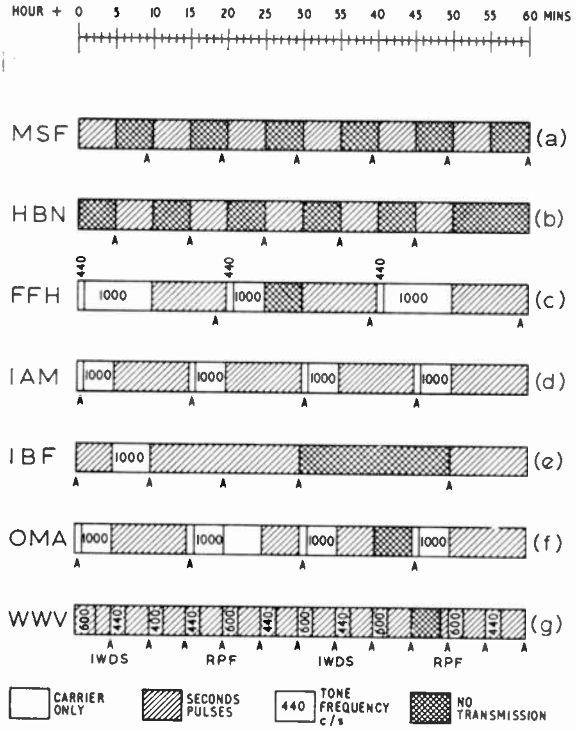


Fig. 19.14. Modulation schedules of standard frequency stations. FFH (Paris) transmits on 2·5 Mc/s from 08.00-16.30 UT on Tuesdays and Fridays, IAM (Rome) on 5 Mc/s from 07.30-08.30 UT and IBF (Turin) on 5 Mc/s from 06.50-07.30 UT and 10.50-11.30 UT Mondays to Saturdays, HBN (Neuchatel) on 5 Mc/s, MSF (Rugby) on 2·5, 5 and 10 Mc/s and OMA (Prague) on 2·5 Mc/s are in continuous operation. MSF and HBN operate on a time-sharing basis on 5 Mc/s but are silent from minutes 55-60 in each hour to permit reception of time signals from RWM-RES (Moscow) at every even hour (UT) during the day.

transmissions from MSF at Rugby on 2·5, 5 and 10 Mc/s while the BBC transmitter at Droitwich on 200 kc/s is also maintained at a very accurate frequency. Similar services in the USA are provided by WWV on 2.5, 5, 10, 15, 20 and 25 Mc/s and some of these signals are normally receivable in the UK. WWVH in Hawaii operates on 2·5, 5, 10, and 15 Mc/s.

The transmissions from MSF on 5 Mc/s are on a time-sharing basis with HBN (Neuchatel) in accordance with the schedule shown in **Fig. 19.14** which also shows the hourly schedules of other standard frequency transmissions useful in Europe. From 0–5 minutes past each hour, MSF transmits carrier and seconds pulses, from 5–9½ minutes past each hour there is no transmission and from 9½–10 minutes past each hour, MSF transmits its call-sign and the amount of the frequency offset (in parts in 10^{10}), each given three times in slow Morse. The cycle is repeated six times in each hour. MSF also transmits on 60 kc/s from 14.29 to 15.30 UT daily.

Absorption Wavemeters

An absorption wavemeter consists simply of a calibrated tuned circuit which absorbs power from the circuit being measured when the circuits are tuned to the same frequency. The power collected by the wavemeter can be made to light a small lamp or operate a sensitive meter. With the wavemeter resonated to the circuit to which it is coupled some energy is absorbed from that circuit. If the wavemeter is held close enough the r.f. current induced into it will be sufficient to light the bulb. Thus, provided that the wavemeter has been previously calibrated, it is only a matter of tuning for a resonance indication on the bulb and reading off the frequency.

A low power stage may not be capable of providing enough r.f. power to light the bulb. Under such circumstances resonance indication can be obtained from the anode current of the low power stage which will rise when the wavemeter absorbs energy or a dip in the grid current of the next stage.

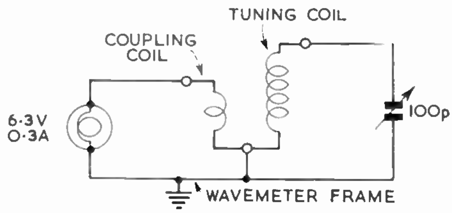
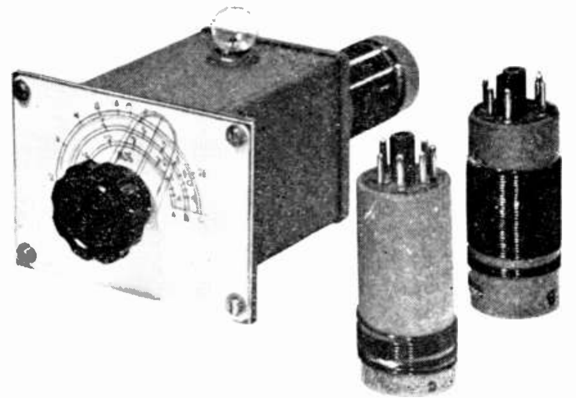


Fig. 19.15. Absorption wavemeter for 1·5-30 Mc/s.

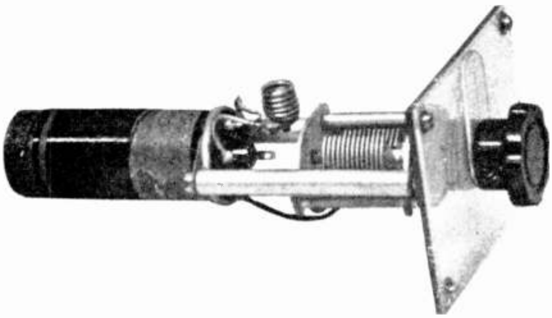
ABSORPTION WAVEMETER FOR 1·5-30 Mc/s

The circuit of a simple absorption wavemeter for 1·5-30 Mc/s is shown in **Fig. 19.15.** It will be seen that the wavemeter frame is connected to one side of the tuning coil, coupling coil, and tuning capacitor. An ordinary 6 V. 0·3 A. bulb is used to indicate resonance. A lower consumption bulb would provide a more accurate indication, but the type specified is more robust and will withstand greater overloads, while showing a sufficiently sharp resonance point for all practical purposes.

The tuning capacitor is mounted directly on the front panel, while the tuning coil holder (an octal valve socket) is off-mounted from the front panel by the use of two tapped aluminium distance pieces. Two stiff lengths of copper wire attached to the coil socket support the bulb holder.



View of the wavemeter of Fig. 19.15 showing the scale, cursor, resonance indicator lamp and spare coils.



Internal construction of the absorption wavemeter.

This is better than a more rigid method of mounting, because the location of the bulb with reference to the metal capacitor cover can be easily adjusted. The cover is cut from 20 s.w.g. tin plate, bent to the required shape and soldered.

The coil formers are made from octal valve bases, force-fitted into a bakelized paper tube 2½ in. long by 1⅛ in. in diameter (internal diameter ⅞ in.). After fitting, the tube and base should be cemented together.

TABLE 19.1

Range	Tuning Coil	Coupling Coil	Wire
1·5—4 Mc/s	80 turns	6 turns	32 s.w.g. enam.
4—12 Mc/s	29 turns	3 turns	22 s.w.g. enam.
12—30 Mc/s	6½ turns	2½ turns	22 s.w.g. enam.

The coils should be wound in accordance with the data given in **Table 19.1**. The illustrations show the type of construction.

Calibration

Calibration may be carried out with the aid of an oscillator or a calibrated receiver. As most amateur stations have an accurately calibrated communications receiver, this method will be described.

With the receiver switched on and the aerial connected, a signal is tuned in at the low frequency end of the band to be calibrated. A coupling coil consisting of a few turns of sufficient diameter to slide over the wavemeter tuning coil is connected in series with the aerial The S meter reading should be observed while the wavemeter is slowly tuned: at one point the reading will drop, indicating that signal frequency energy is being absorbed. This point can now be marked on the prepared dial of the wavemeter. The receiver is then tuned to the next signal higher in frequency, and the process is repeated until finally the whole dial is calibrated.

Some receivers are not fitted with an S meter, in which case the b.f.o. should be switched on, an a.c. rectifier type voltmeter being connected to the phones terminals of the set. A signal is then tuned in until the beat note provides a convenient reading on the voltmeter. As before, the wavemeter (coupled to the receiver aerial) should be tuned slowly until a dip occurs in the reading, indicating a calibration point.

When calibration has been completed, the dial may be removed and permanently marked with Indian ink, after which it should be replaced and covered with a protective sheet of 1/16 in. Perspex, held at the corners by 6 BA screws tapped into the front panel. Perspex of the same thickness is also used for the dial cursor, a central hairline being engraved on it with a stylus.

The size of the front panel is 2½ × 3½ in., and the capacitor cover box is 2⅛ × 2⅛ × 3¼ in. The resonance indicator bulb is pushed through a hole drilled in the top of the cover, and is protected from shock by a rubber grommet.

For checking any apparatus which has a power output of a ½ watt or more, the bulb will indicate resonance quite satisfactorily, but where output is low, as in the local oscillator circuit of a receiver, a meter inserted in the grid-leak earth return will dip as the wavemeter is tuned through resonance. Alternatively, a milliammeter may be inserted in the anode circuit of the local oscillator valve, and this will give an increased reading as the wavemeter is tuned through resonance.

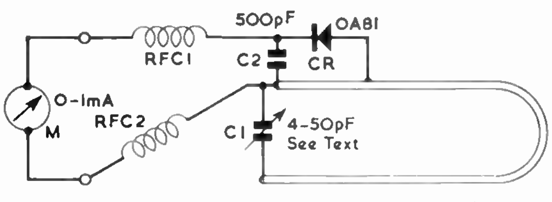
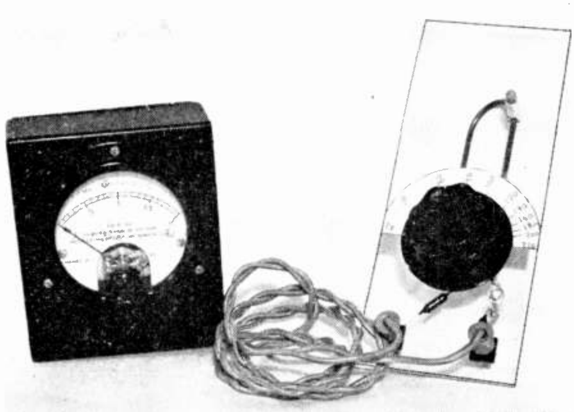


Fig. 19.16. Circuit diagram of the v.h.f. wavemeter. C1 (4—50 pF) is a type C.804 manufactured by Jackson Bros. M should have a f.s.d. of 1-2 mA. RFC1, RFC2 80 turns 40 s.w.g. enamelled wire wound on ½ watt resistor of 1K ohms or more and wax dipped.

SIMPLE ABSORPTION WAVEMETER FOR 65-230 MC/S

The absorption wavemeter circuit shown in **Fig. 19.16** is an easily built unit covering 65 to 230 Mc/s and can therefore be used to check frequencies in Band II (F.m. broadcasting) and Band III (Television) in addition to 70 and 144 Mc/s stages in amateur v.h.f. transmitters. The appearance of the wavemeter with its associated indicating meter can be seen in the photograph.



V.h.f. absorption wavemeter with its indicating meter.

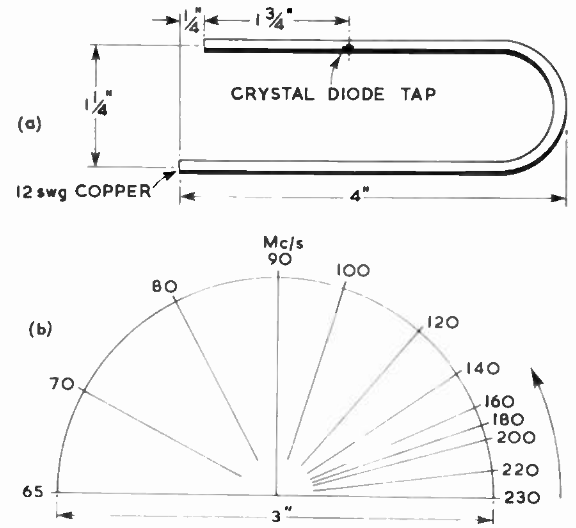


Fig. 19.17(a). Details of the inductance loop made of 12 s.w.g. copper wire. The dimensions should be closely followed if the calibration of (b) is to be used. (b) Dial calibration. The calibration points relative to the base line (anti-clockwise) are; 230 Mc/s-0°: 220 Mc/s—8°: 200 Mc/s—16°: 180 Mc/s—20°: 160 Mc/s—25°: 140 Mc/s—35°: 120 Mc/s—50°; 100 Mc/s—73 : 90 Mc/s—90°: 80 Mc/s—118°: 70 Mc/s—152°: 65 Mc/s—180°.

Construction is straightforward and all the components, apart from the meter are mounted on a Perspex plate measuring $7\frac{1}{2}$ in. $\times$ 3 in. $\times$ $\frac{1}{8}$ in. Details of the tuned circuit are shown in **Fig. 19.17(a)** and should be closely followed. The layout of the other components is not critical provided they are kept away from the inductance loop. To prevent damage to the diode when soldering, a heat shunt should be used.

For accurate calibration a signal generator would be required but provided the inductance loop is carefully constructed and the knob and scale are non-metallic, dial markings can be determined from **Fig. 19.17(b)**. These should be accurate enough for most purposes.

In operation, the unit should be loosely coupled to the tuned circuit under test and the capacitor then tuned until the meter indicates resonance. For low power oscillators a more sensitive meter should be used if available.

The wavemeter can also be used as a field strength indicator when making adjustments to v.h.f. aerial arrays. A single turn coil should be loosely coupled to the wavemeter loop and connected via a low inpedance feeder to a dipole directed towards the aerial under test.

Lecher Lines

Above about 100 Mc/s it is practicable to measure directly the wavelength at which a transmitter or oscillator is operating by using *Lecher lines* which comprise a pair of taut parallel wires, spaced an inch or so apart to form an open wire transmission line, and a bridge to short circuit the wires which can be moved along the line as required.

For transmitter frequency measurement, one end of the line is loosely coupled by a loop to the p.a. circuit. Starting near the coupling loop end of the line, the bridge is slowly moved towards the open end of the line until a point of maximum current in the bridge is found; this will be indicated by a deflection on the anode current meter, or by observing when a flash lamp and loop loosely coupled to the p.a. coil passes through a minimum in brightness. This position should be carefully noted and the bridge then moved further along the line until the next similar position is found. This should be repeated several times until the distance between successive positions is the same. The distance between the two points will be one half of the wavelength at which the transmitter is operating.

Application of the formula

$$\textit{Frequency Mc/s} = \frac{15{,}000}{\textit{Distance between bridge position in cm.}}$$

will enable the frequency of the oscillation to be determined. For example, if the distance between the two bridge positions is found to be 100 cm, then by substituting in the formula,

$$\textit{Frequency} = \frac{15{,}000}{100}\ \textit{Mc/s} = 150\ \textit{Mc/s.}$$

For the most sensitive condition of adjustment, the Lecher lines and the bulb and loop should be very loosely coupled to the tank circuit. This is especially important when measuring the frequency of a self-excited oscillator whose frequency may be altered if the coupling is too tight. Although the accuracy of frequency measurement by this method is not very high the method is of great use to the amateur as the only instrument required is a metre rule. The longer the Lecher lines and the further the two minima are from the coupling loop end, the more accurate will be the reading. An accuracy of 0·1 per cent can be attained with care.

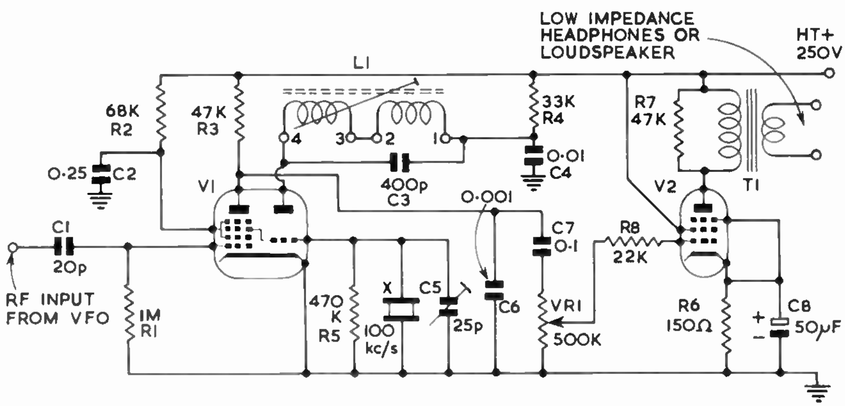


Fig. 19.18. Simple crystal calibrator for addition to a v.f.o. The inductance L1 is an Osmor type OA9. If necessary, the inductance can be increased slightly by squeezing the windings closer together. Final adjustment of the inductance may be made by varying the position of the core. R8 (22 K ohms ½ watt) is wired direct to the valveholder for V2 to act as a parasitic stopper. T1 is a small pentode output transformer. The crystal X is a 100 kc/s unit and details of the circuit should be given when ordering. Suitable valves are V1 ECH81, V2 EF91.

Crystal Calibrator

When using a variable frequency oscillator the absorption wavemeter is invaluable for checking that circuits are tuned to the correct harmonics but it is not capable of giving a sufficiently accurate reading of the actual frequency. It is essential therefore to compare the transmitter against a more accurate frequency standard and the obvious choice is a crystal-controlled oscillator.

CALIBRATION UNIT FOR A V.F.O.

A simple way of adding a crystal-controlled calibrating unit to a v.f.o. is shown in **Fig. 19.18.**

The triode section of V1 operates as a 100 kc/s crystal oscillator with its anode circuit tuned to the crystal frequency. The frequency can be set accurately by adjusting C5 for zero beat against a standard frequency transmission.

R.f. from the v.f.o., or preferably from a buffer stage following the v.f.o., is applied to grid one of the hexode section of V1, in the anode circuit of which the sum and difference frequencies appear. The r.f. components are shorted to earth by C6 while the audio frequency beat note is amplified by V2.

Assuming that the v.f.o. tunes from 3·5 to 3·8 Mc/s care must be taken to ensure that the lowest frequency giving a strong beat note with harmonics of the crystal frequency really is 3·5 Mc/s and not 3·4 or 3·6 Mc/s. This should be established by feeding a weak signal to an accurately calibrated receiver.

Strong beat notes will be obtained at 3·5, 3·6. 3·7 and 3·8 Mc/s, but there will be weaker responses at 3·55, 3·65 Mc/s, etc. due to the second harmonics of the v.f.o. Similarly, there will be still weaker beats at 3·533, 3·567, 3·633 Mc/s, etc. due to the third harmonics of the v.f.o.

10/100 KC/S CRYSTAL CALIBRATOR

One of the simplest frequency standards for use in the amateur station is a 100 kc/s crystal oscillator, particularly if it is employed in conjunction with a multivibrator to provide calibration markers every 10 kc/s. The circuit diagram of an easily built unit of this type is shown in **Fig. 19.19.** Accuracy is assured by the use of the 100 pF capacitor in parallel with the crystal which permits the second harmonic of the oscillator frequency to be adjusted precisely to the frequency of the BBC transmission on 200 kc/s.

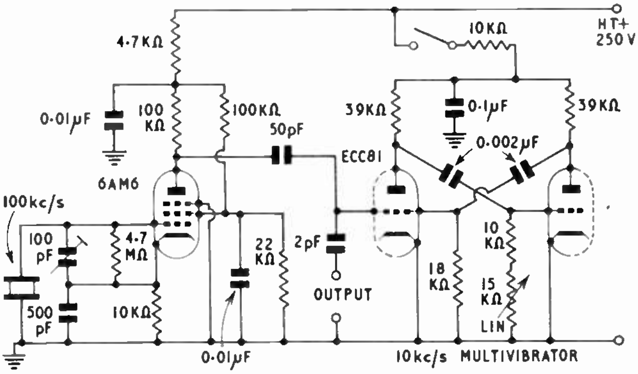


Fig. 19.19. Crystal calibrator giving frequency markers over 10 kc/s and 100 kc/s. An EF91 valve may be used in place of the 6AM6, and a 12AU7 for the ECC81.

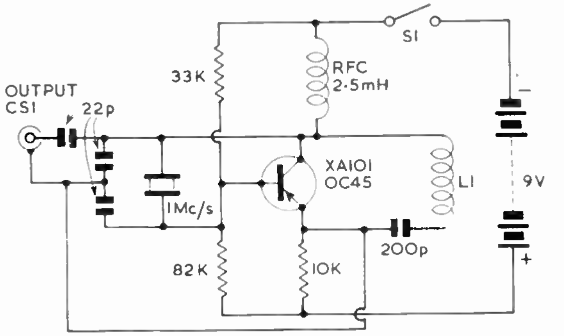
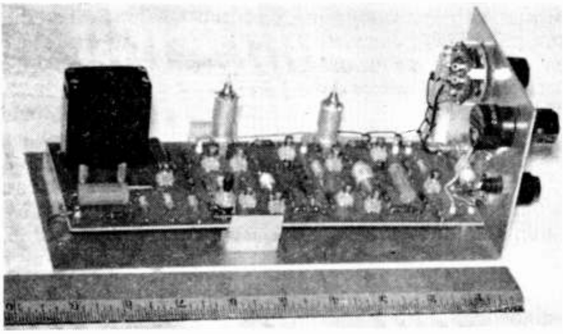


Fig. 19.20. Transistor 1 Mc/s crystal oscillator. The capacitors should all be silvered mica type. L1 is 80 turns 32 s.w.g. pile wound on a ¼ in. diameter former.

TRANSISTOR 1 MC/S CRYSTAL MARKER

A crystal controlled oscillator providing signals every megacycle to well above 30 Mc/s is shown in **Fig. 19.20** and comprises a single transistor, a 1 Mc/s crystal and a few small components. Its frequency can be checked by comparison with the 5 Mc/s transmissions from MSF or WWV.



Top view of G3LRQ Crystal Calibrator

Transistorized Crystal Calibrator

The crystal calibrator shown above is a more elaborate device using a 1 Mc/s crystal followed by dividers to give 100 kc/s and 10 kc/s signals. It can be used to measure frequencies of received as well as transmitted signals with a high degree of accuracy. The 1 Mc/s, 100 kc/s or 10 kc/s ranges may be selected by the switch and passed through an emitter follower to provide a low impedance output.

Referring to the circuit of **Fig. 19.21**, TR1 is a Pierce-type crystal oscillator, the output of which is fed into the base of TR2. TR2 acts as a switch to provide trigger pulses for the first divider circuit. The 1 Mc/s output for the emitter-follower is taken from the collector of TR2.

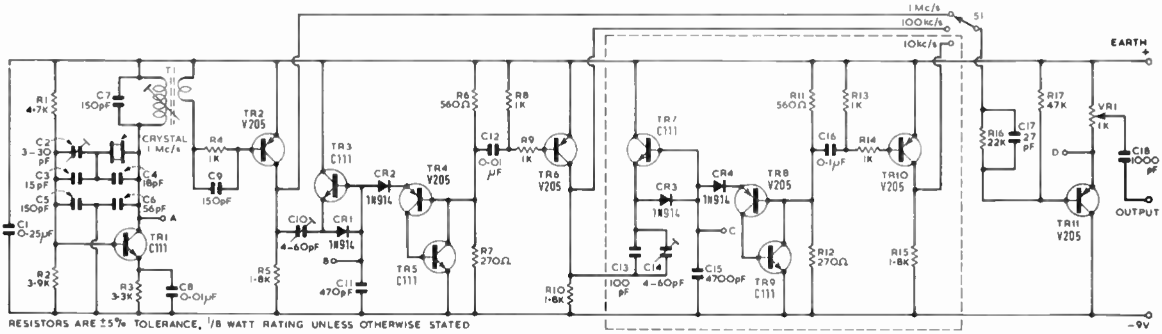


Fig. 19.21. The G3LRQ calibrator circuit. The resistors may all be ½ watt at 5 per cent tolerance. C3, C4 and C17 should be ± 2 pF, while C5, 6, 7, 9, 11 and 13 are ± 2 per cent. C15 and C18 are ± 5 per cent. VR1 is 1 K ohms non-inductive. T1, primary, 130 turns, 38 s.w.g. s.s.e. pile wound. Secondary, 65 turns, 38 s.w.g. s.s.e. interwound with the last 65 turns of the primary on a 3/16 in. diam. former with a dust iron core.

The divider circuits are of the step-counter type, and operate in the following manner. CR1 conducts on the positive pulses from TR2, and C11 charges up a small amount, determined essentially by the ratio of C10 to C11, but this pulse is, however, insufficient to make CR2 conduct. The positive pulse also appears at TR3 base, and TR3 goes into conduction, discharging C10 in readiness for the next pulse. On the following pulse, C11 is charged up further, this new pulse being added to the one already stored. Then, dependent on the ratio of R6 to R7, when C11 has received a small additional charge several times, as shown in **Fig. 19.22**, CR2 will conduct, discharging C11 and operating TR4 and TR5 which act as a complementary high-speed switch producing a negative output pulse at the junction of R6, R7.

This output pulse is fed to the base of TR6, which acts as a switch in the same manner as TR2, the 100 kc/s output for the emitter-follower being taken from the collector of TR6.

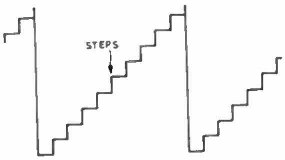


Fig. 19.22. Waveform on C11.

Trigger pulses for the 10 kc/s divider are also taken from the collector of TR6, the step-counter comprising TR7, TR8 and TR9, operating in the same way as the 100 kc/s version. Negative-going 10 kc/s pulses are obtained from the junction of R11, R12 and, these are fed to TR10, a switch, in order that the same level of signal as the 1 Mc/s and 100 kc/s outputs may be applied to the emitter-follower, TR11. The 10 kc/s signal is taken from the collector of TR10.

A switch, S1, is incorporated in the base circuit of the emitter-follower, TR11, to select the desired output. The emitter-follower isolates the output from the frequency control circuits, in order to prevent any load from affecting them. The output is taken from C18, via the potentiometer VR1, which varies its amplitude.

Alignment

Two methods of setting-up the calibrator will be outlined, one using a receiver, and the other employing an oscilloscope.

The **receiver method** is carried out as follows: Connect the output of the calibrator to the aerial terminal of the receiver, from which the aerial should be removed for the entire setting-up procedure, except when adjusting the 1 Mc/s oscillator for zero-beat with a standard frequency transmission, as outlined below.

Switch S1 to the 1 Mc/s position, tune the receiver to a convenient harmonic, e.g., 2 Mc/s, and locate the marker with the aid of the b.f.o. Then, switch off the b.f.o., and the carrier should easily be heard. With the receiver a.g.c. also off, adjust T1 until this carrier has reached its peak amplitude; there should be a maximum within the adjustment range of the core of T1. If the receiver has an S meter, this task is made considerably easier. The frequency of the 1 Mc/s oscillator can be altered slightly by the adjustment of C2 to obtain zero beat when listening to a suitable harmonic beating with a standard frequency transmission, such as MSF on 5 Mc/s. The 1 Mc/s oscillator circuit is then operating properly.

With the switch S1 still in its 1 Mc/s position, turn the b.f.o. on once more, and locate the next harmonic, noting the two readings on the receiver tuning dial (e.g., 2 and 3 Mc/s). Turn S1 to its 100 kc/s position, and, tuning between the two selected harmonics, there should be heard a series of markers. Adjust C10 with an insulated trimming tool until nine markers are heard between the two 1 Mc/s points. This adjustment will require a certain amount of care, but should not be found too critical. When nine markers are heard between the 1 Mc/s points, the 100 kc/s divider is correctly adjusted.

The setting-up procedure for the 10 kc/s divider is identical to that for the 100 kc/s divider except that two adjacent 100 kc/s points are selected, and C14 must be adjusted until nine markers are heard between these two points.

The **oscilloscope method** lends itself readily to setting up a unit of this nature. A fairly low-capacitance probe should be used (of the order of 12 pF) to prevent high oscilloscope capacities from introducing errors during alignment. The position of S1 is not important with this process.

Attach the probe to the collector of TR1 (point *A* in Fig.

19.21) and adjust T1 for maximum waveform amplitude. Adjustment for zero-beat with a standard frequency transmission should be carried out as detailed in the receiver method.

With the probe attached to the " live " side of C11 (Point *B*, Fig. 19.21), a waveform similar to that shown in Fig. 19.22 should be obtained, and C10 should be adjusted until a display with ten steps is achieved.

The probe should then be moved to the " live " side of C15 (point *C*, Fig. 19.21), and C14 should be adjusted until a further 10 steps are obtained on the oscilloscope trace. This completes the adjustments to the 100 kc/s and 10 kc/s dividers.

The output at the emitter of TR11 (point *D*, Fig. 19.21) may be observed, and switching S1 should produce output pulses of 1 Mc/s, 100 kc/s or 10 kc/s p.r.f. depending on its setting.

Calibrator without 10 kc/s Divider

Some economy may be achieved by dispensing with the 10 kc/s divider, the calibrator then only giving outputs of 1 Mc/s, with 100 kc/s marker points. Such a unit would be quite adequate for band-edge marking Top Band and 80m, and on other bands careful interpolation between 100 kc/s points on the receiver tuning dial would give approximate readings. For example, the top end of the 15m band is 21,450 kc/s. Markers would be heard at 21,400 kc/s and 21,500 kc/s. If the receiver tuning scale were linear between these two points, 21,450 kc/s would lie half way between them.

The 10 kc/s divider can be deleted by leaving out the circuit enclosed by the dotted lines in Fig. 19.21, and the 10 kc/s setting-up procedure can then be ignored.

It is worth bearing in mind that the current consumption for the unit is quite high at 30 mA, so that if it is operated by a battery, it should only be switched on for the period of use to prevent excessive battery current drain.

Heterodyne Frequency Meters

The heterodyne frequency meter consists of an accurately calibrated v.f.o. complete with a crystal calibrating source. In some cases, the calibrated oscillator must be compared with the transmitter v.f.o. using a receiver while in others facilities are provided to produce a beat note between a small r.f. voltage from the transmitter and the calibrated oscillator within the instrument itself. Many amateurs use an excellent American instrument of the latter type, the BC221.

AMATEUR BANDS FREQUENCY METER

The circuit of a home-built heterodyne frequency meter is shown in **Fig. 19.23**. It achieves a high degree of accuracy by using a 100 kc/s crystal sub-standard to check a variable oscillator by interpolating between successive harmonics of the crystal in the ranges covered. The crystal itself may be compared with standard frequency transmissions such as those from MSF at Rugby or the BBC transmitter on 200 kc/s.

To obtain as open a scale as possible provision is made for the coverage of the amateur bands only in two ranges: 1·75 to 2 Mc/s, the second harmonic of which includes the 3·5 to 3·8 Mc/s allocation, and 7 to 7·5 Mc/s. On the second range the fundamental, second, third and fourth harmonics serve for measurement in the 7, 14, 21 and 28 Mc/s bands.

The output of the frequency meter is taken from a low impedance point and fed directly into the receiver via a length of coaxial cable thus preventing to a large extent the radiation of oscillator harmonics on television frequencies. A simple attenuator enables the output from the instrument to be controlled over a wide range. Either a c.w. or m.c.w. signal is available.

The addition of a further oscillator in the circuit provides a stable source of oscillations at approximately 1 Mc/s for aid in the preliminary calibration of the meter or for rapid checking of an uncalibrated receiver.

The variable capacitor C18 in series with the 100 kc/s crystal is for setting its frequency against a standard frequency transmission, and is mounted on the chassis inside the case. If it is desired to control the frequency of the crystal to very fine limits it is suggested that either C18 be mounted on the rear drop of the chassis and a hole cut in the back of the cabinet for access or a small additional 10 pF variable capacitor be similarly mounted and connected in parallel with it.

The purposes of V2 is to make audible, and V3 to amplify, the beats between the crystal oscillator harmonics and the variable oscillator when checking the calibration of the latter throughout its ranges or to detect the beat note between an external oscillator and the variable oscillator as is necessary when measuring the frequency of a transmitter. The output from V2 is taken across its unbypassed cathode resistor by way of an ordinary 500–1000 ohm carbon track potentiometer. This gives a good measure of control of the signal fed to the receiver and, as the source is of low impedance, the loss in the coaxial lead is small. To avoid loading the receiver input circuit excessively, as well as to minimize radiation of the frequency meter harmonics from the aerial, it is recommended that a carbon resistor of 1000 ohms or so be connected between the centre conductor of the coaxial line and the aerial terminal of the receiver.

It is often difficult to pick out a frequency meter signal from other strong carriers on a band and it is a definite advantage, therefore, to be able to modulate the local signal if required. In this design modulation is introduced by employing V3 as an a.f. oscillator in conjunction with the coupling transformer between its grid and the anode of V2. The switch S2, cuts the headphone jack out of circuit when modulation is applied.

Good mechanical stability is an essential feature of a frequency meter which is to have any pretence to accuracy. The dimensions of the chassis are 14 in. × 7 in. × 3 in. deep; and the general layout is shown in **Fig. 19.24**. All connections are carried out in push-back wire and extensive use is made of tie points for mounting the smaller parts to prevent vibration.

No matter how well made a frequency meter may be it is next to useless without a dial which can be read clearly to the necessary degree of accuracy. To this end a Muirhead component similar to that used on the old National HRO receiver is used. This dial provides 500 divisions approximately ¼ in. apart with an effective scale length of 12 ft. 6 in.; it is to all intents and purposes free from backlash.

Alignment

Connect a coaxial lead between the frequency meter and a receiver tuned to the 1·8 Mc/s band and check that harmonics of the crystal are audible. If nothing is heard first make sure that C18 is not set to too low a capacity. Then, with the bandswitch on Range 1 and a pair of high resistance 'phones plugged into the frequency meter, make sure that V1 is oscillating by noting the beats between the variable and the crystal oscillators as the tuning is varied. Adjust C2 so that 2 Mc/s is reached at almost minimum capacity and coincides with one of the crystal harmonics. The ZERO SET capacitor, C5 should be at half capacity. Three crystal harmonics lower should bring the variable oscillator to 1·7 Mc/s and this should be checked on the receiver. If the calibration of the receiver is not known to within 100 kc/s the 2 Mc/s point will have to be found with the aid of the 1 Mc/s oscillator.

Switch next to Range 2 and locate 7 Mc/s on the receiver by reference to the amateur band; with the tuning dial at

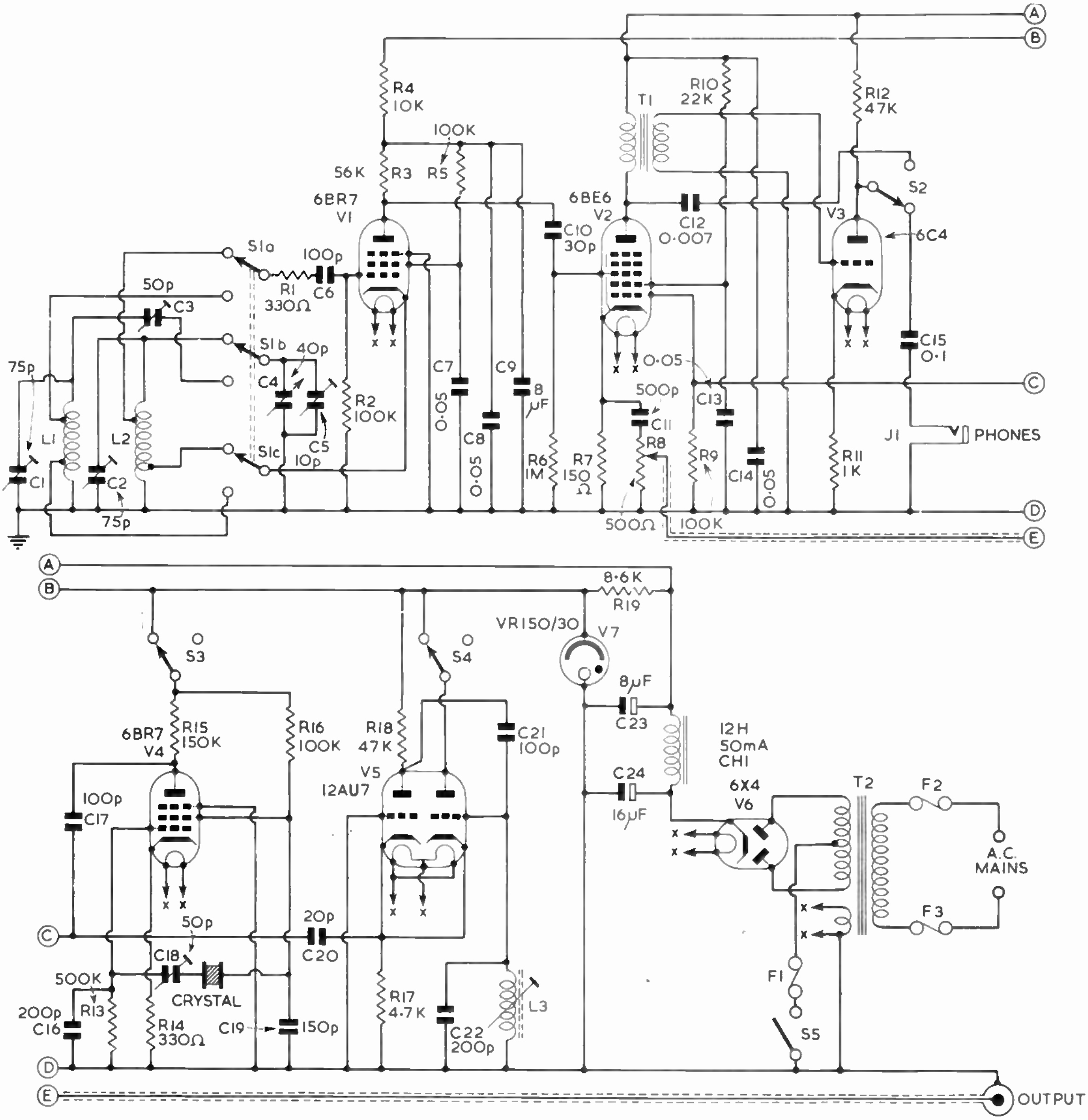


Fig. 19.23. Circuit diagram of the amateur bands heterodyne frequency meter. L1, 16 turns 24 s.w.g. 1 in. dia., ¾ in. long, cathode tap 2 turns, grid tap 11 turns from earthy end: L2, 67 turns 26 s.w.g. enam. 1 in. dia. 1¼ in. long, cathode tap 5 turns, grid tap 41 turns from earthy end: L3, Osmor type QO5. The h.t. line marked B should be decoupled to earth with a 0·05 μF capacitor. The range switch S1 is shown in the Range 1 position. T2, 250-0-250V 60 mA, 6·3V 3A.

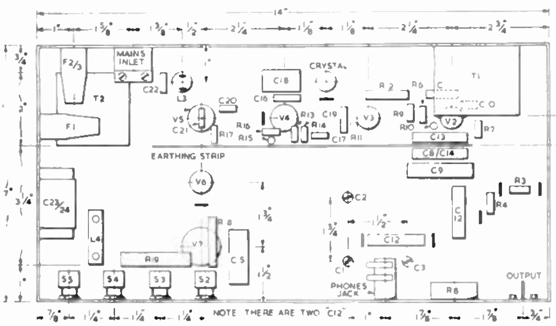


Fig. 19.24. Layout of the principal components beneath the chassis of the heterodyne frequency meter.

about 20° and C3 at about 75 per cent of its maximum capacity, set the variable oscillator to this point by adjustment of C1. To cover the 28 Mc/s band fully the variable oscillator must tune to 7·5 Mc/s and this should occur at about 490° on the dial. Some juggling with the capacities of C2 and C3 will be called for before the required range " fits " the dial, the operation being similar to ganging a receiver. C1 is adjusted at the *high* frequency end and C3 at the *low*.

Calibration

First, the crystal oscillator should be checked against one of the standard frequency transmissions, MSF on 5 Mc/s being convenient. At the same time the 1 Mc/s oscillator can be adjusted, its tuning range (by variation of the iron core of L3) being insufficient for a mistake to be made on the fifth harmonic. Should the receiver cover only the amateur bands the accuracy of the crystal may be checked by reference to the 200 kc/s transmitter on a broadcast receiver, C18 being carefully adjusted until the quality of speech and music are barely affected.

On Range 1 note the dial readings for zero beat for 1·7, 1·8, 1·9 and 2 Mc/s and either draw a graph or record them in a book.

Calibration of Range 2 should be carried out in a similar manner between 7 and 7·5 Mc/s and will be possible to determine the 50 kc/s points without difficulty. Interpolation between these points may be considered to be linear for most purposes but if the user wishes to ascertain the 25 kc/s positions exactly the receiver should be tuned to the 28 to 29·7 Mc/s band where the variable oscillator appears to tune four times as fast as on the fundamental and therefore each coincidence between the oscillator and the 100 kc/s points will represent a movement of 25 kc/s at the fundamental frequency.

It will be found that some slight changes take place from time to time in the frequency of the variable oscillator and these should be corrected before readings are taken by setting the dial for the previously ascertained reading for the 100 kc/s point nearest to the frequency it is required to measure and setting the oscillator stage (V1) accurately to zero beat by means of capacitor C5.

The tuning rate on Range 1, assuming coverage from 1·7 to 3 Mc/s, is 1¼ divisions per kc/s and on the 28 Mc/s band nearly 4 kc/s per division. On the latter range, if full coverage to 29·7 Mc/s is not required, the tuning rate may be considerably improved by setting C3 for greater bandspread.

If a modulated note cannot be obtained the connections to either the primary or secondary of T1 should be reversed. The modulation frequency depends largely upon the capacity of C12. The 0·007 μF capacitance used in the original model was obtained by series connection of 0·05 and 0·01 μF capacitors, but this would not be necessary if a satisfactory note were forthcoming with the aid of a single component. T1 is not critical and almost any transformer with a ratio of between 3·5 and 5 to 1 should be satisfactory. A change from these ratios would almost certainly require a different value of C12.

Use

To measure the frequency of an external oscillator a short wire probe may be necessary and should be plugged into either the end of the coaxial cable or the output socket so that the incoming signal can be heterodyned by the variable oscillator. The latter is corrected, if necessary, at the nearest crystal harmonic to the unknown frequency but when actually taking a reading both the crystal and 1 Mc/s oscillators should be off.

FREQUENCY MARKERS FOR THE 144 MC/S BAND

The normal crystal calibrator is of limited use on the v.h.f. bands because its output is very small at v.h.f. and also because its h.f. output will be received strongly by the tunable i.f. which forms part of most v.h.f. receivers. To overcome this problem, it is necessary first to generate the higher order harmonics of the markers and then amplify them selectively in the required band. A suitable design is shown in Fig. 19.25. The 1 Mc/s oscillator and its buffer amplifier are transistorized and built into a well-lagged tin box remote from the rest of the calibrator. After the 6AK5 amplifier, the 1 Mc/s signal is fed into the 100 kc/s and 10 kc/s multivibrators.

This feeds the harmonic generator which uses two diodes, any point contact types being suitable. The output from the diodes is fed to an EF91 (6AM6) which is an amplifier with its anode circuit tuned to 145 Mc/s. The calibrator output is taken from a link on this coil via a coaxial cable to the converter. A triode-pentode such as an ECF82 (6U8) could be used instead of the 6J6/EF91 combination if preferred.

The harmonic generator is a high impedance device and is thus rather prone to picking up other frequencies from receiver local oscillators, etc., which give rise to many spurious beats in the 2m band. It is essential, therefore, to screen the two diodes and the associated resistor and capacitor in a small can and to keep the leads short and close to the chassis. The output coil and tuning capacitor in the anode circuit of the EF91 (6AM6) are also in a screening can to eliminate stray pick-up. These precautions result in only a very few spurious signals being found between 144 and 146 Mc/s, and these are weaker than the 10 kc/s markers.

The rest of the calibrator is quite conventional except that the 1 Mc/s oscillator is transistorized and encased in a well lagged tin remote from the rest of the equipment. A buffer amplifier transistor is also incorporated in the tin and the unit is coupled by coaxial cable to the rest of the calibrator. The 1 Mc/s crystal oscillator is adjusted to zero beat with MSF or WWV and enables frequencies to be measured on 2m to an accuracy of $\pm$500 c/s.

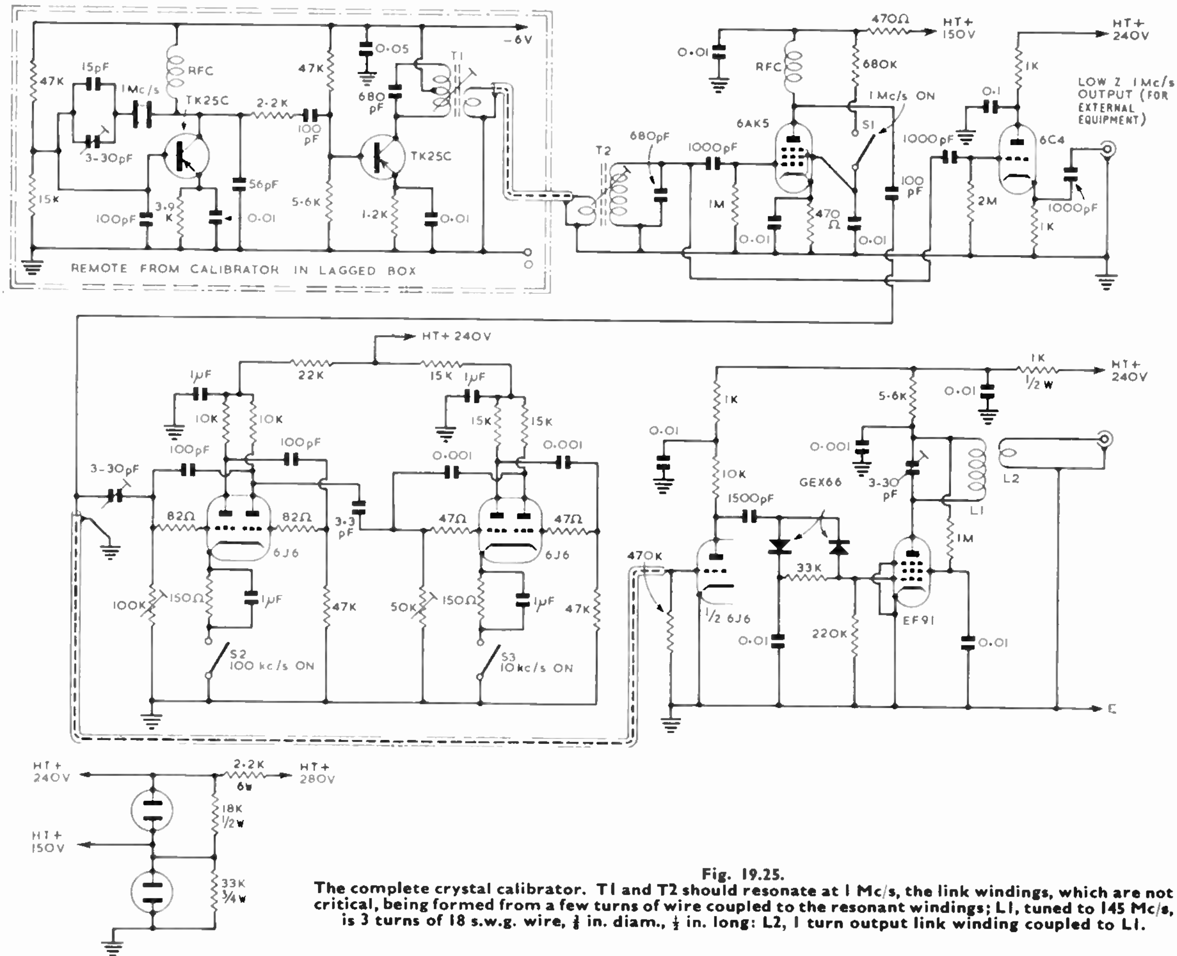


Fig. 19.25.
The complete crystal calibrator. T1 and T2 should resonate at 1 Mc/s, the link windings, which are not critical, being formed from a few turns of wire coupled to the resonant windings; L1, tuned to 145 Mc/s, is 3 turns of 18 s.w.g. wire, $\frac{3}{8}$ in. diam., $\frac{1}{2}$ in. long: L2, 1 turn output link winding coupled to L1.

If an output is required for the lower frequency bands it could be taken from the anode of the 6J6 amplifier at high impedance or via a cathode follower if a low impedance feed is required.

Grid Dip Oscillators

Another type of frequency measuring device which is extremely useful is the grid dip oscillator. This is really nothing more than a calibrated oscillator which can be tuned over a wide range of frequencies and which has a moving coil meter indicating the grid current. If it is coupled to an external tuned circuit the grid current meter will dip when the oscillator is tuned to the reasonant frequency of that circuit because the circuit will absorb energy from the oscillator, thus reducing the amount of r.f. feedback to the grid and, in consequence, reducing the grid current.

Since the grid dip oscillator provides its own r.f. energy it does not require the circuit being checked to be energized. It is therefore useful for checking the resonant frequency of tuned circuits, r.f. chokes and aerial systems. The g.d.o. is also useful as an absorption wavemeter, signal monitor or simple signal generator.

A TRANSISTORIZED DIP OSCILLATOR FOR 0·85—150 MC/S

The circuit of a typical dip oscillator which is simple to build is shown in **Fig. 19.26.**

Basically, the circuit comprises a multi-frequency range transistor oscillator, covering 0·85 to 150 Mc/s in seven ranges, using plug-in coils, a diode detector and a transistor d.c. amplifier operating a meter. The unit contains its own 9 volt battery, the " g.d.o." actually running from a 6·8 volt Zener-stabilized line, an arrangement which helps to reduce the effects of battery voltage variation. The total current consumption is 5 mA. A generous overlap is provided between the frequency ranges, L2, L3, L4 and L7 covering two amateur bands each and L5 three amateur bands (**Table 19.2**).

The oscillator circuit (Fig. 19.35) is a grounded-collector Colpitts, with only part of the tuning capacitance tapped for connection to the emitter. If the 15 pF variable capacitor C1, shown ganged to the 150 pF variable C2 in the circuit diagram, were in fact fixed, the coupling of the transistor input and output circuits to the tuned circuit would vary with rotation of the main 150 pF tuning capacitor, the

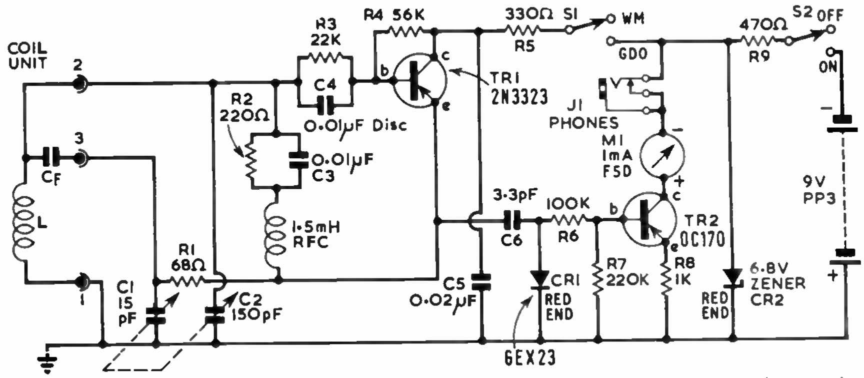


Fig. 19.26. The G3HBW dip oscillator covering 0·85-150 Mc/s. The r.f. transistor TR1 is an inexpensive v.h.f. type manufactured by Motorola. Alternatively, the AF102, AF118, AF139, AFZ12, 2N1742, GMO290 or GMO378 may be used. S1 is provided to enable the dip oscillator to be used as a sensitive wavemeter. CR1 may be a GEX23, GEX54, OA70, OA71 or similar diode. CR2 may be any small 6·8V Zener diode.

effective positive feedback falling as the value of the tuning capacitance was increased. The combination of the stray capacitances in the circuit with the ganged 15 pF variable capacitor results in almost constant feedback being obtained over all ranges except the highest frequency one, where the feedback mechanism is somewhat different. The required value of the capacitance C_F varies with the frequency range in use and so the appropriate value of C_F is built into each coil range. Increasing the value of C_F reduces the feedback ratio.

The reactance of the 3·3 pF coupling capacitor C6 and the essentially resistive impedance of the diode and its load, form a potentiometer which delivers a nearly constant d.c. to the base of TR2, except on the highest frequency range, where the meter reading is reduced to about one third of full scale deflection. The coil is used as part of the d.c. return circuit for the oscillator transistor emitter current to avoid shunting the greater part of the tuned circuit with the 1·5 mH r.f. choke.

Provision is made to remove the d.c. supply from the oscillator transistor when required so that the g.d.o. may be used as a sensitive wavemeter and also as monitor by employing the 'phones jack in the d.c. amplifier collector.

The battery is mounted in a small clamp inside the lid and connected to the g.d.o. proper by means of flying leads. A normal toggle switch is used for the main on-off function and a small slide switch to select either GDO or WAVEMETER operation. This was done so that there would be no confusion as to whether the instrument was switched off or not, which might have occurred if two similar switches had been used. Any type of insulated jack socket may be utilized, provided that it is of the shorting variety.

Construction

The dip oscillator is built into a small 18 s.w.g. aluminium box, provided with a close-fitting, flanged lid, the box and lid being available ready made.

The coils for each range plug into a socket on one end of the box. Ordinary three-pin battery plugs and sockets are used, the plug being glued, and then forced into a length of Paxolin tubing of suitable diameter on which the coil is wound.

First, the Perspex dial cursor and aluminium battery clamp should be made according to Fig. 19.27 (a) and (b) respectively. It is necessary to be very careful when cutting and drilling the $\frac{1}{16}$ in. thick Perspex sheet. A hand drill is to be preferred to a power-operated one to avoid the risk of splitting. The large centre hole should be opened out from a suitable smaller size with a repairman's reamer or a round file. A useful ancillary to be employed when marking the dial may be made from 18 s.w.g. aluminium sheet of the same dimensions as the Perspex cursor. Mark and drill the holes as in the cursor proper, using the latter as a marking-out template. Draw the centre line on the aluminium strip along

TABLE 19.2

Coil Winding Details

Coil No	Range Mc/s	No. of turns	S.W.G.	Forner o.d.	Former Length	Winding Length	CF total
1	0·85 to 2·0	180	28	$\frac{7}{8}$ in.	$4\frac{1}{8}$ in.	$3\frac{1}{16}$ in. (close wound)	3 × 1000pF
2	1·8 to 4·0	54	28	$\frac{7}{8}$ in.	$2\frac{1}{2}$ in.	$\frac{7}{8}$ in. (close wound)	2200 + 820pF
3	3·4 to 8·0	27	22	$\frac{7}{8}$ in.	$2\frac{1}{2}$ in.	$\frac{7}{8}$ in.	2200 + 680pF
4	6·7 to 16	13	22	$\frac{7}{8}$ in.	$2\frac{1}{2}$ in.	$\frac{7}{8}$ in.	3300pF
5	13·5 to 34	6	22	$\frac{7}{8}$ in.	$2\frac{1}{2}$ in.	$\frac{7}{8}$ in.	2200pF
6	33 to 85	3	18	wind on $\frac{3}{8}$ in. drill (see Fig. 19.29(d))	shroud $\frac{7}{8}$ in. long	coil $\frac{1}{2}$ in. long	330pF
7	50 to 150	1	18	(See Fig. 19.29(c))			68pF

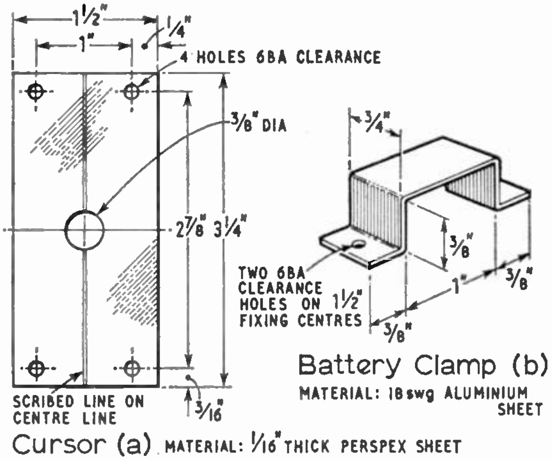


Fig. 19.27(a). The perspex cursor dimensions, and (b), the method of forming the battery clamp.

its greater dimension and then mark out and drill four 1/16 in. diameter holes along this centre line, at distances of 7/16, 5/8, 13/16 and 1 in. from one end. Finally, cut carefully along the centre line, through the four holes and as far as the centre hole, with a fine hacksaw.

When making the battery clamp, preform the inner bends first and then the outer, finally marking and drilling the holes. The holes may now be drilled in the box and lid (see Fig. 19.28). The four holes for fixing the lid should be marked out through the corresponding holes in the box and then drilled and tapped 6 BA.

The shaft of the tuning capacitor is cut off so that only about 5/16 in. of its length is left protruding. Both capacitor trimmers, if fitted, are then opened out to their fullest extent and the capacitor is mounted in the box by means of three 4 BA countersunk head screws, *not more than ½ in. long*, using two 2 BA full nuts on each as spacing washers and screwing tightly into the tapped holes in the capacitor end-plate. *If the screws used are too long, they are liable to damage the vanes of the capacitor beyond repair*. Before the capacitor is bolted home, insert a ⅜ in. long 6 BA cheesehead screw into the lug fixing hole for the slow-motion drive, the head of the screw being inside the box. Next, the operating shaft of the slow-motion drive is cut off so that only about ½ in. of it remains, the drive is fitted over the capacitor shaft and the slotted lug is bolted to the box, using a single 6 BA spacing washer and a 4 BA nut as packing, between the lug and the box. The clamping screws may now be tightened on the capacitor shaft by manipulating a small screwdriver through the hole in the side of the box, which has been provided for the purpose.

A scale disc, 2⅝ in. in diameter is made from thin card or, better, from 1/16 in. thick white plastic sheet, such as Ivorine, and a 5/16 in. diameter hole is made in the centre together with two 8 BA clearance holes ⅝ in. apart, so that the disc may be screwed to the drive flange, using two 8 BA 3/16 in. long, countersunk head screws. The two tagstrip-mounting screws are ¾ in. long and serve also to support the lower end of the Perspex dial cursor, with two 19/32 in. long pieces of 3/16 in. o.d., 21 s.w.g. wall brass or aluminium tubing as packing pieces between the top of the box and the cursor. The other end of the cursor is supported in a similar way using two more 6 BA screws, nuts and pieces of tubing.

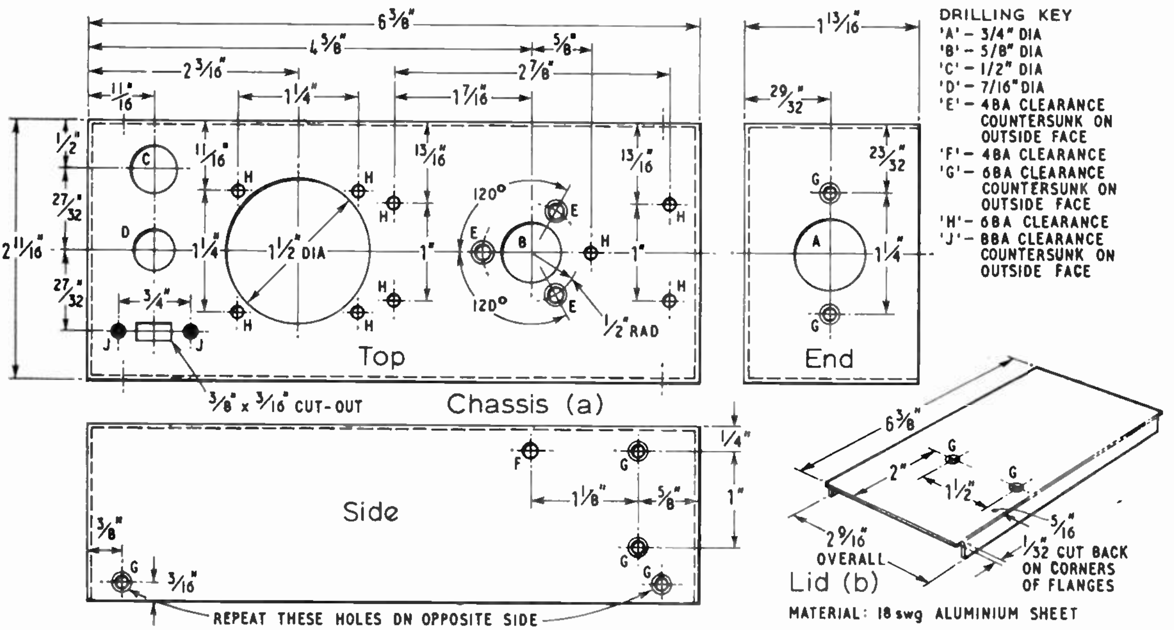


Fig. 19.28. Drilling details suitable for a standard 6⅜ × 2 11/16 × 1 13/16 in. aluminium box. The cursor is spaced above the top of the box, over the slow motion drive which is also mounted on the exterior of the box.

Wiring up the g.d.o. is a simple process. First, the double tagstrip arrangement should be partially completed as a unit inserting R4, R5, R6, C5, C6 and CR1 but not TR1. Then bolt the tagstrips in position in the box. Connect R1, R3, C4, R2, C3, RFC and also the interconnecting wires between the tuning capacitor, coil socket and tag-strips. The parallel pair, R2, C3, should be brought up vertically from coil socket connection 2, between connection 1 and the pin sockets of 2 and 3. The 2N3323 transistor, TR1, may then be soldered into position using a heat shunt to protect the transistor. Next, attach R7, R8 and TR2 to the centre tagstrip and R9 and CR2 to the slide switch. Two unused poles of the d.p.d.t. slide switch are used as anchoring points for R9.

Coil Construction

The general arrangement of the various coils is shown in Fig. 19.29. For the five lowest frequency ranges, covering

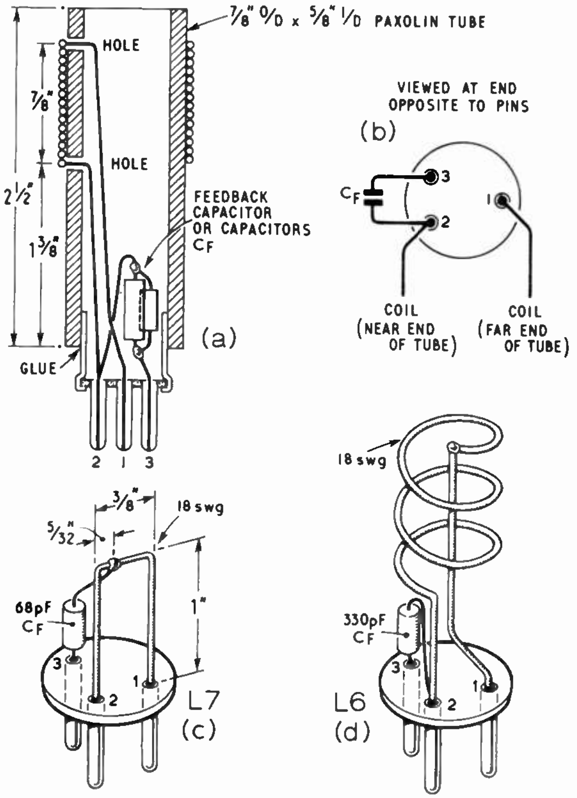


Fig. 19.29. The coil assemblies, the five lower ranges are wound on Paxolin tubes, while ranges 6 and 7 are airwound.

0·85 to 34 Mc/s, the form of construction shown in Fig. 19.29(a) is adopted. The v.h.f. coils, L6 and L7 are made self-supporting, as shown in Figs. 19.29(d) and 19.29(c) respectively. In all cases, the feedback capacitor (or capacitors) is mounted on the plug base, its wires being soldered into the pins.

First, cut one 4⅛ in. length and four 2½ in. lengths of ⅞ in. o.d., ⅛ in. thick wall Paxolin tubing, filing the ends at right angles and smoothing them off. Take the longest tube and drill 1/16 in. diameter holes 3/16 in. and 3⅜ in. from one end on a line along the length of the tube. Do the same with the four shorter lengths, drilling the holes 5/16 in. and 1¼ in. from one end. Remove the burrs and sharp edges from these holes, both inside and out. Now cut off and file square two lengths of ¾ in. o.d., 1/16 in. thick wall Paxolin tube, both ⅞ in. long (the other size of tubing will do if this is not available). Take the seven aluminium plug-shrouds and, supporting them carefully with a pair of narrow-nosed pliers, cut off the constricted portion with a fine hacksaw, leaving a 3/16 in. length of the ⅝ in. o.d. aluminium tube portion with, of course, the four tabs still attached. Square off the end remote from the tabs and file a small chamfer on the outside at this end to provide a lead, when inserting into the Paxolin tubes. Put a smear of Durofix on the outside of the aluminium tubes and then force each of them into one of the five longer Paxolin tubes, at the end remote from the holes, so that a 1/16 in. length of the aluminium tubes, together with the four tabs, are left protruding. When the glue is dry, the coils may be completed.

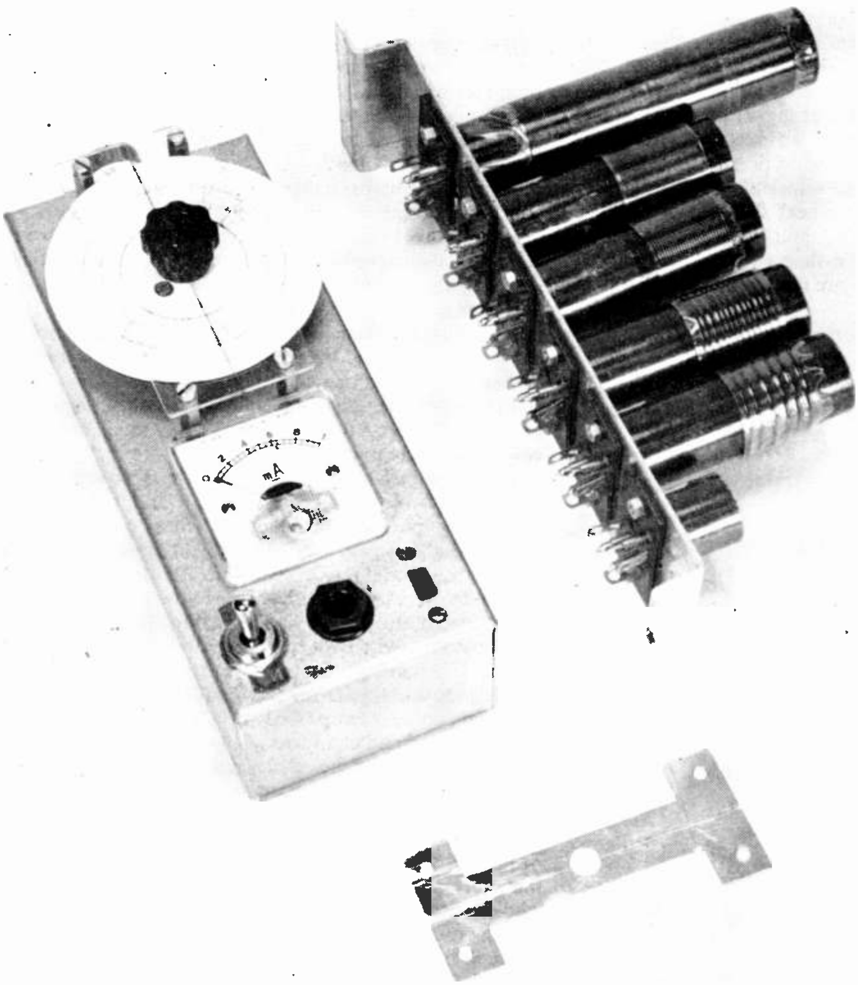
The three lowest-frequency coils are close-wound (Table 19.2). Winding the coils for Ranges 1 and 2 is made easier if a simple procedure is adopted. Start at the plug end, passing the 28 s.w.g. wire through the hole and then temporarily anchoring the free end with Sellotape, hold the plug end of the former in the left hand and wind on about ten turns at a time, with a small spacing. Then, keeping the tension on the wire with the right hand, use the index finger of the left hand to push the turns together. When the coil is completed, pass the free end of the wire through the hole and anchor it temporarily with Sellotape, as at the start. The beginning and end turns of the coil may be secured with adhesive Melinex tape if desired but ordinary Sellotape should not be used for a permanent job, as it is hygroscopic.

The coils for Ranges 4 and 5 should be spaced out after winding to fill the available winding space. Take both ends of the coil down through the plug end of the tube, cut off so that only about 1 in. is left protruding and then bare the whole of this length, sliding an inch or so of loose-fitting sleeving on to each lead. Take the feedback capacitor appropriate to the range (Table 19.2). If two or three are required, parallel them as shown in Fig. 19.29(a). Then pass one capacitor lead into pin 3 of a plug base, cut off and solder in position, leaving the capacitor pointing away from the plug base, as shown. Bend the other capacitor lead over and push into pin 2, but do not solder. Then offer up the plug base to the coil, pushing the coil leads into the appropriate pins, cut off the leads and solder. Test the coil in the g.d.o. before folding over the tabs.

The coil for Range 6 is wound on a ⅜ in. diameter drill or rod as a mandrel. The forms of construction of L6 and L7 are self-evident from Figs. 19.29(d) and 19.29(c). The two short pieces of ¾ in. o.d. Paxolin tubing already prepared are used as protecting shrouds for these two v.h.f. coils. The finished appearance is shown on page 19.22.

Testing and Calibration

The circuit connections should first be checked carefully. If all is well, plug in the Range 1 coil (0·85 to 2·0 Mc/s) and, with the function switch set to GDO and the tuning capacitor at mid-scale, switch on. A meter reading of from 0·5 to 0·7 mA should result. If no meter reading is produced, try the next Range 2 coil. Should success still not be obtained,



Complete transistorized dip oscillator. Coils should be handled with care, preferably at the base, and mounted on a stand as shown to avoid damage which could result in calibration errors.

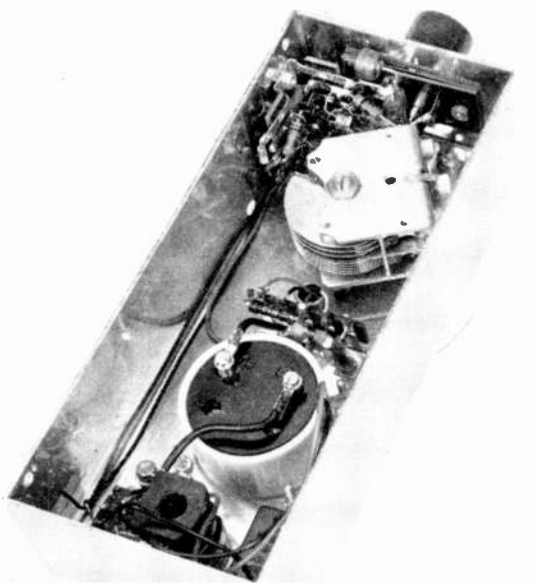
check the polarity of the detector diode CR1, whose red (positive) end should be connected to the chassis. Correct operation of the d.c. amplifier may be ascertained by momentarily bridging the collector and base connections on the TR2 tagstrip with a 470 K ohm resistor, which should give a reading of about 0·4 mA on the meter. If this is successful, listen for the signal from the g.d.o. on a receiver. If nothing is heard, disconnect one end of CR1 and listen again, as a diode with very poor reverse characteristics may prevent oscillation. Using methods such as these, the fault should be localized.

When all the ranges have been made to oscillate, check the frequency coverage, which should not differ by more than a few per cent from the frequencies quoted in Table 19.2. If the frequencies all seem too low, particularly at the h.f. ends, the trimmers on the tuning capacitor have probably not been unscrewed. It will usually be found easier to break the strip leads to the trimmers with a small pair of pliers, or unsolder them carefully.

The greatest meter deflection is usually obtained near the h.f. end of Range 3 or 4. This should be almost full scale. If the deflection is too large or too small, it may be adjusted by reducing or increasing, respectively, the value of R7.

It may, perhaps, be found that some coil ranges will not oscillate over the whole of the tuning capacitor travel; usually the h.f. ends are affected. If only one or two ranges are defective, the value of the built-in feedback capacitor should be reduced by about 5 or 10 per cent. However, if several ranges are unsatisfactory, a 2N3323 transistor of exceptionally low gain may have been used for TR1 and, in this case, it should either be changed or the value of the negative feedback resistor, R1, should be reduced to 56 ohms or even lower, until the trouble is cured.

Check all the coil Ranges 1 to 7 in the oscillator, tuning right round on each range. The indicated currents should vary smoothly across each range, without spurious dips or peaks. It will be found possible on any particular range to use the self-resonance of the coil two ranges lower to check



The underside of the transistor dip oscillator.

the g.d.o. action. Oscillation should be completely stopped at resonance, with only moderately close coupling between the coils. The meter deflection on Range 7 becomes rather small below about 60 Mc/s, but only the upper portion is required, in any case, as the Range 6 coil covers the frequencies below 85 Mc/s.

The g.d.o. is now ready for calibration. Four concentric circles are drawn on the dial, using differently coloured ball-point pens. The radii of the four circles are, respectively, $\frac{5}{8}$, $\frac{13}{16}$, 1 and 1 $\frac{3}{16}$ in. The capacitor is then completely unmeshed and a line is drawn along the diameter of the disc.

The easiest method of calibration is to listen for the g.d.o. signal on a general coverage receiver. Above 30 Mc/s, a 2 or 4m receiver may be used and, by listening to harmonics of the g.d.o., the sub-multiple frequencies as well may be checked. A TV or f.m. receiver will provide further calibration points. Simple absorption wavemeters may be employed to check that spurious signals are not being received.

Both the upper and lower halves of the dial are calibrated, reversing the dial-marking template to suit. It is best to mark the Range 1 and 2 calibrations on the outside of the $\frac{5}{8}$ in. radius circle, Ranges 3 and 4 on the outside of the $\frac{13}{16}$ in. radius circle and so on up to Range 7, leaving one space on the outside circle as a " spare." An " H " pencil is probably the best wiring instrument for this job.

Grid Dip Oscillator for the 430 Mc/s Band

Many designs exist for grid dip oscillators using plug-in coils and covering all bands from 1·5 Mc/s to 144 Mc/s but it is not practicable to extend such designs much beyond 200 Mc/s so that a separate instrument is required for the 430 Mc/s band. The oscillator shown in **Fig. 19.30** uses an A2521 in a Colpitts circuit with the inductance connected directly across a split-stator capacitor. The instrument covers from 400-450 Mc/s.

The inductance, which is also the pick-up loop, consists of a hairpin of copper $\frac{3}{16}$ in. wide and $\frac{7}{8}$ in. long; if the oscillator is contained in a closed box, about $\frac{1}{2}$ in. of the loop should project through a clearance hole about $\frac{3}{4}$ in. diameter. There is usually no difficulty in coupling this small loop to the circuits to be measured but it is possible to use a low impedance line with a small loop at each end if the grid dip oscillator cannot be brought close to the unknown circuit.

The oscillator is constructed on an aluminium chassis measuring 4½ in. × 3¼ in. × 3 in. The layout can be seen from the photograph and is not critical provided that leads are kept reasonably short. The oscillator requires an h.t. supply of about 150 volts at 8 mA and the grid current, which is measured on an external meter, is about 0·5 mA. If any dips in the grid current are found as the oscillator is tuned over the band, this can usually be reduced by modifying one or more of the chokes.

Using the G.D.O.

Determination of the Resonant Frequency of a Tuned Circuit. The resonant frequency of a tuned circuit is found by placing the g.d.o. coil close to that of the circuit and tuning for resonance. No power should be applied to the circuit under test and the coupling should be as loose as possible consistent with a reasonable dip being obtained on the indicating meter. The size of the dip is proportional to the Q of the circuit under test, a circuit having a high Q producing a more pronounced dip than one having only a low or moderate Q.

Absorption Wavemeter By switching off the h.t. and coupling the g.d.o. in the usual manner, the instrument may be used as an absorption wavemeter. In this case power has

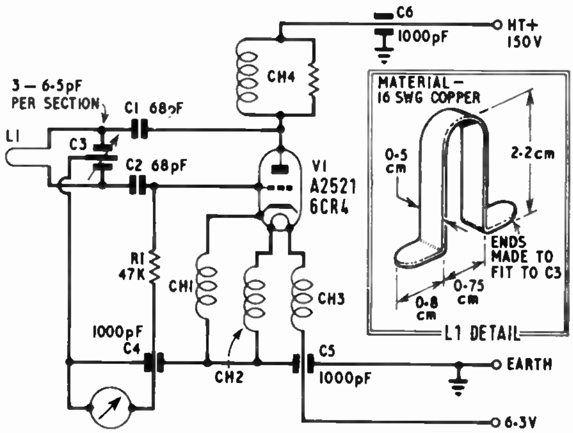
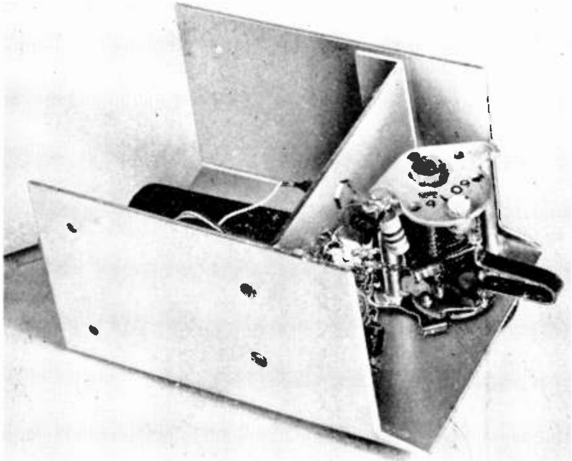


Fig. 19.30. Circuit diagram of the 450 Mc/s grid dip oscillator, with tuned inductance detail.



The 450 Mc/s grid dip oscillator with cover removed to show construction and layout.

to be applied to the circuit under test. Resonance is detected by a deflection on the g.d.o. meter due to rectified r.f. It should be noted that an absorption wavemeter will respond to harmonics if harmonic power is present and the wavemeter is tuned to the frequency of the harmonic.

Capacitance and Inductance. Obviously, if an instrument has the ability to measure the frequency of a tuned circuit, it can also be used for the determination of capacitance and inductance. To measure capacitance, a close tolerance capacitor (C_S) is first connected in parallel with a coil (any coil will do, providing it will resonate at a frequency within the range of the g.d.o.). This circuit is coupled to the oscillator and its frequency (F_1 Mc/s) noted. The unknown capacitor (C_X) is then connected in place of C_S and the resonant frequency again determined. If this is now F_2 Mc/s, the unknown capacitance is given by:

$$C_X = \frac{F_1^2 C_S}{F_2^2} \text{ pF}$$

Similarly, inductances may be measured by connecting the unknown coil in parallel with a known capacitance and applying the formula

$$L = \frac{25{,}300}{CF^2}$$

where L is in μH, C is in pF and F_2 is the resonant frequency expressed in Mc/s.

Signal Generator. For receiver testing the g.d.o. may be used to provide unmodulated c.w. signals by tuning the oscillator to the required frequency and placing it close to the aerial terminal of the receiver. The amplitude of the signal may be controlled by adjusting the distance between the g.d.o. coil and the aerial terminal.

A.f. signals can be injected at the jack socket on the g.d.o. to provide a modulated test signal, the modulation depth being dependent on the level of the modulation voltage, 5 volts being adequate.

C.W. and Phone Monitor. By connecting a pair of headphones to the instrument, it may be used as a c.w. monitor. It should be tuned to the transmitter frequency and the distance between the oscillator and transmitter adjusted for optimum signal strength. To monitor telephony transmissions the h.t. is switched off.

Checking Crystals. It is possible to check the activity and frequency of oscillation of a crystal by inserting it in the coil socket of the g.d.o. and setting the frequency dial to maximum (i.e., minimum tuning capacitance). The meter reading will vary in accordance with the activity of the crystal and the frequency may be checked by locating the oscillation on a calibrated receiver. Generally speaking, this procedure may only be applied to fundamental mode crystals.

TRANSMITTER POWER OUTPUT MEASUREMENT

From time to time most amateurs wish to know the actual power output from a transmitter to ensure that it is operating at its correct efficiency. Setting up a valve as an r.f. oscillator or power amplifier according to a published design or the information in the manufacturers' data and assuming that the stated power output will be automatically achieved is not sufficient.

Probably the most practical and inexpensive method of measuring r.f. power output is to use a two lamp comparative method, where one lamp is connected to the equipment and the other to a variable source of power. It is essential that both the lamps should be of the same type and power rating and it is desirable that before use the consumption data is recorded for both lamps so that allowances may be made for any differences. The actual difference between two lamps of the same type and rating is unlikely to be sufficient to make calibration essential for practical purposes but is necessary for those who wish to be as accurate as possible.

As may be seen from the typical curves shown in **Fig. 19.31** the light output from a lamp varies rapidly with changes of voltage or current and it is important therefore that direct comparison of the output is made. This is best arranged by fitting the two lamps side by side into a box having a ground glass screen over the lamps and with a dividing panel between the two lamps. The two lamps will then each illuminate the two separate halves of the ground glass screen.

Matching is easier using a ground glass screen instead of viewing the lamps directly and this is particularly true at the higher filament temperatures. When the lamps are enclosed in the box there is also much less interference by stray room lighting. A suitable arrangement is shown in **Fig. 19.32.** A variable power supply for the standard lamp is required and either a voltmeter and ammeter or a voltmeter and a calibration curve of the lamp is necessary for making measurements.

If a Variac transformer or other iron-cored device is used to control the applied a.c. voltage instead of a resistor it will be necessary to take account of waveform error by using a moving iron voltmeter and ammeter which read true r.m.s.

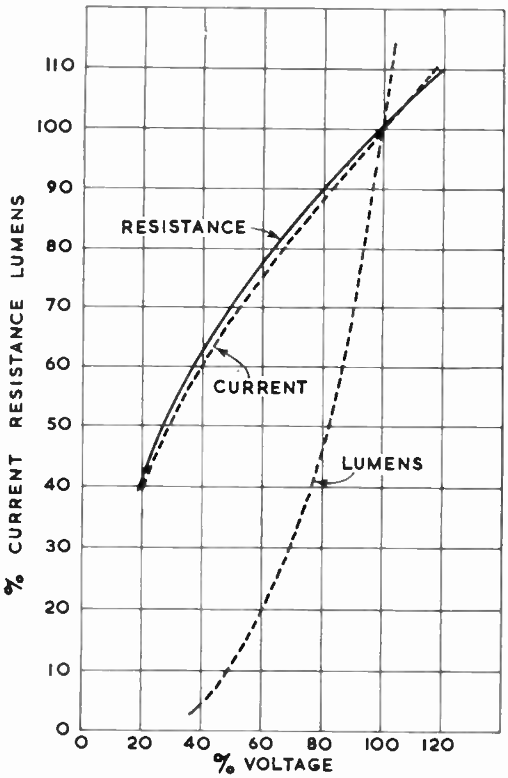


Fig. 19.31. Characteristics of gas filled lamps.

values. This point may be a trifle academic in view of the inherent inaccuracies of this form of power measurement, but it may be of interest to those seeking the highest possible accuracy.

Alternative Single Lamp Method

In this case the lamp is supplied with power either from the equipment or a measured source, the light output being observed by means of a selenium photo cell and micro-ammeter or a photographic exposure meter. The procedure is then as follows:

(i) First supply r.f. power to the lamp from the transmitter.

(ii) Take a reading of the light output using the indicator (selenium photo cell and microammeter or exposure meter) at a suitable and measured distance.

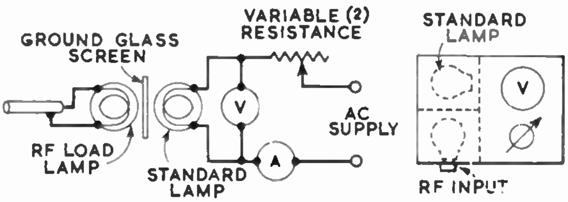


Fig. 19.32(a). Circuit arrangement of the two lamps. (b) Suggested layout. An effective " ground glass " screen may be made by using a sheet of Perspex, one side of which has been roughened with a household abrasive such as Vim. Suitable rotary variable resistors are available from P.X. Fox, Curtis and Berco.

(iii) Next, supply the lamp with a measured source of power, with the indicator at the same distance as before, increasing the light output until the indicator reads the same as in step (ii). The actual power output can then be computed from the calibration data of the lamp being used.

Details of some small low voltage lamps suitable for use with low power equipment and with operating resistances near to nominal output impedances of modern transmitters are given in Tables **19.3** and **19.4**. The lamps listed cover a power range of 2·2 to 25 watts and a resistance range of 43 to 100 ohms. For higher powers two or more may be used in series but this should not be done at frequencies above 30 Mc/s. The series inductance of the lamps can be tuned out at any one frequency by a suitable capacitor (for example, a Philips trimmer).

TABLE 19.3

Lamp Class	Volts	Watts	Amp	R (ohms)
Aircraft (General)	24	6	0·25	96
Aircraft (General)	28	18	0·645	43·5(*a*)
Aircraft (General)	24	10	0·415	58
Trolley Bus	35	15	0·43	81.5
Trolley Bus	35	36	1·0	36(*b*)
Car Side	24	12	0·5	48
Car Side	16	6	0·375	43
Small Projector	50	25	0·5	100
Indicator	16	3	0·188	85
Indicator	12	2·2	0·183	65·5

(*a*) Two in series for 36 watts at 87 ohms
(*b*) Two in series for 72 watts at 72 ohms

TABLE 19.4

Lamp, Small Projector Type 50V 25W

Voltage	Current	Watts	Resistance	Lumens
50	0·5 A	25	100 ohms	100%
45	0·47 A	21·5	95 ohms	70%
40	0·44 A	17·6	90 ohms	46%
35	0·425 A	14·9	84 ohms	31%
30	0·375 A	11·25	78 ohms	20%
25	0·34 A	8·5	71 ohms	10·5%

It is obviously desirable to select a lamp which will have a resistance similar to the feeder impedance. In this connection it should be noted that the characteristics of the lamps listed are very much nearer to those required than those of types normally used for domestic lighting. Bearing in mind that as the power is reduced the lamp resistance falls, a lamp should be chosen that has the appropriate resistance to match the feeder impedance at the expected power level. The figures for R quoted only apply at full brilliance, i.e. normal voltage and current.

Lamps cannot, of course, be regarded as non-inductive and some care is necessary in making connections to keep the inductance as low as possible particularly when using them at v.h.f. and u.h.f. It is worth noting that the screw and single contact type cap is more readily fitted to coaxial plugs or sockets than the double contact bayonet type.

AERIAL AND FEEDER MEASUREMENTS

Radio frequency ammeters and voltmeters do not reveal the true power unless the line in which they are connected is correctly matched, but they may always be used as a means of tuning up the transmitter and the aerial coupler. The *reflectometer* is, however, an instrument which can distinguish between the forward and reflected waves in a standing-wave system, and can therefore be used to indicate a true match of impedance; because only the forward wave is radiated, such an instrument can be calibrated to read true power flow into the aerial.

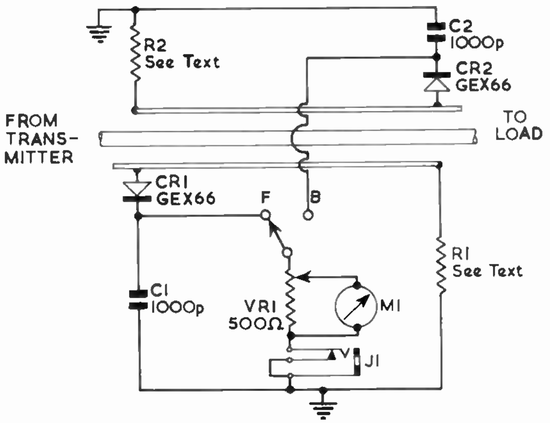


Fig. 19.33. Circuit arrangement of a reflectometer for the h.f. amateur bands.

Reflectometers are Wheatstone-bridge arrangements, balanced to indicate flow in one direction only along a line. Several varieties have been described in radio publications in which the bridge comprises physical components built into the concentric line, but these require careful construction for successful results. That described here uses the properties of coupled transmission lines, and two arms of the bridge are formed by the electric (capacitive) and magnetic (inductive) couplings. The coaxial line impedance and a balancing resistor in the reflectometer make up the remainder of the bridge.

H.F. REFLECTOMETER

The reflectometer is essentially simple in principle and in construction. **Fig. 19.33** shows the main conductor of a coaxial line and coupled to it two secondary lines, each fitted with a diode detector and a terminating resistance. The forward wave from transmitter to aerial can only induce a wave flowing in the opposite direction on the secondary line. This is a fundamental property of coupled lines. The lower line detector CR1 therefore receives a voltage proportional to the amplitude of the forward wave, the flow from transmitter to aerial. In the upper line the induced wave travels towards R2, and if this also matches its line, the induced wave is absorbed and nothing reaches detector CR2. On the other hand, if the aerial is not matched to its line, then it will reflect power back to the transmitter: the forward diode CR1 now shows total power going to the aerial and the backward diode CR2 the reflected portion.

Fig. 19.34 illustrates a construction suitable for insertion in a coaxial line using coaxial connectors. Only the essential dimensions are given. The concentric line is converted to a parallel plate line, comprising a $\frac{1}{4}$ in. conductor between two "earth" strips $\frac{3}{4}$ in. wide and $\frac{5}{8}$ in. apart which are bolted to the two end plates holding the connectors. The secondary lines, made of 16 s.w.g. wire, are supported in sawcuts in two polystyrene blocks, and the diodes and resistors are soldered to these lines. The complete instrument is boxed in by a metal wrapping screen over its end plates.

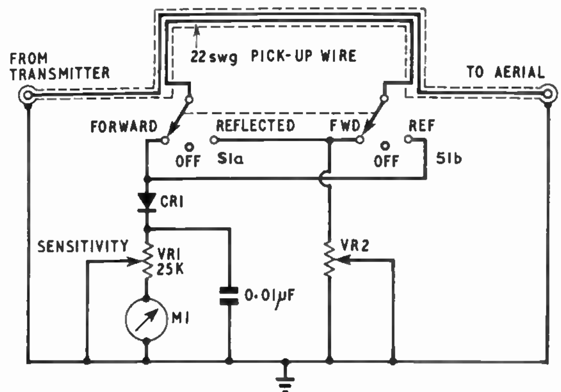


Fig. 19.35. Reflectometer for the h.f. bands.

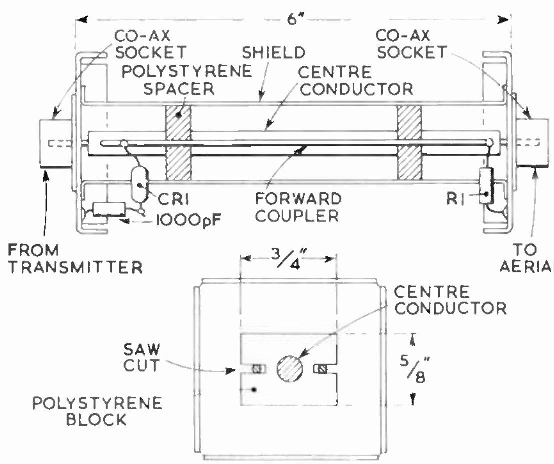


Fig. 19.34. Practical details of a reflectometer of Fig. 19.33. For 3·5–28 Mc/s the length should be 12 in.

The impedance of the strip line is 50 ohms, but it may also be used in 70 ohm circuits because R1 and R2 can be varied to adjust the balance of the equivalent bridge circuits. These resistors may be of $\frac{1}{2}$ watt rating but must be carbon and not wirewound. The value is 150 ohms for a 50 ohm reflectometer, and 100 ohms for 70 ohm use. The indicating meter is connected across a potentiometer VR1 which adjusts the sensitivity. This meter may be 0–1 mA f.s.d. for 50 watts on 14, 21 and 28 Mc/s, but for lower frequencies a more sensitive meter may be needed, since the output of the coupling lines is proportional to frequency; alternatively, the instrument may be made twice as long as suggested in Fig. 19.34 to double its sensitivity. Small wire-ended germanium diodes (e.g. GEX66, OA71 or 1N34) are suitable detectors.

It is necessary to balance the coupling lines, and this is done by moving them parallel to the main line to make a small adjustment to their characteristic impedance. For this purpose it is necessary to terminate the aerial socket in a pure resistance equal to the line characteristic impedance but unless high-power accurate r.f. loads are available this adjustment must be done at low power, in which case the meter is not sensitive enough. The procedure then is to replace the meter by a pair of headphones and use some audio modulation on the transmitter or its power driver, using very low output coupling and a 1 or 2 watt carbon resistor as a standard load of value equal to the coaxial line impedance. With this arrangement the backward-wave line is adjusted for zero output and then the instrument is reversed and the forward line set up. The adjustment should bring the secondary lines to about ¼ in. from the main conductor. It has been found that an improved balance is obtained if the diode is connected about ½ in. from the end of its line to form a small end-capacitance. The balancing adjustment should always be made at the highest frequency at which the reflectometer is to be used, because there the errors are most important, and it is also most sensitive there. When the adjustments are finished the wires can be cemented in position.

In use, the potentiometer VR1 is set for full-scale meter reading when switched to FORWARD; the meter will indicate reflection coefficient when changed to the backward position. If there is a reasonable resistance, say at least 1000 ohms, in series with the meter and the diode, the readings will be reasonably linear, proportional to reflection amplitude. Without damping the readings are nearer to square law and are proportional to power.

The reflection coefficient K is the best matching parameter to use, since it is zero when the line is matched, but if it is preferred to work in terms of v.s.w.r. the meter can be calibrated by means of the formula:

$$V.S.W.R. = \frac{1+K}{1-K}$$

The reflectometer can be used "in the field" as a very helpful aid to aerial matching operations, but its normal use is in the station, either in the aerial feeder or in the link line between transmitter and aerial coupler. Where two or more aerials are used on one band it is a big advantage to have each aerial matched into the line (either directly or via an aerial coupler) because then the transmitter can be adjusted for a matched line load, and change of aerial can be made instantly as required. A good match is indicated by zero or very low reflection when the meter is switched to the backward diode (CR2). When the reflection is zero, the forward detector indicates true power flow to the aerial and can obviously be used for transmitter adjustment and as a permanent monitor. For a.m. phone work side tone may be obtained by plugging headphones into J1.

A rough power calibration of the forward detector can be made by using a 25 or 100 watt lamp as a load on the aerial coupler, and matching its brilliance with a similar lamp supplied with mains power. The r.f. lamp is tapped into the aerial coupler as if it were the aerial and adjustments made until it matches the link line. This procedure needs patience because the lamp resistance changes rapidly with power input. The reference lamp is connected with meters to read the current through it and the voltage across it, and is supplied from a suitable source of power with a variable resistance in series. The two lamps should, of course, be of the same type.

A simplified form of reflectometer for the h.f. bands using modified coaxial cable is shown in Fig. 19.35.

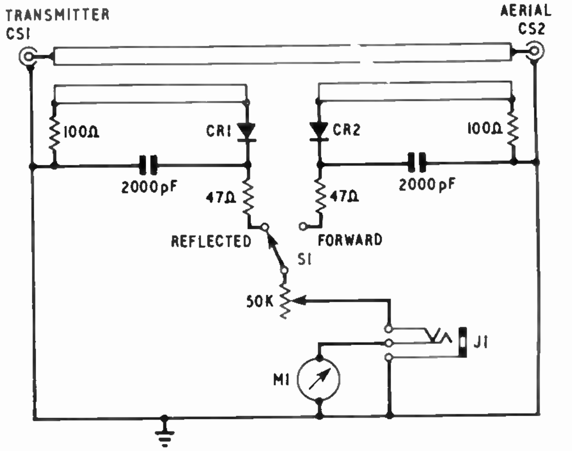


Fig. 19.36. Circuit diagram of the v.h.f. reflectometer.

It should be remembered that when the diodes are passing rectified current they are also generating harmonics and may therefore cause some degree of TVI. For this reason the complete instrument should be properly screened in a closed metal box and if the meter is detached, its leads should be screened and decoupled. A choke or single section TVI filter between the switch and potentiometer will help prevent harmonic generation and radiation.

The instruments shown in Figs. 19.34 and 19.35 will operate quite well at 145 Mc/s but a reflectometer specially designed for v.h.f. use is shown in Figs. 19.36 and 19.37.

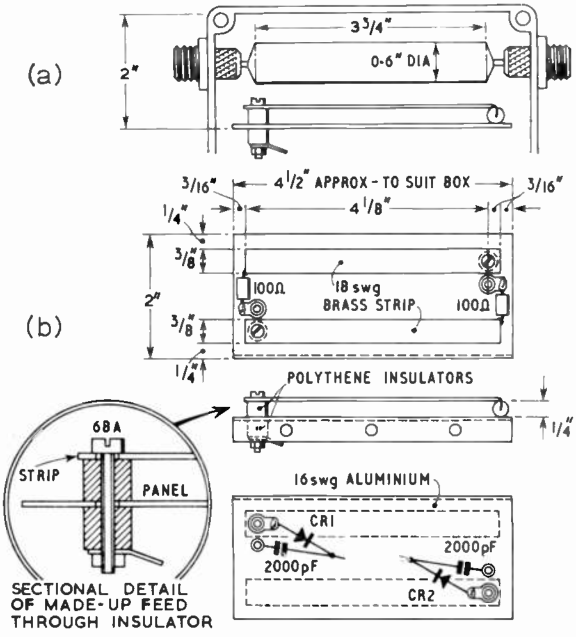


Fig. 19.37. Constructional details of the circuit of Fig. 19.36.

FIELD STRENGTH METER

A field strength meter is a useful tool for use in adjusting aerial systems and as a visual indication of a signal being radiated. As may be seen from **Fig. 19.38**, it comprises a tuned circuit, switched for the required bands, and a crystal rectifier feeding a single stage transistor d.c. amplifier. The signal level is indicated on a 0–1 mA meter. The instrument should be constructed in a metal box, its pick-up aerial being connected via a suitable insulated terminal.

To avoid obtaining confusing results, a field strength meter should be situated several wavelengths away from an aerial under adjustment.

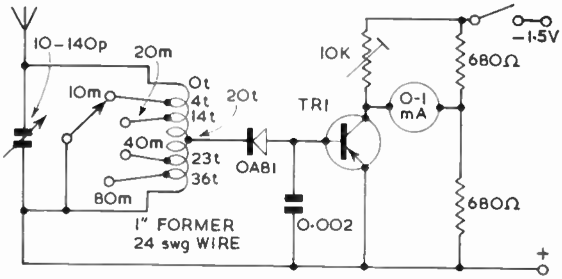


Fig. 19.38. Simple field strength meter for 10-80 metres. TR1 may be any general purpose transistor. A short rod aerial should be used (3-6 ft.).

SIGNAL GENERATORS

In setting up and testing radio and electronic equipment it is frequently necessary to have an a.f. or r.f. signal of suitable frequency. A.f. oscillators are used in the testing of speech amplifiers and modulators and in setting up single sideband transmitters while r.f. signal generators are of great value in the alignment of communication receivers and converters.

SIMPLE A.F. OSCILLATOR

An audio oscillator for amateur use need not be very complicated and in fact the essential part of the circuit to be described consists simply of two equal *RC* time constants, one in parallel and one in series, which are connected to a regenerative system in which the gain and phase relationship will sustain oscillation at a frequency which depends solely on the *RC* values.

In the circuit shown in **Fig. 19.40** the twin triode oscillator (V1) feeds a cathode follower (V2) so that it remains unaffected by the load into which it works. The switch S1ab provides six ranges covering 32 c/s to 33 kc/s.

The frequency can be readily calculated from $f = \dfrac{1}{2\pi RC}$

where R is the value to which one section of the twin potentiometer VR1, VR2, has been adjusted and C is the capacity of the respective range capacitor. For this relationship to hold, the R and C values in each section should be within ± 1 per cent of each other.

Each section of the potentiometer VR1, VR2, is 100,000 ohms. Other values can be used, e.g., 50,000 ohms, in which case all the associated capacity values should be doubled, or 200,000 ohms with all the capacity values halved. Beyond these extremes, the R/C ratios become rather big in one case and small in the other, and oscillation is erratic at the ends of each range.

The lamp in the cathode circuit of V1 is a 230 volt type of 5 to 15 watts consumption. The resistance of the lamp varies with the thermal effect of the current flowing through it and keeps the amplitude of oscillation reasonably constant and independent of the frequency. The filament does not glow but its operation can be verified by watching the a.f. voltage across it on an oscilloscope. The thermal lag of a few seconds can be observed and the lamp therefore requires to accommodate itself after a frequency change has been made.

The variable resistor VR3 is a regeneration control and has a marked effect on the distortion present. If VR3 is at its maximum value the output is more or less distorted at all frequencies. Its resistance should be reduced until the circuit just oscillates satisfactorily on all ranges. Oscilloscope traces of the output look clean with the exception of those on the lowest ranges which may still have a flattened crest. This can be remedied by shunting C7 to earth with a 500,000 ohm resistor R1.

The maximum audio voltage available is about 7 volts R.M.S. For many purposes, lower voltages are required and VR5 in conjunction with S2 permits the output to be adjusted from 0 to 150 mV.

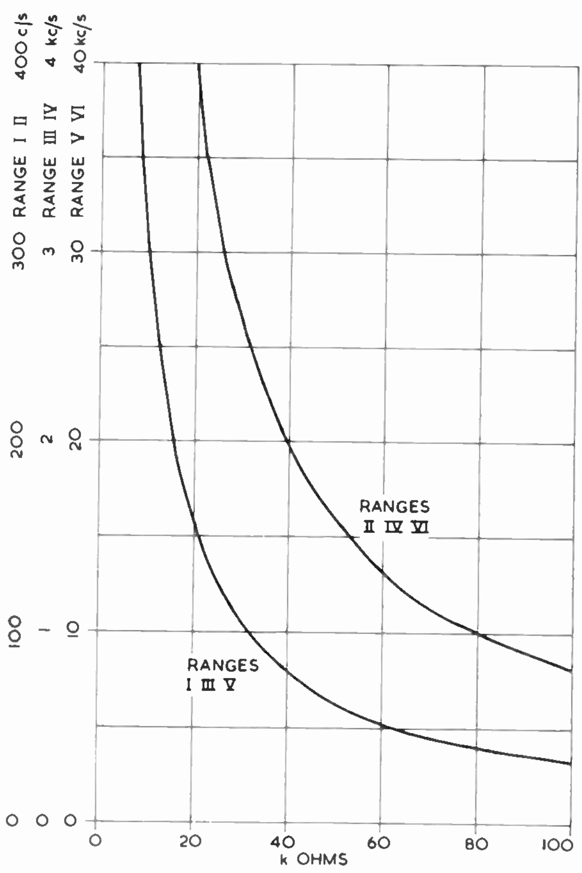


Fig. 19.39. Frequency v. resistance curve for the circuit and values of Fig. 19.40.

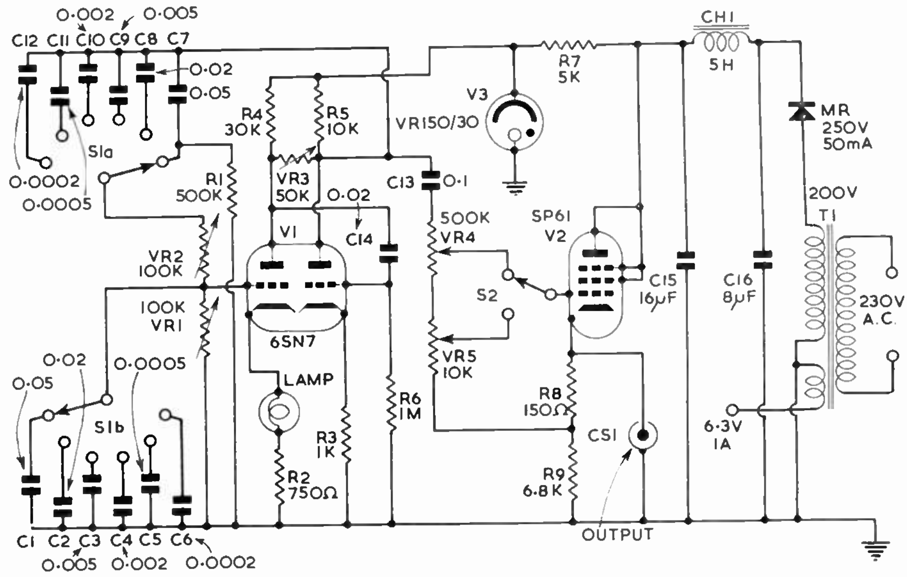


Fig. 19.40. Circuit of the audio oscillator. VR1 and VR2 are a twin potentiometer, 100 K ohms each section. To provide d.c. isolation, CS1 should be fed from V2 cathode via a 0·1 μF capacitor.

Fig. 19.39 shows the frequency versus resistance curve of the circuit described. It is clear that 10 octaves can be covered without the cramped part of each range being used by working from 100,000 ohms to 20,000 ohms.

Lissajous' figures are often suggested as the best means of calibrating an instrument of the type described, and indeed they are excellent for the lower frequencies. The method of using them is as follows. With 50 c/s (mains frequency) a.c. voltage fed into the horizontal deflecting plates of an oscilloscope, the output of the audio oscillator should be applied to the vertical plates and the tuning control adjusted until the pattern of **Fig. 19.41(a)** appears on the screen. The oscillator frequency is then 25 c/s and the scale may be appropriately marked. The tuning control may then be rotated until the pattern of **Fig. 19.41(b)** appears, which indicates a frequency of 50 c/s. Similarly, the calibration points relating to the range 100–250 c/s may be fixed by reference to the patterns shown; finally the whole scale may be calibrated accurately by interpolation. There are of course many intermediate Lissajous' figures which may be obtained in addition to those shown. Readers who wish to work out the frequencies for which these apply may do so from the method given in **Fig. 19.42.**

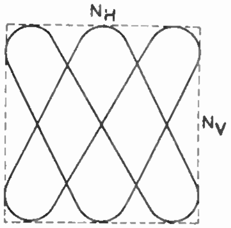


Fig. 19.42. If a sine-wave of unknown frequency is fed to the horizontal deflecting plates of an oscilloscope and a signal from a calibrated source is fed to the vertical plates, a stationary pattern such as that shown above may be obtained by adjusting the frequency of the calibrated source. The number of loops in both horizontal and vertical planes should be counted, and the unknown frequency may then be calculated from the following equation: Unknown frequency = NV/NH × frequency of calibrated source. In the diagram, the unknown frequency is ⅔ of the calibrating frequency.

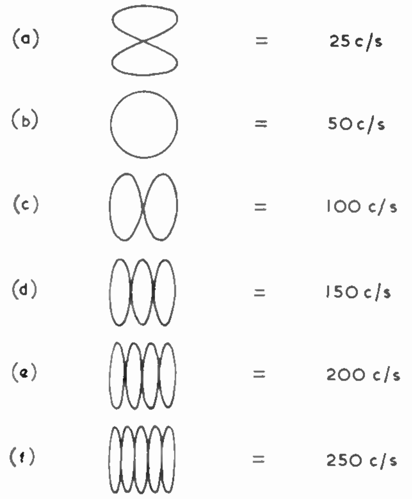


Fig. 19.41. Calibration of an audio oscillator by means of Lissajous' figures using 50 c/s a.c. mains.

Above 400–600 c/s calibration against the mains ceases to be practicable. One way to go on from there is to use a second oscillator which may be set to an intermediate point (say 200 c/s) and used to calibrate the first oscillator by means of Lissajous' figures up to approximately 1000 c/s. The auxiliary oscillator may thereupon be set to 1000 c/s, and the process repeated in sequence until the full range has been covered. The principal requirement of the auxiliary oscillator is frequency stability; purity of waveform is unimportant.

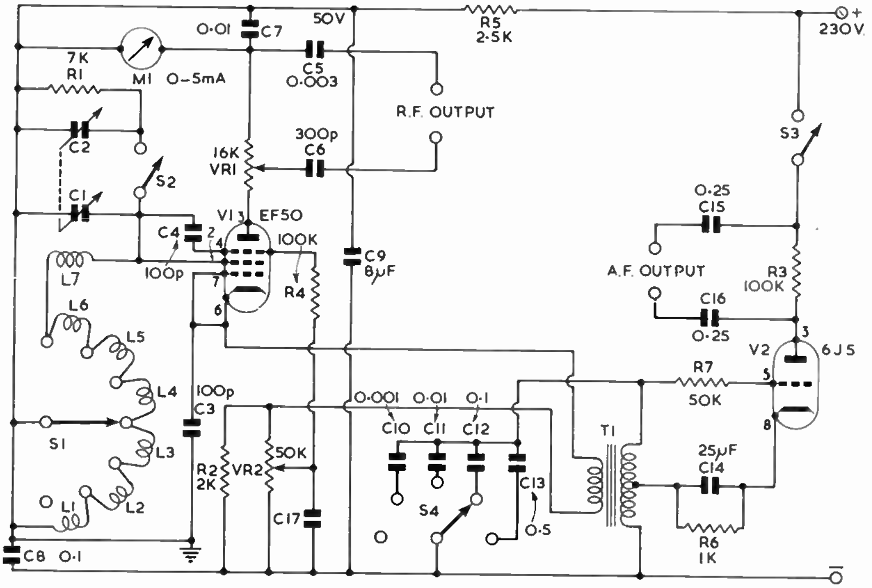


Fig. 19.43. Circuit of a transitron signal generator covering 80 kc/s–56 Mc/s. L1, two long wave b.c. coils in series; L2, medium wave b.c. coil with one quarter of turns removed; L3, 23 turns 40 s.w.g. d.c.c. close wound on 1¼ in. dia. former; L4, 17 turns 24 s.w.g. enam. close wound on 1¼ in. dia. former; L5, 8 turns 20 s.w.g. enamel close wound on 1¼ in. dia. former; L6, 7 turns 20 s.w.g. enam. close wound ⅜ in. inside dia. (no former) L7, 7 turns 20 s.w.g. close wound 5/16 in. inside dia. (no former). A 6·8 K ohms resistor may be used for R1 and a 10 K ohms potentiometer for VR1 if desired. T1, midget speaker transformer. C17 should be 0·001μF.

R.F. SIGNAL GENERATOR

The r.f. signal generator shown in Fig. 19.43 covers 80 kc/s to 56 Mc/s and employs a transitron oscillator (V1) modulated by a conventional a.f. oscillator (V2).

Approximately 3–4 mA anode current and 6–8 mA screen current is drawn by V1, the sum of the two being practically constant for any setting of VR2. The value of R5 depends on the h.t. supply and should be arranged to produce 50 volts at the point indicated. Construction is simplified if this 50 volt rail is earthed instead of the negative as is usual practice.

The seven range coil assembly consists of inductances in series which are tapped by S1. One end of the highest frequency coil L7, should be mounted directly on the screen grid pin of the valveholder, the switch being arranged close to it. Winding of the coils should commence from the L7 end to make sure that there are no gaps. Some experiment may be necessary with the number of turns.

A 500pF twin gang capacitor with ceramic insulation is used for C1, C2. The C1 section is reduced to approximately 100pF by removing vanes; it is used on Ranges 4 to 7. For Ranges 1 to 3, C2 is switched in parallel with C1 by means of S2.

The frequency ranges covered by the capacitors suggested and the coils specified, are as follows: *Range 1*, 80 to 220 kc/s; *Range 2*, 210 to 760 kc/s; *Range 3*, 660 to 2700 kc/s; *Range 4*, 2·6 to 6·5 Mc/s; *Range 5*, 5·6 to 15·8 Mc/s; *Range 6*, 14 to 29 Mc/s; *Range 7*, 24 to 56 Mc/s.

Attenuation of the r.f. output is provided by VR1, a non-inductive carbon type potentiometer. The arrangement is unorthodox but it does provide an easy means of attenuating the output. Modulation is by cathode injection, the audio oscillator V2 providing four fixed tones selected by S4.

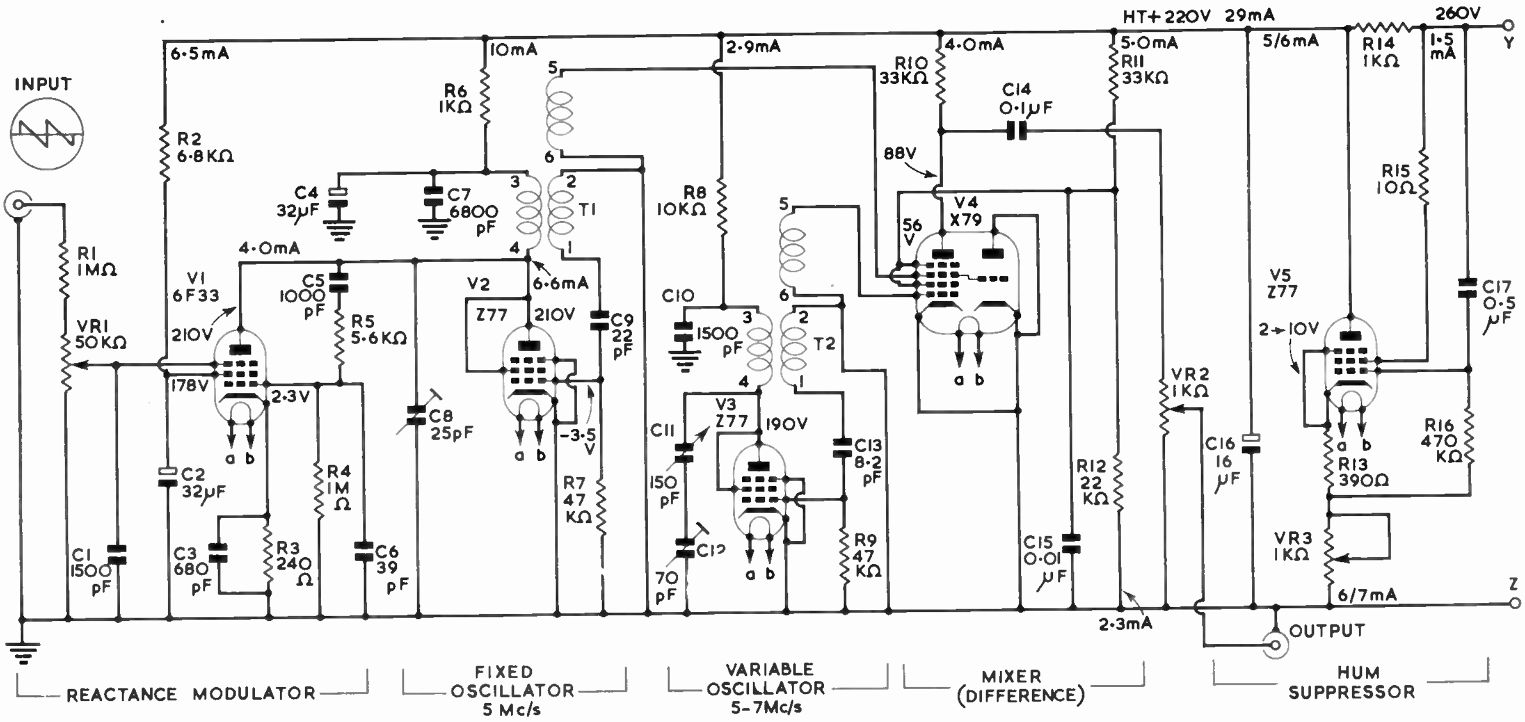
Screening of the entire unit, including the power pack, is desirable in order to avoid signal leakage. The tuning dial should be fitted with a vernier so that it may be read to one-tenth of a degree.

Calibration can be carried out with the aid of a frequency meter or by listening on a receiver for beats with known broadcast stations and short-wave stations.

WOBBULATOR FOR RECEIVER ALIGNMENT

The main function of a wobbulator is to simplify the alignment of i.f. strips and bandpass filters to produce a clean, flat-topped, steep skirted response with minimum side lobes. Using a signal generator and v.t.v.m., the time consumed can be considerable, for a new graph has to be constructed following every major adjustment; with a wobbulator and oscilloscope, the pass-band can be observed at a glance, and even the effect of an adjustment actually being made can be continuously monitored.

As there are a number of different intermediate frequencies in general use, extending from 50 kc/s to 2 Mc/s and often beyond, a simple oscillator to cover this range would require several stages of coil switching. An alternative arrangement, however, was chosen which produces an output continuously variable over the desired range. This is accomplished by mixing the output of a 5 Mc/s fixed frequency oscillator with that of a 5 to 7 Mc/s variable oscillator. The process of frequency modulating the output occurs within part of the tuned circuit of the fixed oscillator, which ensures that the frequency deviation remains constant regardless of the output frequency, this being another reason for the choice of the particular method of generation. The maximum permissible sawtooth wave injection to the 6F33 reactance modulator (V1), Fig. 19.44, to produce a linear frequency deviation is able to shift the nominal 5 Mc/s by approximately 60 kc/s, which should be adequate for all normal purposes. In order to present a single display on the oscilloscope, the signal injected into V1 should be of the same repetition frequency as the X-scan on the oscilloscope, and also a clean sawtooth. The wobbulator has therefore been designed without the inclusion of a sweep-frequency oscillator, for the output



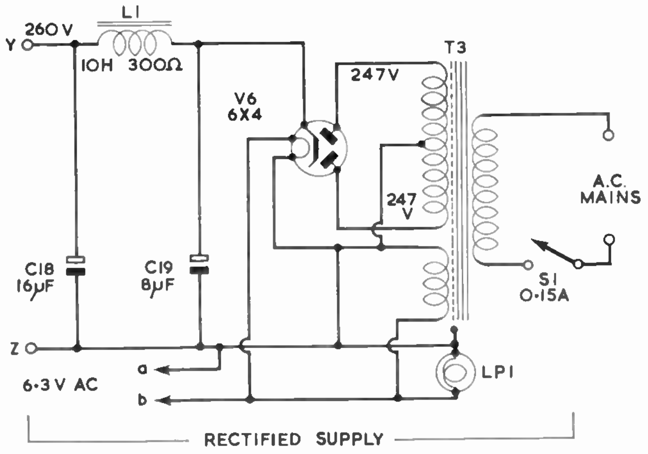


Fig. 19.44. Circuit of the wobbulator and power supply.

Details of Oscillator Coils

Fixed Frequency Oscillator

T1—7/16 in. dia. Aladdin F804/PP5892 former with dust core.

Main tank winding	34 turns of no. 32 s.w.g.
Feedback winding	16 turns of no. 32 s.w.g.
Coupling winding	4½ turns of no. 26 s.w.g.

All windings single layer, close-wound, enamel. Feedback winding on top of main winding near earthy end.

Coupling winding spaced ⅛ in. from main tank winding. Main tank winding inductance approximately 17 μH, Q approximately 70.

Variable Frequency Oscillator

T2—7/16 in. dia. Aladdin former with dust core.

Main tank winding	30 turns of no. 32 s.w.g.
Feedback winding	12 turns of no. 32 s.w.g.
Coupling winding	4¾ turns of no. 26 s.w.g.

All windings single layer, close-wound, enamel. Feedback winding on top of main winding near earthy end.

Coupling winding spaced ⅛ in. from main tank winding. Main tank winding inductance approximately 15·6 μH, Q approximately 70.

of the timebase of the oscilloscope in use should be perfectly satisfactory. Indeed, any irregularities in the wave-shape become unimportant when the same oscillator is used in both cases. This method also eliminates the necessity for synchronization of oscillators.

The 6F33 (V1) was chosen as the reactance modulator because of its superior linear mutual conductance/suppressor grid voltage characteristic. It is important to see that both the anode and screen are adequately decoupled with respect to the sweep frequency.

The oscillator (V2, V3) frequencies were originally chosen as a compromise between stability of output centre frequency and ease of obtaining the required tuning range. As the output frequency is dependent on the difference frequency between the two oscillators, 5 Mc/s was not considered unduly high for stability. The oscillators being electrically similar, conditions causing a shift in one oscillator will similarly affect the other. The difference frequency should thus remain constant.

Should the necessity arise, the frequency range can be extended by the modification of one or both oscillators, with due consideration for unwanted mixing products.

The oscillator outputs are fed into the mixer (V4) and as more than sufficient output was available, the anode load of V4 was deliberately made low.

The instrument will work satisfactorily from an h.t. supply of 220 to 250 volts, and to avoid low frequency cyclic distortion of the display, the smoothing must be of a high order. A conventional capacitance input filter followed by a hum suppressor was found to give better results than two identical filter sections. The a.c. ripple is minimized by adjustment of the potentiometer in the cathode circuit of the hum suppressor, and in the case of the prototype was reduced to less than 2 mV R.M.S. on full load (30 mA) with no sweep voltage applied.

If the power supply is to be contained within the same case as the instrument, adequate ventilation must be provided.

The layout of the unit is flexible within reason, although a logical design ought to be adhered to as far as possible. Satisfactory operation was achieved without enclosing the oscillators in screening boxes, but they could be added as an additional precaution if desired. All the valves should be fitted with screening cans, and the entire instrument enclosed in a metal case.

Following construction, the wiring must be checked thoroughly. After switching on, the panel light and valve heaters should glow, and then, provided no components become unduly hot, the unit should be allowed to run for at least 20 minutes before beginning alignment. The oscillator frequencies must subsequently be corrected, and the hum suppressor adjusted with the aid of a valve voltmeter or sensitive oscilloscope.

Establishing the correct nominal frequencies may be accomplished in one of three ways. After reducing the variable oscillator tuning capacitor to minimum, the fixed oscillator may be aligned by alteration of the slug of T1, and C8, together with any one of the following three methods of frequency determination:

(i) Absorption wavemeter.

(ii) Zero beat with a receiver tuned to 5 Mc/s having b.f.o. on.

(iii) Zero beat with a signal generator tuned to 5 Mc/s, the outputs being mixed and displayed on an oscilloscope. The circuit of a suitable mixer is shown in **Fig. 19.45.**

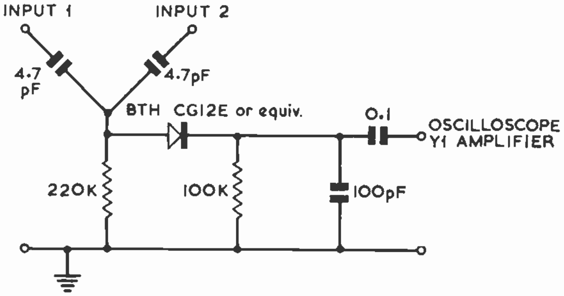


Fig. 19.45. A simple mixer for use when aligning the fixed and variable frequency oscillators of the wobbulator.

The variable oscillator must be set to cover the range 5 Mc/s to 7 Mc/s. For 5 Mc/s, C11 should be positioned at maximum capacitance, and the slug of T2 adjusted for zero beat with the fixed oscillator. If, with this setting, C11 is not capable of tuning the full 2 Mc/s, alteration of the series padder C12 will improve the coverage, but the initial 5 Mc/s adjustment will have to be repeated.

Calibration

The variable oscillator capacitor is calibrated directly in terms of difference frequency, and hence " centre frequency " when a sweep waveform is applied. It is recommended that the process of calibration be attempted when the outputs of the oscillators are not being mixed, as direct 5-7 Mc/s frequency measurements leave less likelihood of unintentional measurement of spurious mixer products.

Calibration of the deviation control is the next step. With VR1 at minimum (zero sweep) the output can be mixed with a reference frequency from a signal generator, and tuned to zero beat. The combined output should, for convenience, be displayed on an oscilloscope. On variation of the generator frequency, by say, 10 kc/s (**Table 19.5**), the zero beat pattern will immediately vanish, until VR1 is altered to

TABLE 19.5

Trace Deviation Linearity Measurements

Frequency (kc/s)	Trace Deviation (cm)	Frequency (kc/s)	Trace Deviation (cm)
270	0·0	1850	0·0
280	0·7	1860	0·5
290	1·1	1870	0·9
300	1·4	1880	1·2
310	1·7	1890	1·5
320	2·0	1900	1·8
330	2·3	1910	2·1
340	2·6	1920	2·4
350	2·8	1930	2·7
360	3·1	1940	3·0
370	3·4	1950	3·3
380	3·7	1960	3·7
390	4·0	1970	4·0
400	4·3	1980	4·4
410	4·6	1990	4·8
420	5·0	2000	5·2
430	5·3	2010	5·6
440	5·7	2020	6·0
450	6·1	2030	6·4
460	6·5	2040	7·1
470	7·0	2050	8·0
480	7·8		

resume the condition. The process should be repeated until the deviation control is at maximum, calibration marks being applied to the dial of VR1 at each step. The deviation will always remain constant, regardless of the centre frequency, since the maximum output frequency of the frequency modulated oscillator cannot vary.

Amplitude calibration of the output is unfortunately impracticable. The reason is that with variation of the centre frequency, the output voltage varies in sympathy. Reference divisions of from, say, 1 to 10, however, are very useful.

A TWO TONE TEST OSCILLATOR

A simple transistorized two tone test oscillator for the alignment of s.s.b. transmitters is shown in **Fig. 19.46.**

The oscillators provide frequencies of 1 kc/s and 2 kc/s the output from each being independently adjustable giving a maximum of 1 mW into 600 ohms combined. The power supply required is 12-15 volts at 10 mA.

The two test oscillators TR1 and TR4 are both single stage phase shift oscillators using three section ladder RC networks. This type of oscillator has been chosen for its simplicity and economy of components, though it does suffer from considerable dependence on transistor parameters. In many circuits embodying this type of oscillator, potentiometers are used to vary the conditions on the base or emitter in order to establish satisfactory oscillation. This should be unnecessary in this case but, should transistors other than those specified be used, it is useful to remember that small changes to R3 or R15 will affect the waveform and amplitude of the oscillator outputs.

The signals at the collectors of the oscillators are coupled by the isolating resistors R6 and R13 to the gain controls VR1 and VR2. This d.c. coupling ensures that the controls vary both the signal amplitude and d.c. bias on the bases of the output amplifiers. Thus, when only a low output amplitude is required, the standing current of the class A amplifiers is also low, ensuring maximum efficiency of these stages. The cold ends of the amplitude controls are taken to a fixed d.c. potential determined by R10 and C6. This prevents TR2 and TR3 from being cut off at very low amplitude settings.

The switches S1 and S2 each apply power to both oscillators and one of the output amplifiers. The output of each oscillator may therefore be obtained independently for setting up levels or response checks.

The common collector load for TR2 and TR3 ensures linear mixing of the two test tones and provides the required output impedance. Coupling to the load is through the 100 μF capacitor C5. This is suitable for driving into the low impedances encountered in transistorized modulators, but, if the application is restricted to valve amplifiers, the use of a 0·01 μF capacitor here may be necessary to reduce the effects due to the changing d.c. conditions in the output amplifier.

The complete oscillator with batteries may be accommodated in a die-cast aluminium box 4¾ in. × 3¾ in. × 2¼ in. All the wire ended components can be mounted on a printed circuit board or pin board 4¼ in. × 2½ in. which will fit, components downward, within the box leaving room for the controls and switches along one side and the batteries (eight U7 or U12) on top of the board. Distribution of the components within the box and a specimen printed board layout are shown in **Fig. 19.47.**

A strip of synthetic sponge can be used to hold the batteries in place and has been found simpler and more effective than a metal strap.

MEASUREMENT OF NOISE FACTOR AND EQUIVALENT NOISE TEMPERATURE

The importance of evaluating the noise performance of a receiver in relation to the noise produced in the other parts of a communication system has already been discussed in Chapter 15 (*Noise*). The purpose of this section is to describe the principles and practice involved in the construction of noise generators, and the method of using a thermionic diode noise generator to determine the noise factor of a receiver. In addition, details are given of a germanium diode noise source which, while not suitable for absolute measurement of noise factor, is convenient to use when making adjustments to the input circuits of a receiver to determine whether any improvement in noise factor has been achieved.

The anode current of a diode contains noise components due to shot effect, and the frequency spectrum covered by the noise output is exceedingly wide, from zero up to the kilomegacycle region and so may be used as a source of

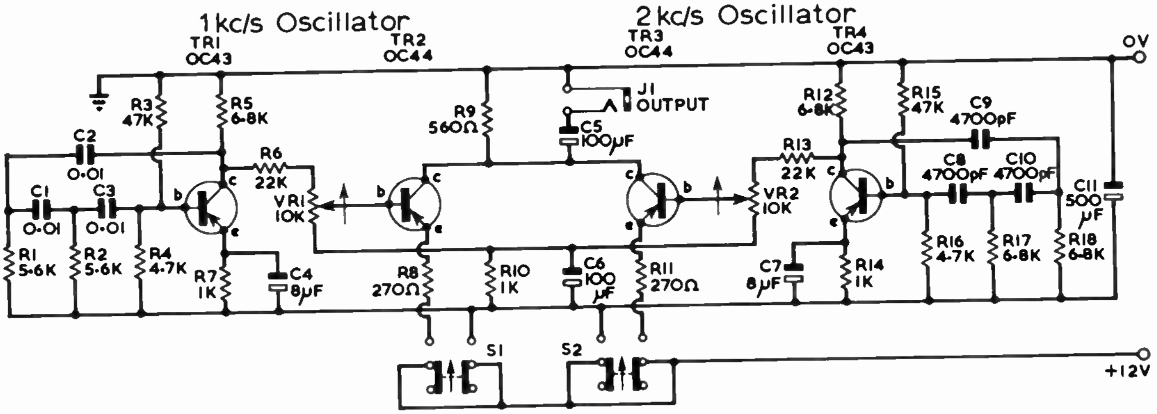


Fig. 19.46. The circuit of the two tone test oscillator.

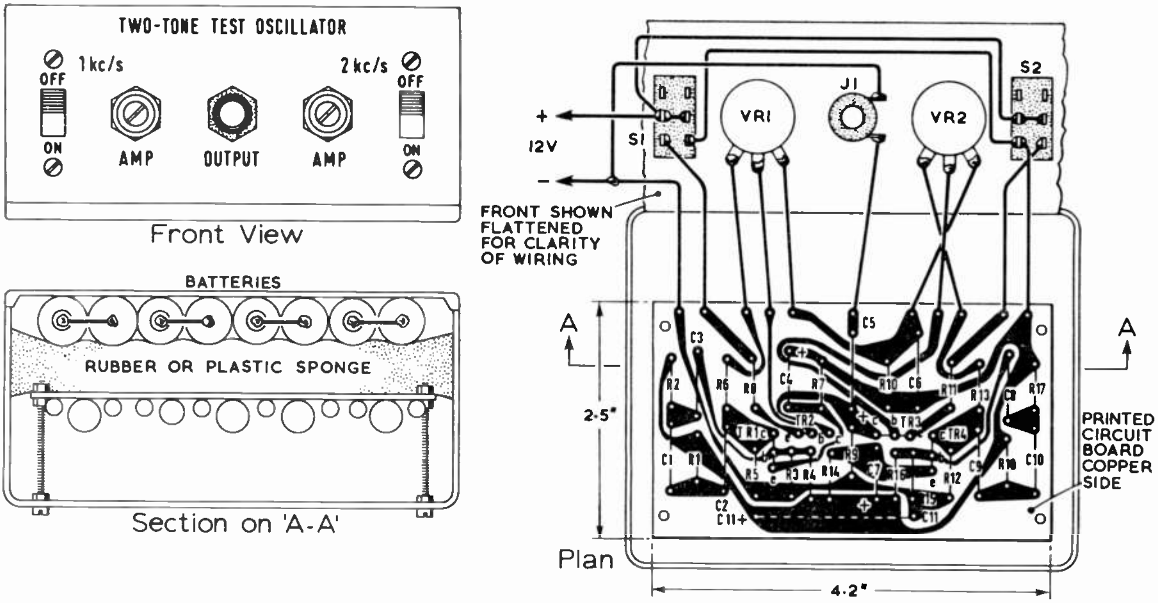


Fig. 19.47. A printed curcuit board layout suitable for the two tone oscillator.

noise signal at any frequency within this range. When the diode is operated under temperature limited conditions, i.e. the emission is controlled only by cathode temperature and not by anode to cathode potential, the noise power output may be shown to be directly proportional to anode current. Thus by varying the filament current, and hence its temperature, a close control of noise power output may be achieved.

The germanium diode generates considerable noise power when a current of a few milliamperes is passed through it in the reverse conduction (high resistance) direction. It has been found that low turnover voltage diodes produce the greatest noise output. Unfortunately the law relating diode current to noise output varies for each diode so that a germanium diode source requires calibration by a thermionic generator if it is to be used for absolute measurements.

Design Considerations

The noise output of the generator is developed across a load resistor equal in value to the source resistance from which the receiver under test is designed to operate. The anode resistance and output capacitance of the noise diode appear in shunt with the load resistor and care must be taken in the practical layout to keep the r.f. connections as short as possible to reduce the stray capacitance to a minimum. At the higher frequencies it is desirable to tune out the stray capacitance by means of shunt inductance if the greatest accuracy is required. This problem arises because the value of the load resistor appears in the formula used to calculate the noise factor and any shunt capacitance will lower the effective value of the load to an extent directly proportional to the frequency of measurement. The anode resistance of the diode will be much higher than the load resistor as the diode is operated with sufficiently high anode voltage to give anode current saturation and the resistive component of the shunt impedance may be neglected for all practical purposes. The d.c. anode supply to the diode must be fed in via r.f. chokes to ensure that all the noise generated is passed to the load resistor. It is not necessary for very elaborate smoothing to be used as the anode current will be independent of anode voltage variations under saturation conditions. A suitable noise diode is the A.2087(CV2171) made by M-O Valve Co. Ltd. This is a B7G based valve with a short mount, which minimizes the lead inductance associated with the valve itself. An h.t. supply of 100–150 volts at 20 mA is required together with a filament supply capable of varying the applied voltage up to a maximum of 4·4 volts. The tungsten filament takes 0·64 amp at the maximum filament voltage. The anode current of the CV2171 should not be allowed to exceed 20 mA.

For use at 450 Mc/s, the M-O.V. CV2398 is preferable, although the CV2171 is satisfactory at this frequency. The CV2398 is a noval based flying lead valve, requiring an unscreened p.t.f.e. lead mounting. A suitable component is made by McMurdo Instruments Ltd., under the code FB9A/F. The CV2398 has a tungsten filament rated at 6·0 volts, 1·15 amps and requires an h.t. supply of 100–150 volts at 45 mA maximum.

VALVE-TYPE NOISE GENERATORS

For ease of operation it is desirable to use coarse and fine variable resistors for the filament supply so that the noise diode anode current may be set accurately to any required value. For complete versatility, the noise generator should be capable of operating in an unbalanced or balanced condition in order to cater for both types of receiver input circuit. The simplest method of doing this is to provide two noise diode circuits fed from a common power unit, the noise diode being plugged into the appropriate valve holder as required. Suitable circuit arrangements are shown in **Figs. 19.48 and 19.49.** The meter M1 has three ranges, 1, 5 and 25 mA f.s.d., obtained by switching in appropriate shunts across the basic 0–1 mA meter. The maximum values of noise factor which may be measured on each range are shown in **Table 19.6.** When measurements at 75 ohms only are required then a 0–5 mA meter shunted to read 25 mA will be adequate.

The figures given in Table 19.6 are calculated by substitution in the formula $F(db) = 10\ log_{10}\ 20\ IR$, where R is the source resistance in ohms, and I is in amperes.

At frequencies up to 150 Mc/s it is permissible to connect the noise generator to the receiver by a short coaxial or balanced line but for the highest possible accuracy of measurement at 450 Mc/s the noise diode and its associated r.f. circuit should be constructed in the form of a probe, which can be connected directly to the input of the receiver under test.

The load resistor used should be of a type having minimum associated series inductance and parallel capacitance, and a coaxial resistor is recommended, such as the type DSR/4/1 made by the Welwyn Resistor Company in various values and tolerances. For this application a tolerance of ± 5 per cent will be adequate. The coaxial resistor is formed in the shape of a thin disc of ceramic coated with resistive material. Electrical connections are made to the centre and circumference of the disc by silver plated metal contacts. The method of mounting the resistor is shown in the illustration of the germanium diode noise generator

When shunt inductance compensation is used, the inductance is adjusted to resonance with the stray capacitance at the frequency of test as indicated by a rise in the noise output of the generator. Strictly speaking inductance compensation is accurate at one frequency only, but the resulting error in calculated noise factor is negligible over the frequency bands of interest to the amateur. A practical point to observe when using a noise diode is to permit anode current to flow for the minimum time possible when making measurements as the expected life of a valve of this type does not exceed a few hundred hours and is much less if the filament voltage is maintained at the maximum value for long periods.

TABLE 19.6

Meter F.s.d.	Noise Factor	
	75 ohms	300 ohms
1 mA	1·76 db	7·78 db
5 mA	8·75 db	14·77 db
25 mA •	15·74 db	21·76 db

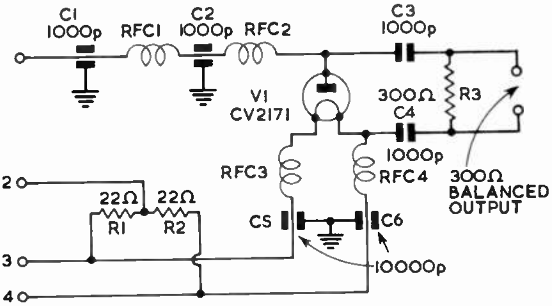


Fig. 19.49. Noise diode circuit suitable for balanced input. C1, C2, 1000 pF ceramic lead-through capacitors; C3, C4, 1000 pF tubular ceramic; C5, C6, 10,000 pF ceramic lead-through; R1, R2, 22 ohm ½ watt; R3, 300 ohm 5 per cent (Erie type 5B); RFC1, RFC2, 50 turns 30 s.w.g. close wound on 3/8 in. dia. former; RFC3, RFC4, 36 turns 18 s.w.g. self-supporting ½ in. internal diameter.

GERMANIUM DIODE NOISE GENERATOR

The circuit of a germanium diode noise generator is given in **Fig. 19.50.** The construction, illustrated overleaf, is quite straightforward and follows the principles outlined for valve diode generators. The generator requires 100 volts at 5 mA which may be obtained from any convenient supply. A 0–5 mA meter inserted at J1 measures the crystal current. With some germanium diodes it may be found that the noise output first increases and then falls off before rising again with increase of crystal current. This departure from linearity will not be a disadvantage as the germanium diode noise generator is used principally for qualitative assessment of receiver performance.

Method of Using the Noise Generator to measure the Noise Factor of a Receiver

It is essential, when a noise factor measurement is to be made, to ensure that the measuring system is linear, as the definition of noise factor is based upon this assumption. The receiver under test may be assumed to be linear for this purpose if doubling the input power produces a corresponding doubling of the output power. This relation may

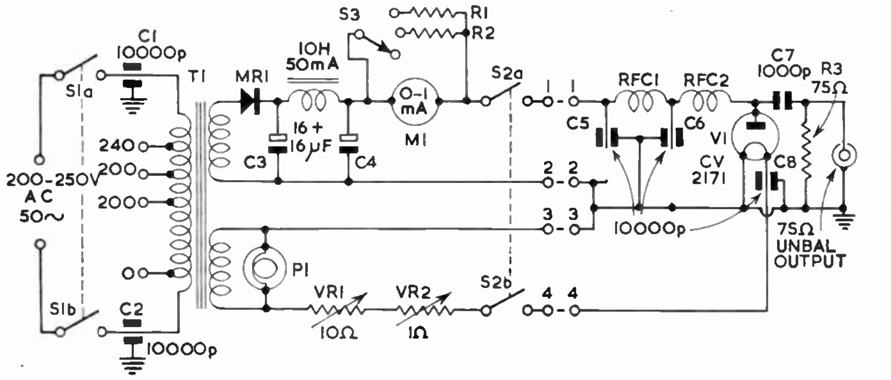


Fig. 19.48. Noise generator using CV2171, suitable for unbalanced input. C1, C2, C5, C6, C8 are 10,000 pF lead-through capacitors. C1, 1000 pF tubular ceramic; VR1, 10 ohms 5 per cent wire wound linear (Reliance type TW); VR2, 1 ohm 10 per cent wire wound linear (Reliance type TW); R3, 75 ohm 5 per cent coaxial resistor (Welwyn type DSR/4/1). T1, 125 V 50 mA, 4·5 V 1A; S1a, b, S2a, b, d.p.s.t. toggle; S3, s.p. three way wafer; RFC1, 2, 50 turns 30 s.w.g. close wound on 3/8 in. former; MR1, S.T.C. type C2H.

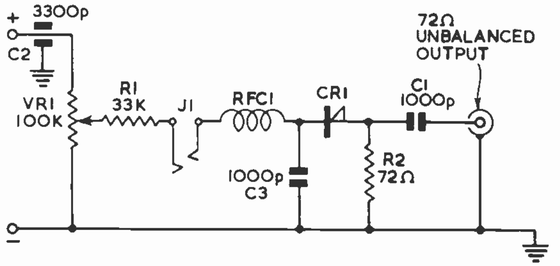
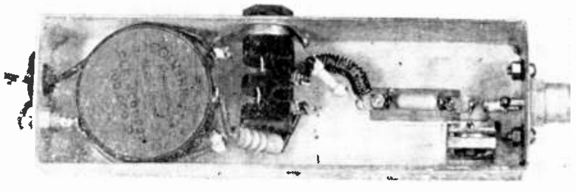


Fig. 19.50. Circuit of a germanium diode noise generator, C1, C3, 1000 pF tubular ceramic; C2, 3300 pF lead-through; CR1, GD8, 1N21 or equivalent; R1, 33 K ohm 10 per cent ¼ watt; R2, 72 ohm 5 per cent coaxial resistor (Welwyn type DSR/4/1); VR1, 100 K linear wire wound potentiometer (Colvern); RFC1, 25 turns 22 s.w.g. enam. air wound ¼ in. i.d. The germanium diode CR1 should be a low reverse voltage type.

not hold over a wide range of signal levels and settings of the receiver gain controls, but provided that the measurement is carried out at signal levels and gain settings at which the receiver is known to be linear then the conditions imposed by the definition of noise factor which follows will be satisfied. The *noise factor* of a linear receiver, or any other four terminal network having its input terminals connected to an impedance of stated value and temperature, is the number of times by which an addition to the noise power available from this impedance must exceed the thermal noise at signal frequency in order to double the noise power available from the output terminals of the network, provided that all sources of noise give the same frequency spectrum at the output terminals.

Two methods of using the noise generator to measure the noise factor of a receiver will be described. The first is a simple method, involving one measurement only of noise diode anode current while the other involves three such measurements and a little more calculation, but has the advantage of incorporating a linearity check.

Both methods require some form of power measuring device which is connected to the receiver at a suitable point to measure the a.f. noise output. In the event that an a.f. output meter is not available, a low range a.c. volt meter may be connected across the primary of the receiver output transformer. Alternatively an 0–100 μA d.c. meter may be placed in series with the earthy end of the detector diode load. A fairly heavily damped meter will assist in reducing the fluctuations observed when reading noise output. It should be remembered that power is proportional to voltage or current squared so that to double the power output when reading in microamperes or volts the meter reading should increase by a factor of $\sqrt{2}$.



A view inside the germanium diode type noise generator of Fig. 19.50.

In the first method of measuring the noise factor the noise generator is connected to the input terminals of the receiver but not switched on. The i.f. and a.f. gain controls are adjusted to give a convenient reading on the output meter. If a d.c. microammeter in the detector diode load is being used, the a.f. gain control is ignored and the i.f. gain control adjusted to give a reading of 20–40 μA. The noise diode is then switched on and the anode current increased until the receiver power output is doubled or the detector diode current increases by $\sqrt{2}$ times. The noise factor of the receiver in decibels is given by $F\,db = 10\,log_{10}\,\frac{20\,IR}{1000}$ where I is the noise diode current in milliamperes and R is the value of the noise generator source impedance.

The procedure described above is only valid if the receiver is known to be linear over the range of input levels involved in making the measurement. If there is any doubt then the three measurement method should be used as follows:

With the noise generator connected but not switched on the receiver gain is adjusted to give a convenient meter reading, P_1, due to its own noise. The noise generator is then switched on and the noise diode anode current adjusted to give a considerable increase in receiver noise power output, but not enough to cause saturation. Let this output power be P_2 and the noise diode anode current required to produce it I_1. The gain is now reduced appreciably, and the noise diode currents I_2 (usually made equal to I_1 by suitable adjustment of the gain control) and I_3 required to give the original output readings P_1 and P_2 are determined, and the three values of I substituted in the following formula:

$$F(db) = 10\,log_{10}\,\frac{20\,I_1\,I_2\,R}{I_3 - I_2 - I_1}$$

where R is the noise generator load impedance and I_1, I_2 and I_3 are in milliamperes. It should be noted that if $I_1 + I_2$ is nearly equal to I_3, a small error in readings gives a large error in F, and this condition should be avoided by making P_2/P_1 as large as possible.

The equivalent noise temperature of the receiver is given by $Tr = (F - 1)T_0$ where F is the numerical value of noise factor and T_0 is the equivalent noise temperature of the load resistor, which may be assumed to be 290° K (°K = °C + 273). The numerical value of the noise factor F is a ratio and not the figure given in db quoted in the valve makers' data. In order to substitute for F in the equation for equivalent noise temperature, it is necessary to take the antilogarithm of the quoted noise factor in db divided by 10, i.e. numerical value = antilog $\frac{F(db)}{10}$.

Using the Germanium Diode Noise Generator

The germanium diode noise generator cannot be used to measure the absolute value of noise factor as already stated but it is very useful as a means of assessing improvements to receiver aerial coupling circuits and r.f. stages. When used in this manner the generator is connected to the receiver and the increase in noise output produced when the generator is turned on is recorded. Next the desired modification is made and the procedure repeated. If after the modification less diode current is required to produce the same increase in noise output from the receiver, then it can be assumed that an improvement in noise factor has resulted. The converse is also true.

Noise Generation above 500 Mc/s

Current practice at frequencies in excess of 500 Mc/s is to use an Argon-filled fluorescent discharge tube as a noise source, the noise power being extracted by an external coupled helix terminated at one end by a load resistor equal to the required source impedance, while the other end is connected to the receiver under test by a short coaxial cable of suitable impedance. When the tube is conducting, it generates noise at an equivalent noise temperature of 10,100° K, i.e. approx. 15·5db excess noise power, which is fixed by the type of tube. A variable attenuator is employed at i.f. to reduce the gain of the receiver under test until the noise power output of the receiver is the same with the fluorescent noise source switched on or switched off. The value of the noise factor F in db using a fluorescent noise source is given by $F(db) = 10 \log_{10}(C-1)\left(\frac{P_{ON}}{P_{OFF}} - 1\right)$

where C = the ratio of the noise temperatures of the source when switched on and switched off, and P_{ON}/P_{OFF} is the loss inserted by the attenuator to maintain the receiver noise output constant. Corrections are applied which allow for the incomplete coupling of the fluorescent tube to the helix and also for the mismatch introduced by the tube and helix in the extinguished condition. If the receiver has an appreciable response at the image frequency a further correction is necessary. This last correction will apply to the measurement of noise factor by any form of noise generator and amounts to as much as 3db if the image response is equal to the required response.

CATHODE RAY OSCILLOSCOPES

The oscilloscope is one of the most versatile instruments an amateur can possess and permits the visual display of a.f. and r.f. signals. It is particularly useful for monitoring phone transmissions.

Commercial instruments are often unnecessarily elaborate for amateur purposes but it is perfectly feasible to build simple units at home.

The heart of the oscilloscope is the cathode ray tube. If it is desired to examine a waveform, the input is connected to the vertical plates while a sweep or time base circuit is connected to the horizontal deflection plates. For many applications it is sufficient to derive the sweep voltage from the a.c. mains.

By using readily available and inexpensive tubes, and instrument designed expressly for 'phone monitoring may be constructed quite cheaply. Of the government surplus tubes the 3BP1 has features which make it perhaps the most suitable. Its screen is large enough to give an unambiguous picture, and although it is a short and handy tube, it has a reasonably high deflection sensitivity so that a high input is not required. This obviates the need for deflection amplifiers in monitor service. A number of tubes of this type have been tested and all were found to have the electron guns so well centred that the spot appeared in the centre of the screen without the assistance of a shift-control network. This simplifies construction and reduces cost. The anode voltage requirements are modest; many tubes will give a quite well focused spot at voltages as low as 400-500, and nearly all will work excellently at 700 volts and above.

For transmission monitoring it is sufficient to provide a heater supply only, because the existing power supply for the transmitter final stages will almost invariably be capable of providing the few milliamperes drawn by the cathode ray tube and its control network. It is unfortunate that external fields are liable to cause trouble unless the final anode and the deflector plates of the tube are operated at earth potential. It is therefore highly desirable to provide the few components needed to furnish a negative h.t. supply for the cathode.

MONITORING DISPLAY UNIT

The circuit of a cathode ray oscilloscope embodying the foregoing principles is given in Fig. 19.51. This, though deliberately modest, is adequate for setting up s.s.b. transmitters and monitoring a.m. transmitters. The lead from the heater of V1 may be connected to either end of the high-voltage secondary of the p.a. power supply transformer, and after smoothing in the RC filter comprised of C5, R8 and C4, a negative d.c. output of almost 1·4 times the r.m.s. half-secondary voltage will be provided for the cathode-ray tube. Due to the negligible current drain, RC filtering is adequate. Brilliance of the pattern is controlled by VR3, and focus by VR2. It should be noted that a high-intensity spot must not be permitted to remain stationary on the screen for any length of time, otherwise it will leave a permanent blemish on the fluorescent phosphor. Horizontal sweep is taken from the 50 c/s a.c. mains It is therefore necessary to make sure that C1 is connected to the live side of the mains supply. If no sweep is obtained on switching on, it is clear that C1 is connected to the neutral lead, and the fault should be rectified either by transferring it to

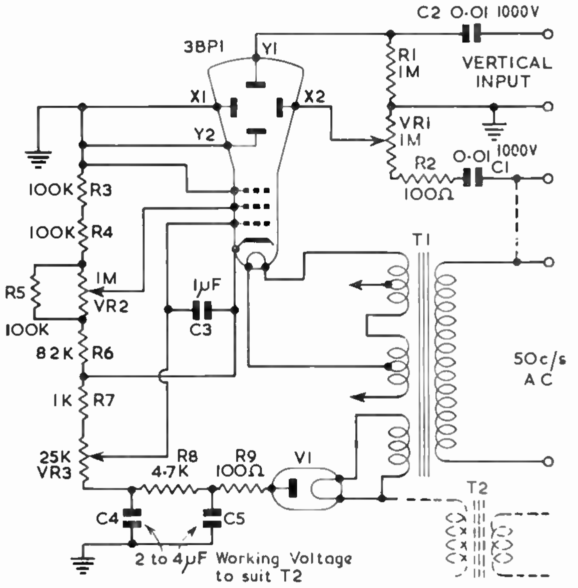


Fig. 19.51. Oscilloscope display unit for monitoring telephony transmissions. T1 must be adequately insulated for the voltage in use. T2 is the existing p.a. power supply transformer (see text). V1 may be a U27 or other suitable rectifier.

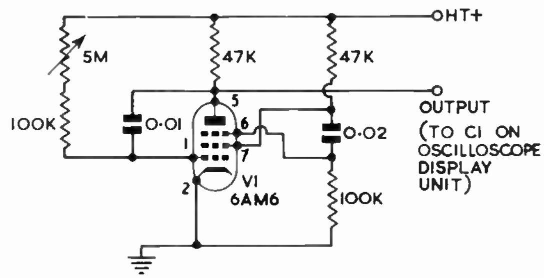


Fig. 19.52. Miller-transitron sawtooth time base generator for use with the oscilloscope of Fig. 19.51.

the other side of the heater transformer primary or by reversing the mains plug in its socket. A 50 c/s sweep causes some distortion of the display, but this may be kept within reasonable limits by using a high sweep amplitude so that only the centre portion of the sweep appears on the screen. The resulting trace is linear enough for most monitoring work. The sweep amplitude may be reduced by VR1, but this control should be set at maximum when monitoring is being carried out.

If other work is contemplated, it is worth including some form of linear horizontal sweep. The Miller-transitron oscillator, shown in its most elementary form in **Fig. 19.52,** has much to commend it. Any sharp cut-off pentode may be used in this circuit, so long as it has its suppressor grid brought out to a separate pin. As the valve requires only a few milliamperes of h.t. at 200-300 volts, it will probably be possible to "borrow" the necessary supply from an

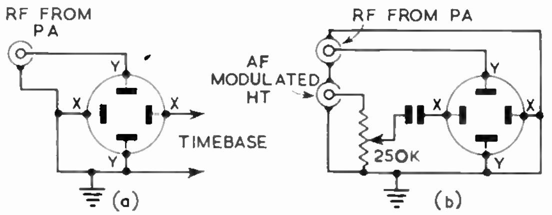


Fig. 19.53(a). Oscilloscope connections for envelope pattern. (b) Connections for trapezoidal pattern.

existing power-pack. If the linear sweep of Fig. 19.52 is preferred to the 50 c/s mains sweep, C1 in Fig. 19.51 should be disconnected from the primary of the heater transformer and connected to the anode of the oscillator valve.

The small $\frac{1}{10}$th watt resistors R2 and R9 are the only components which remain to be explained. They are included merely to act as fuses in the event of other components breaking down.

This simple display unit can be employed for monitoring a.m. transmissions using the method described, with illustrations of typical patterns, in Chapter 9 (*Modulation*). Connections for envelope and trapezoidal patterns are shown in **Fig. 19.53.**

Calculation of modulation depth using the trapezoidal method is possible by measuring the sides of the trapezium and using the abac of **Fig. 19.54.** A disadvantage of the trapezoidal method is the necessity to tap off from the secondary of the modulation transformer an audio signal for the *X* (horizontal) plates. However, this can be overcome by building a small tuned circuit and demodulator unit (**Fig. 19.55**) which can be connected to the output of the transmitter to provide the necessary audio signal. The unit should be made up on a small sub-chassis and fitted as near to the

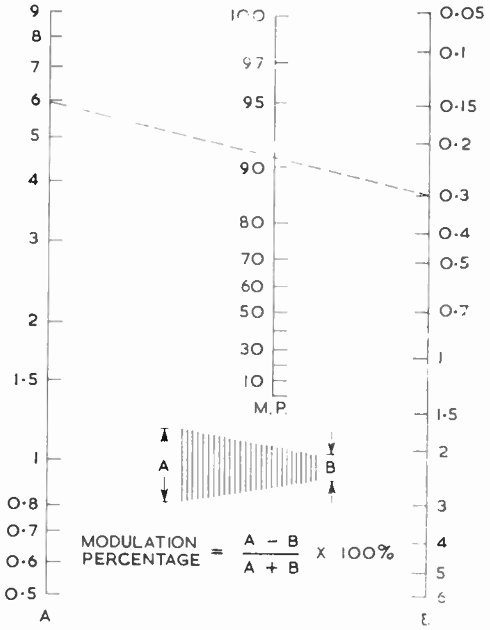


Fig. 19.54. Abac for the calculation of modulation depth from the trapezoidal pattern. The dotted line illustrates an example in which the large side (A) is 6 units long and the shorter one (B) 3 units indicating a depth of modulation of just over 30 per cent.

holder of the cathode ray tube as possible. If the signal is fed to the *X* amplifier input the trapezoidal display will be obtained. When it is fed to the sync. terminal on an oscilloscope the waveform pattern will be shown.

S.s.b. transmissions may be conveniently checked by the 45° method which calls for the application of similar voltages from the input and output of the linear amplifier. This necessitates the use of two detectors, as shown in **Fig. 19.56,**

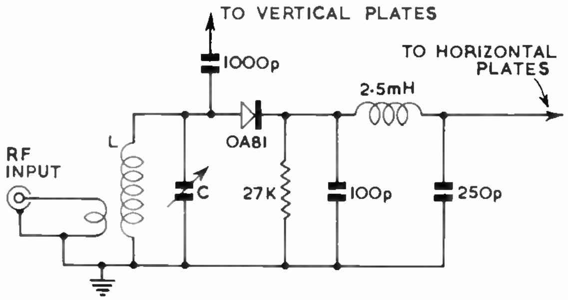


Fig. 19.55. Demodulator unit to provide an audio signal for connection to the X plates of the oscilloscope in the trapezoidal method of measuring modulation depth.

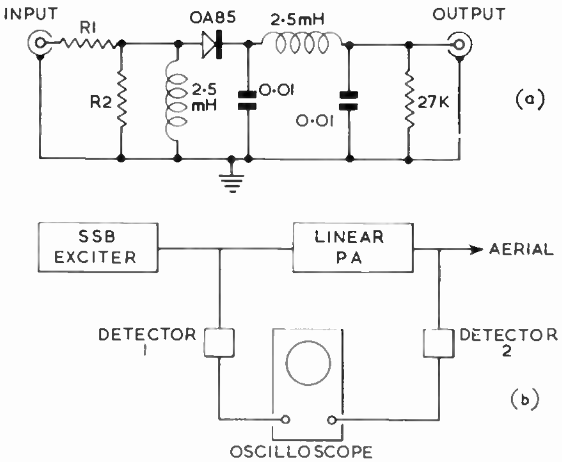


Fig. 19.56(a). Demodulator unit. (b) Circuit arrangement for checking s.s.b. transmissions by the 45° method.

the resistors R1 and R2 should be adjusted so that the voltage output from each is similar. The interpretation of the various displays obtained when a sinusoidal tone is fed into the s.s.b. transmitter is shown in **Fig. 19.57.** It should be noted that the use of the demodulator of Fig. 19.56 does not permit checks on linearity to be made.

BUILT-IN MODULATION MONITOR

Availability of the 1CP1 c.r. tube makes it possible to build a monitor into a transmitter, the display screen taking its place on the front panel with the grid and anode current meters. A suitable circuit, provided by Electronic Tubes Ltd., is shown in **Fig. 19.58.** It comprises a small cathode ray tube. a time base generator and a phase splitting *X* amplifier. This type of monitor may be used to display the modulation envelope or the modulation trapezium, the display being selected by means of S1.

While the *Y* deflection of the tube is simplified in this case by the fact that the tube is designed for asymmetric *Y* deflection, the *X* deflection will still require symmetrical deflection. (Asymmetric *X* deflection could be used, directly from the time base, but would probably introduce trapezium distortion on the display. This could be misleading when displaying the modulation trapezium.) For this purpose the phase splitting amplifier V2 is introduced. This amplifier also serves the purpose of amplifying the modulating waveform—which may be taken at a low level—when the trapezium is to be displayed.

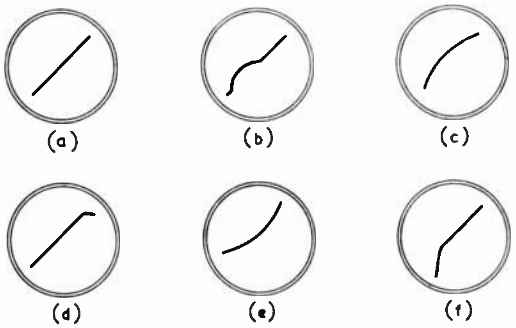


Fig. 19.57. Interpretation of displays obtained with the arrangement of Fig. 19.69. (a) Linear condition. (b) Incorrect grid circuit loading. (c) Incorrect bias. (d) Amplifier overloaded. (e) Insufficient standing current in amplifier. (f) Pattern obtained from speech input to a correctly adjusted linear amplifier.

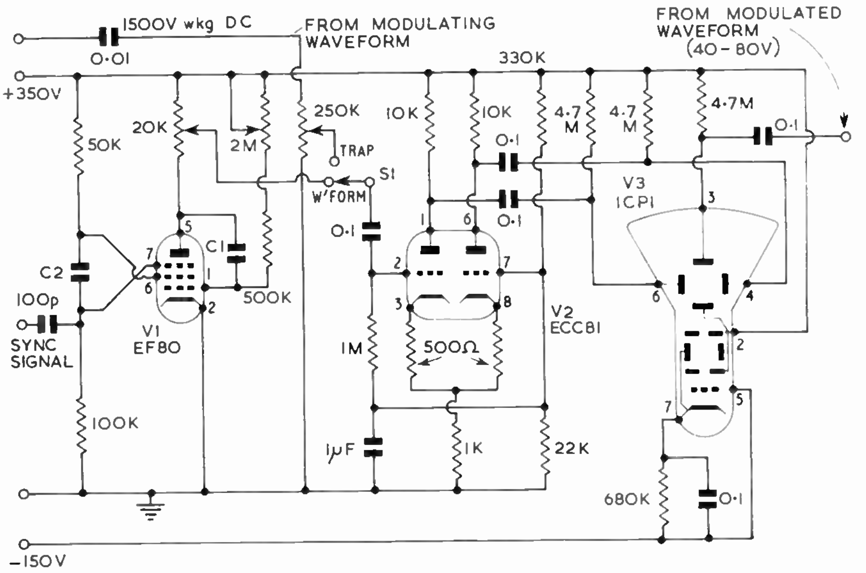
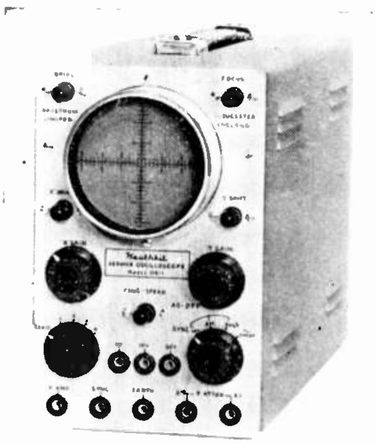


Fig. 19.58. Modulation monitor employing a 1CP1 tube. The unit is suitable for building into a transmitter. The time base speed depends on the value of C1 which may be 300 pF (1·2-6k/cs), 1000 pF (400 c/s to 2 kc/s) or 3000 pF (120 c/s to 600 c/s). C2 should be five times the value of C1.

The time-base generator V1 is a transitron-Miller circuit which in spite of its simplicity, provides a linear waveform. Synchronization of the time base sweep may be effected by applying a small amount of the audio to be displayed through a small capacitance to the suppressor grid of the pentode, provided that the time-base repetition rate and the modulating frequency are not very greatly different.

The cathode ray tube 1CP1 operates from a 500 volt supply and, therefore, requires an extra 150 volts in addition to the 350V h.t. line. The arrangement shown in **Fig. 19.59** is a simple and inexpensive means of obtaining this voltage. Operation of the tube is simplified by the use of automatic biasing in place of the usual brilliance

The Heathkit Model OS-1 oscilloscope. The complete circuit diagram is shown in Fig. 19.61.



control and by the fact that the tube is designed to focus automatically.

If the monitor is to be used in an s.s.b. transmitter, the circuit can be further simplified by omission of the time-base generator.

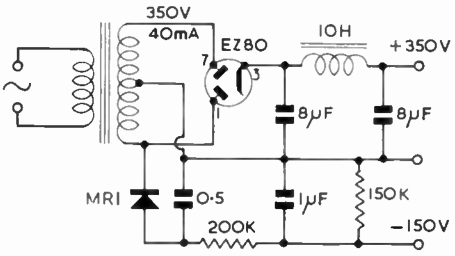


Fig. 19.59. Power supply for the modulation monitor of Fig. 19.58

SERVICING OSCILLOSCOPE

The circuit diagram of a complete oscilloscope employing a 2¾ in. tube type 3AFP1 is shown in **Fig. 19.61.** This instrument, the Heathkit Model OS-1, has *X* and *Y* amplifiers and a Miller transistron time-base generator covering 15 c/s to 150 kc/s. The frequency reponse of the *X* amplifier is ± 3db from 150 c/s to 500 kc/s and the *Y* amplifier ± 3db from 10 c/s to 2·5 Mc/s. It is thus suitable for all the work likely in the amateur station—waveform investigation, testing audio amplifiers and servicing communication and television receivers.

Two input sockets for the *Y* amplifier are provided, one being direct, the other introducing an attenuation of 10 : 1. The first section of V1 is a cathode follower, gain being controlled by VR1 (Y GAIN). Vertical positioning of the tube trace is provided by VR2.

V3b is used as a sync. amplifier, the valve operating with zero bias and producing at its anode an amplified and limited signal of the sync. input. This output is positive going and is fed to the suppressor grid of the time-base generator V4. three methods of synchronization may be used: internal, external or 50 c/s. Internal synchronization is obtained from the cathode of V1 through an isolating resistor R52.

Coarse frequency control of the time base (V4) is provided by the ganged switches S_B and S_C and fine speed control by VR4. Output from the anode of V4 is taken through an isolating capacitor to the X AMP socket on the front panel so that while the time base is operating a saw tooth waveform is available at this point. One of the positions of S_B and S_C allows the time base to be switched off and an external signal or time base fed into the *X* amp socket. Amplitude of the signal is variable by means of VR3. Output from VR3 (X GAIN) is fed to V5 connected as a cathode coupled pair. Horizontal positioning of the trace is adjusted by VR5 (X SHIFT). A 50 c/s sinusoidal time base is provided by a position on switch S_A which switches out the time base and connects V5 to a source of 50 c/s. Calibration voltages of 1, 10, and 50 volts peak-to-peak at 50 c/s are available from sockets on the front panel.

E.h.t. supplies for the tube are derived from a resistor chain across the supply from the rectifier MR1 (type K8/40 selenium). The maximum voltage fed to pin 8 via VR7 is 700 volts. VR6 is a spot shape or astigmatism control and is adjusted in conjunction with VR7 (BRILLIANCE) and VR8 (FOCUS) to produce a well-defined trace. Flyback suppression is achieved by taking a negative pulse from the screen of V4 This pulse is limited by CR1 and the resulting flat topped waveform fed to the grid of the tube to blank the retrace or flyback of the time-base sweep.

MODULATION MONITOR FOR USE WITH A RECEIVER

The circuit of a modulation indicator for use with a receiver is shown in **Fig. 19.60.** L1 and L2 are the coils of an i.f. transformer of the same frequency as the receiver intermediate frequency. The signal is fed through a small capacitor connected to the final i.f. valve via coaxial cable to point *A*, and to Y1 via the 0·005 uF capacitor. With S1 in position 1, the signal is fed inductively to the *X* plate, with 90° phase shift, thus creating a circular trace for every carrier tuned in. Depth of modulation is shown by the extent to which the circular trace is thickened inwards and outwards, full modulation being shown as a disc. Overmodulation will produce brightening at the centre due to trace overlapping.

When an a.c. source is connected to position 2 of S1, and with the switch in this position, the envelope trace for modulation results.

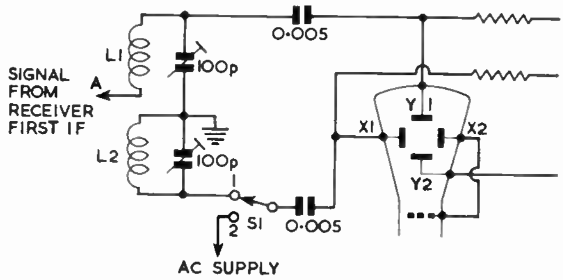


Fig. 19.60. Modulation monitor for use with a communications receiver L1 and L2 are the windings of an i.f. transformer of appropriate frequency.

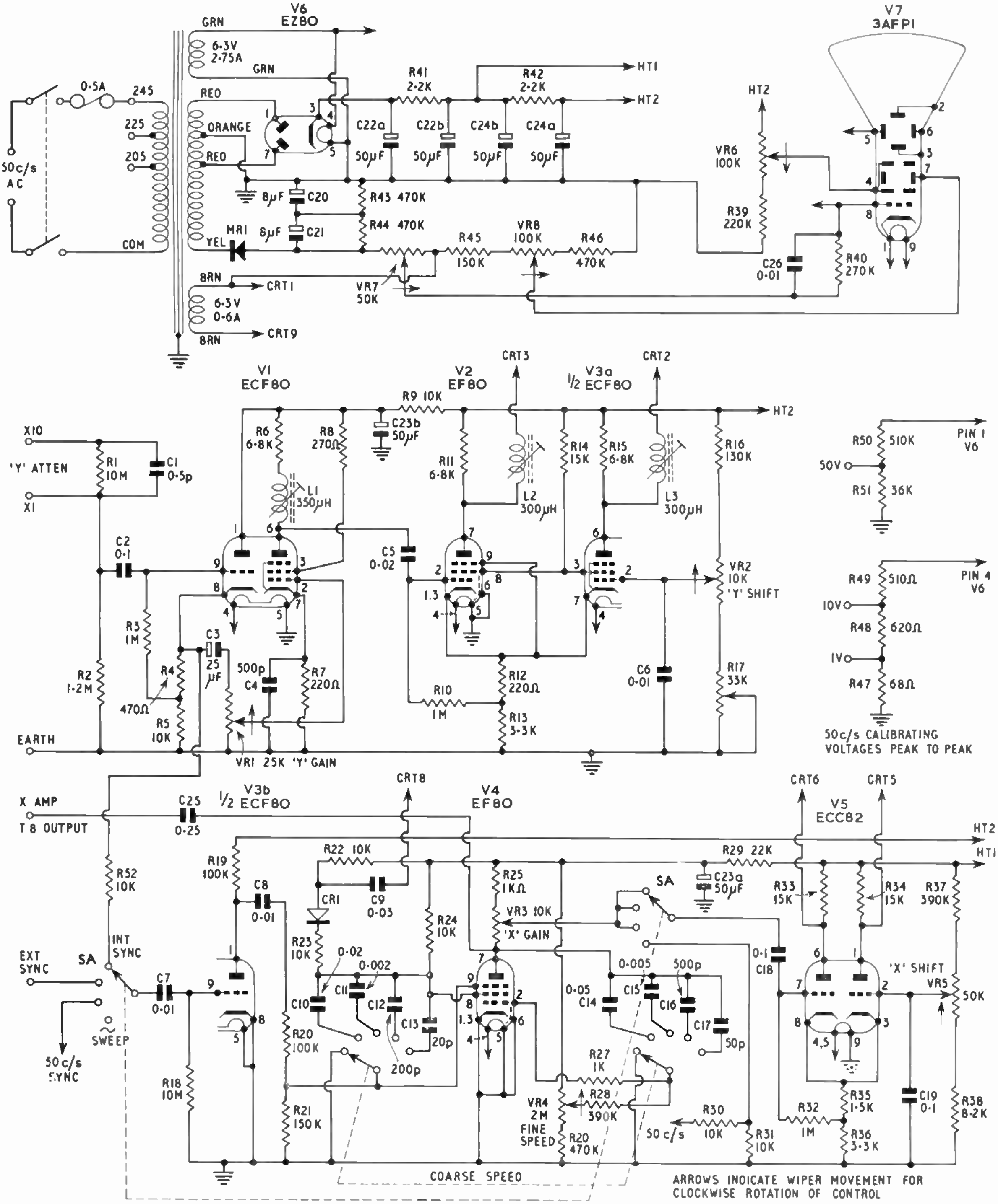


Fig. 19.61. Circuit diagram of the Heathkit Model OS-1 cathode ray oscilloscope. *(By courtesy of Daystrom Ltd.)*

On setting up the monitor, a strong steady carrier should be accurately tuned in, and the 100 pF trimmers adjusted for the best circular trace, although it may be found that the trace is rather elliptical. The cores of the i.f. transformer should be initially withdrawn sufficiently for the tuning to be carried out with the 100 pF trimmers only. The 0·005 μF capacitors should be rated to suit the tube voltages and the a.c. supply at position two of the switch. For safe working, if d.c. voltages only are shown on capacitors, this figure should be divided by three to obtain the safe a.c. working voltage.

OTHER APPLICATIONS OF THE OSCILLOSCOPE

It is important to realize that an oscilloscope is virtually a two dimensional meter, i.e. it is intended for making *measurements*, not simply for displaying an interesting picture. Consequently the temptation to continually adjust the X and Y sensitivities to make the picture fill the screen should be firmly resisted. If the Y gain is continuously variable it is worthwhile calibrating it against a voltmeter on 50 c/s a.c. (remembering that the meter reads R.M.S., not peak-to-peak, 1 volt R.M.S. $= 2\sqrt{2}$ volts peak-to-peak) so that the settings for say 1, 3, 10, 30 and 100 volts/cm, can be found rapidly. Time calibration can be similarly checked against a crystal calibrator, with a divider to 10 and 1 kc/s and the 50 c/s mains. If the oscilloscope is provided with a triggered time base it is convenient to calibrate in terms of time/cm, but with the older slightly less convenient synchronized time base this has to be variable to synchronize, so a frequency (or period) calibration is more suitable.

The uses to which an oscilloscope can be put can be conveniently classified by the inputs to the Y and X plates.

(a) Signal to Y, nothing to X. This mode uses the electron beam as a meter pointer, and does not fully exploit an oscilloscope, but is convenient when a valve voltmeter would otherwise be required, as for example measuring the d.c. anode potential of a high gain audio amplifier. In practice, there is no need to switch off the timebase for d.c. measurements, although this may be convenient to measure the peak-to-peak amplitude of an a.c. signal. Care should be taken to avoid burning the screen in either case.

(b) Signal to Y, timebase to X. The additional dimension of time reveals much more information about a signal. The frequency can be measured against a calibrated timebase, and any distortion or unwanted signal traced. The oscilloscope, like the camera, can however lie. The input capacitance of the average Y amplifier is some 30pF and this, if connected to a high impedance point, can severely restrict the bandwidth. The use of coaxial cable leads with an additional capacitance of 22 pF per foot will make this much worse; sometimes this effect can be used to remove unwanted r.f. from a low frequency signal, but in general it is preferable to use either a high impedance probe unit, or failing this, short leads of unscreened wire.

The bandwidth of an oscilloscope amplifier may be quite low, so that measurements at, say, 28 Mc/s cannot be made. Often it is possible to work slightly outside the nominal frequency range if a reduced sensitivity is tolerable, but this should be done with caution since excessive signal outside the bandwidth can cause distortion and even modulate the timebase speed. Many oscilloscopes have provision however for inputs direct to the Y plates, via a blocking capacitor, thus extending the usable bandwidth although with reduced sensitivity. As the frequency becomes higher the period of the wave becomes comparable with the transit time of an electron between the Y plates and the deflection sensitivity falls.

(c) One signal to Y, another to X. The most familiar form of this mode is the trapezoidal display of a modulated amplifier; but the input and output of any device may be so displayed. The input/output linearity of a diode detector can therefore be displayed and compared with a product detector, taking care to avoid leakage of the b.f.o. signal into the oscilloscope. The maximum undistorted output from an amplifier can be measured by increasing the drive till the diagonal line on the screen starts to curve.

The wobbulator, or swept alignment oscillator, is another application in which the X signal is related to the frequency of an oscillator, the output of which is fed to the r.f. amplifier under test, and the amplifier output, possibly after detection is fed to the Y amplifier. Provided the rate of frequency sweeping is sufficiently slow, a graph of the amplifier response is presented on the cathode ray tube.

Finally, a word of caution. It has been known for the timebase harmonics of an oscilloscope to cause TVI. This possibility should be borne in mind and suitable screening and lead filtering applied if necessary.

MISCELLANEOUS TEST GEAR

P-N-P AND N-P-N TRANSISTOR TESTER

A test set providing quick and accurate direct measurement of current gain and collector leakage current of *p-n-p* and *n-p-n* transistors under static, grounded emitter conditions is a useful tool in the modern radio workshop. The unit described here will also give a good indication of the state of diodes when compared with a similar type in good working order.

A simple form of protection is included which will guard against damage if a transistor with an internal short is connected for test. Collector leakage, I'_{co}, is read on a 0-1 mA meter. The current gain (*beta*) is arranged in two switched ranges: 0-100 at 100 μA base current and 0-250 at 40 μA base current. On *beta* test, the meter is shunted to read 10 mA full-scale deflection (f.s.d.).

Testing a transistor under grounded emitter conditions is somewhat similar to testing a valve for mutual conductance, the emitter, base and collector of the transistor corresponding to the cathode, grid and anode of the valve. The base is the controlling element: if there is a small change in base current there will be a much larger change in the collector current. the grounded emitter current gain, termed *beta* or α' is the ratio of the collector and base currents, and the measurement of it is here simplified by using a defined base current and indicating the collector current on the meter.

Fig. 19.62 is the complete circuit of the tester. A 0-1 milliammeter is used for indicating the collector leakage current and was chosen as a good compromise between sensitivity and expense.

The operation is as follows: connect a transistor to the appropriate terminals, close switch S1 and read off the collector leakage current. (The value of this will normally be under 100 μA.) Next close switch S2, allowing a base current of 100 μA to flow, and then read off the collector current. A full scale reading will indicate a gain of 100 (the

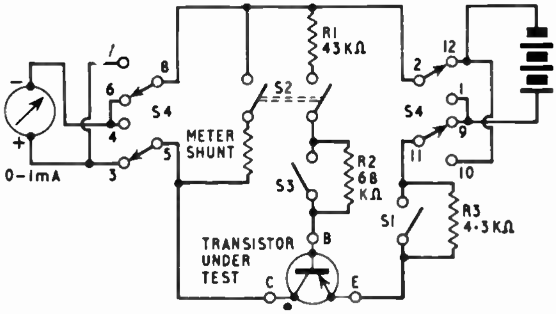


Fig. 19.62. Complete circuit diagram of the transistor tester.

ratio of 10 mA collector current, divided by 100 μA base current). An OC71 for example may give 5 mA, thus indicating a gain of 50, and similarly for other transistors.

To enable *n-p-n* transistors to be tested, a four-pole changeover switch is used to reverse the battery and meter connections. Some transistors such as the OC44 have a high gain and a small allowable dissipation. R2, in conjunction with R1 (68 K and 43 K ohms respectively), limits the base current to 40 μA. In this position, full scale deflection of the meter indicates a gain of 250 (10 mA collector current divided by the 40 μA base current is equivalent to a gain of 250). The purpose of R3 (4·3 K ohms) is to limit the meter to full scale deflection should a transistor with an internal short be connected. In this case, the push switch must *not* be pressed, thus avoiding meter overload, and the faulty item should be discarded.

Some readers building this tester may wish to use a meter which they have on hand. In such cases the required shunt-resistor can be calculated from the formula on page 19.1. With a 0-1 mA meter the scale multiplying factor is 10. With a 250 μA meter it would be 40.

The limiting resistor R3 should be changed to 8·2 K ohms if a 0-500 μA meter is used and to 15-18 K ohms with a 0-250 μA movement.

The layout is shown in **Fig. 19.63,** the components used fitting comfortably into an Eddystone diecast box.

The switch wafer is wired as shown in **Fig. 19.64** but on the under side for neatness. The contacts should be of the non-shorting type; that is, break before make.

R1 and R2 are standard value ¼-watt resistors of close tolerance and high stability. If another type of resistor is used, any small variation from the values quoted should be toward a lower resistance, certainly not a higher. The meter shunt resistor must be selected carefully if accuracy of measurement is to be obtained. The meter and its shunt should be compared with an instrument of known accuracy before fitting into the tester.

Operation

Transistor Test. Open both toggle switches, set switch S4 to the *p-n-p* position. In the first instance only, use a voltmeter to check that the collector terminal is negative with respect to the emitter terminal. This being so, connect a *p-n-p* transistor, such as an OC44 or OC71. If the meter reads full scale, an internal short is indicated and the transistor is unserviceable. If the meter is not indicating full scale deflection, press the button and read off the collector leakage current directly on the 0-1 mA scale. As transistors of the germanium type are very temperature conscious, an abnormally high leakage, perhaps double the normal, will be indicated if the transistor is handled for too long with warm hands especially near a fire or a hot radiator. The transistor should be at normal room temperature when tested.

Next close switch S2 and read off the current gain on the 0-250 scale. If the gain is under 100, close switch S3 and read on the 0-100 scale. It is quite normal for the transistor gain to be a little higher (or lower in some cases) on the 0-100 scale due to operation on a different part of its characteristic curve. (This curve is not quite linear.)

For *n-p-n* transistors, set switch S4 to the *n-p-n* position and carry out the testing procedure as above.

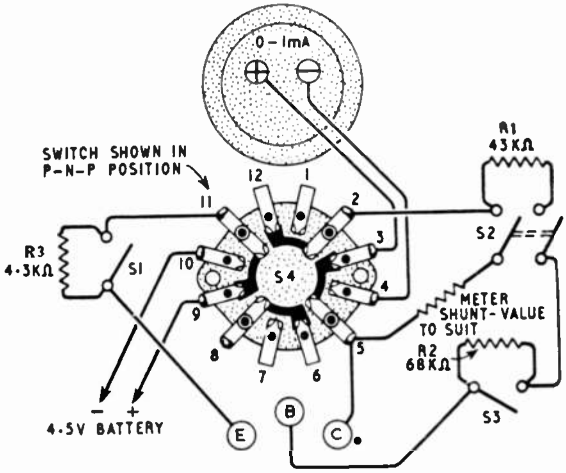


Fig. 19.63. The layout viewed from the rear of the diecast box.

Diode Testing. With switch S4 in the *p-n-p* position and switch S2 open, connect the cathode (bar and red end) of the diode to the collector terminal and the anode to the emitter terminal. The meter should give a full scale deflection, indicating a low forward resistance. Move switch S4 to the *n-p-n* position (this reverses the voltage on the diode) and the meter should drop to zero, indicating a high back resistance. As this is only a simple check, and not a measurement, it may be advisable to compare the results with a diode of the same type which is known to function correctly. In this test, the push button S1 is not used.

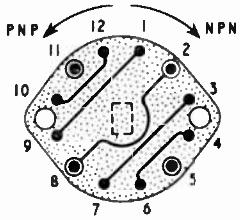


Fig. 19.64. Wiring of the wafer switch.

OPERATING TECHNIQUE AND STATION LAYOUT

SUCCESS in any branch of Amateur Radio depends primarily on the operator's ability to use his equipment correctly and his knowledge of correct operating technique. However much time or money has been spent on equipment, results will depend on the way in which the station is operated. The two basic principles of good operating are applicable to any mode of operation and are quite simply *judgment* and *courtesy*.

While studying to pass the Morse test the newly licensed amateur will probably have spent a considerable time listening on the various amateur bands and will have noticed the widely differing standards of operating technique to be heard. The good operator is not necessarily the man who can send fast, but the man whose sending is easy to read. Automatic Morse, derived from an automatic keyer which delivers perfectly formed characters, is the easiest to read, and therefore the amateur should always strive to send Morse that sounds as if it were being transmitted by an automatic sender. The ability to listen to one's own transmission is essential if a good style is to be achieved and maintained: this practice is sometimes known as *side-tone*.

Similarly, in voice operation care should be taken to speak as clearly as possible; this includes speaking slowly and choosing simple " basic " words when the other operator is obviously not fluent in the language being used.

An amateur should never send c.w. at a faster rate than he knows he can send perfectly, because a very quick deterioration results from even slightly exceeding his maximum speed capability. The best speed is the maximum speed at which the other operator can receive the whole of the message being transmitted. It is generally a good principle to send at the speed used by the station being worked.

DX Operating

In order to be successful in working DX it is necessary to spend much time listening, learning the operating habits of stations in the parts of the world with which contacts are desired. It is usually much more rewarding to call specific stations, rather than to call CQ. However there are exceptions to this rule, for example when a band appears to be inactive. In these circumstances it is a good idea to put out a CQ call as this may reveal the fact that the band is in fact open but that everyone is listening! When about to call CQ, find a quiet frequency, listen there for a short while, and then make a short transmission. This is very important: many contacts have been lost because the station at the far end tired of waiting for the end of a CQ call. A satisfactory CQ would consist of:

CQ CQ CQ de G2XYZ G2XYZ CQ CQ CQ de G2XYZ G2XYZ CQ CQ CQ de G2XYZ G2XYZ AR K

After sending this a very careful check should be made of the frequency for possible callers; if there are none, the procedure may be repeated, at the same time making sure that there is no interference on the frequency. When it is desired to make contact with a specific area, a directional CQ call may be made:

"CQ VK CQ VK CQ VK de G——".

This should result in a reply from a station in the desired country only. Conversely, never call another station who has just put out a directional call, unless situated in the area he is calling.

When answering a CQ make short transmissions, sending the call-sign of the station being called not more than four or five times followed by one's own call two or three times only. If there is no reply this should be repeated. If long calls are made the other station may well be half way through and over by the time the call ends. A fast operator may even have finished his first transmission.

Most DX stations find that once they have made an initial CQ call, they have a number of callers standing by waiting for contacts and therefore there is no need to put out any more calls. This means that those who call intelligently will obtain a contact more quickly. It is a good plan to try to find the frequency of the station being worked, and set the transmitter near to this spot, then when the QSO is over a short call will often result in a contact. Unfortunately this is not always possible, and in this case a call just off the DX station's frequency may be successful. In the case of some of the more proficient DX operators the listening frequency will be announced from time to time. On c.w. this is usually signified by " U 5 " (meaning " I am listening 5 kc/s higher than my own frequency "), or " L 5 " (which means " I am listening 5 kc/s lower than my own frequency "). Under these circumstances, calls should never be made zero beat. Under " DXpedition " conditions, when the DX station is anxious to give as many contacts as possible in limited time, it is most important to keep calls very short, putting the emphasis on one's *own call*, not on the call of the DX station. When contact is established the exchange should be limited to reports only, unless indicated otherwise. If these ideas are translated into practice it will be found that the ultimate in equipment is not needed, and that common sense and courtesy will reap a rich reward.

TABLE 20.1

Morse Code and Sound Equivalents

A	di-dah	**S**	di-di-dit
B	dah-di-di-dit	**T**	dah
C	dah-di-dah-dit	**U**	di-di-dah
D	dah-di-dit	**V**	di-di-di-dah
E	dit	**W**	di-dah-dah
F	di-di-dah-dit	**X**	dah-di-di-dah
G	dah-dah-dit	**Y**	dah-di-dah-dah
H	di-di-di-dit	**Z**	dah-dah-di-dit
I	di-dit	**1**	di-dah-dah-dah-dah
J	di-dah-dah-dah	**2**	di-di-dah-dah-dah
K	dah-di-dah	**3**	di-di-di-dah-dah
L	di-dah-di-dit	**4**	di-di-di-di-dah
M	dah-dah	**5**	di-di-di-di-dit
N	dah-dit	**6**	dah-di-di-di-dit
O	dah-dah-dah	**7**	dah-dah-di-di-dit
P	di-dah-dah-dit	**8**	dah-dah-dah-di-dit
Q	dah-dah-di-dah	**9**	dah-dah-dah-dah-dit
R	di-dah-dit	**0**	dah-dah-dah-dah-dah

An abbreviated form of 0 (zero) is sometimes used and consists of one long dah. Similarly an abbreviated 9 in the form of an N (dah-dit) is in quite frequent use, and can save considerable time during contest work where many numbers are being exchanged.

Punctuation

Frequently employed in Amateur Radio

Question Mark	di-di-dah-dah-di-dit
Full Stop	di-dah-di-dah-di-dah
Comma *	dah-dah-di-di-dah-dah

*Often used to indicate exclamation mark.

Procedure Signals

Stroke	dah-di-di-dah-dit
Break sign (=)	dah-di-di-di-dah
End of Message (+ or $\overline{\text{AR}}$)	di-dah-di-dah-dit
End of Work ($\overline{\text{SK}}$)	di-di-di-dah-di-dah
Wait ($\overline{\text{AS}}$)	di-dah-di-di-dit
Preliminary call ($\overline{\text{CT}}$)	dah-di-dah-di-dah
Error	di-di-di-di-di-di-di-dit
Invitation to transmit (K)	dah-di-dah
KN	dah-di-dah-dah-dit

* * *

One dah should be equal to three di's (dit's).
The space between parts of the same letter should be equal to one di (dit).
The space between two letters should be equal to three di's (dit's).
The space between two words should be equal to from five to seven di's (dit's).

On some bands it will be found that the listening frequencies are widely separated from the calling frequencies. On Top Band (160m), for example, it is current practice for North American stations to operate near the low end of the band and listen around 1825 kc/s, and for European amateurs to operate around 1825 kc/s and listen at the bottom end of the band. There is a special set of Q codes for use under these circumstances:

QLM I will listen from the low-frequency end of the band up to the middle.
QML ..... middle down to the low-frequency end.
QHM ..... high-frequency end down to the middle.
QMH ..... middle up to the high-frequency end.

In normal circumstances a station should not be called until he has completed a contact, as this is a bad practice and labels the caller as a bad operator. It also encourages further misbehaviour by other stations who may be standing by, and often causes confusion and interference. The only time when a short call may be made during a contact is when a DX station is allowing the practice known as " tail ending " to take place. The technique here is to drop in one's call almost zero beat with the station working the DX just as he is signing on his final transmission. Unfortunately this practice can very easily lead to abuse, and frequently produces chaotic results.

Single Sideband

This mode of transmission is gradually replacing a.m. and enables users to carry on telephone type conversations where both stations are making full use of voice controlled transmission (vox). Any interference on the frequency may be detected immediately, and questions answered as they arise in the course of conversation. It is also very easy for a third person to join in the contact at any time. It should be remembered however before trying to break in on a contact that this is in effect a private conversation between two individuals, and that therefore common courtesy should be exercised over the manner of joining the contact.

Regulations require that during prolonged contacts, station identification should be made at intervals not exceeding 15 minutes (every 10 minutes in USA). When identifying on telephony it is suggested that the phonetic alphabet shown in Table 20.2 is used. It must be remembered that phonetics used must not be facetious or misleading.

Helpfulness

A good operator will always be helpful, not only to the newcomer but to any amateur who is in need of co-operation in order to carry out tests on his transmitter or aerials in order to bring them up to the desired performance. In the same way it would be a friendly gesture to inform any operator, whose signals are defective, that his transmission is not as it should be (and possibly violating the licence conditions) and at the same time to offer reports on any adjustments that may appear necessary to cure the faulty transmission.

A good operator will also always give an honest and useful report to the stations he works. He should endeavour to give the sort of reports that he himself would appreciate,

TABLE 20.2

Phonetic Alphabet

A	Alfa	**N**	November
B	Bravo	**O**	Oscar
C	Charlie	**P**	Papa
D	Delta	**Q**	Quebec
E	Echo	**R**	Romeo
F	Foxtrot	**S**	Sierra
G	Golf	**T**	Tango
H	Hotel	**U**	Uniform
I	India	**V**	Victor
J	Juliet	**W**	Whiskey
K	Kilo	**X**	X-ray
L	Lima	**Y**	Yankee
M	Mike	**Z**	Zulu

remembering that signal strength is not the sole criterion of a good transmission and that the bandwidth occupied by a telephony signal or the keying waveshape of a morse signal is even more important than its strength.

Codes and Abbreviations

A great deal of information can be transmitted in a short time on c.w. if full use is made of the standard abbreviations and codes. They may also help at times during contacts with amateurs who are experiencing difficulty on telephony due to language differences. However, they should generally not be used on s.s.b. or a.m. Many listeners with no special knowledge of Amateur Radio have receivers which tune the amateur bands, and are bound to be unfavourably impressed by individuals using apparently senseless expressions. In these days of pressure on frequency allocations it is very important that the Amateur Service should be looked upon as responsible serious experimenters by casual listeners, and anything which may give the impression of irresponsible transmission behaviour avoided.

Originally the only code used for reporting signals was the one used for loudness and was graded in the arbitrary scale R1-R9. Later, when interference became a problem it was necessary to describe the readability of the signal, since under heavy interference even a strong signal might be only partly readable. The scale QSA1 to QSA5 was introduced for this purpose, and at the same time the signal strength was denoted by the scale QRK1 to QRK9. Later, when more attention was being paid to the tone of c.w. signals, another code was evolved. This is the now universally used RST code Table 20.3, which indicates *Readability* (R1 to R5), *Signal Strength* (S1 to S9), and *Tone* (T1 to T9). On phone it is the custom to give the report as *Readability* (1-5) and *Strength* (1-9). The SINPO code, Table 20.4, may also be used.

There are five signals which are used at the end of transmissions under different sets of circumstances. It is very important to employ and understand the correct use of these letters in order to reduce misunderstandings and confusion. The letters $\overline{\text{AR}}$ are sent at the end of a call to a specific station, before contact has been established; for example "G3XXX G3XXX G3XXX de G2ZZZ G2ZZZ G2ZZZ $\overline{\text{AR}}$." The letter K is sent at the end of a CQ call, or at the end of a transmission during an established contact when there is no objection to other callers joining the contact. When a call is made to a specific station, and replies are not desired from any other, or at the end of a transmission during a contact when calls from other stations are not wanted it is customary to end with the letters $\overline{\text{KN}}$; for example, "G3XXX de G2ZZZ $\overline{\text{KN}}$." At the end of the last transmission of a contact the letters $\overline{\text{SK}}$ are sent, immediately before the call-signs, e.g. ". . . $\overline{\text{SK}}$ G3XXX de G2ZZZ." This means that G2ZZZ is finished and is ready to receive other calls. If the station is closing down and does therefore not wish to receive further calls the letters CL should be sent after the call-signs, e.g. ". . . $\overline{\text{SK}}$ G3XXX de G2ZZZ CL." If a station desires another contact after signing with another station he should not use the letters QRZ? unless he has already heard someone calling him, but should call CQ, always remembering that it is courteous to move away from the frequency and operate elsewhere if the station just contacted was originally operating there.

TABLE 20.3

The RST Code

Readability

R1	Unreadable.
R2	Barely readable, occasional words distinguishable.
R3	Readable with considerable difficulty.
R4	Readable with practically no difficulty.
R5	Perfectly readable.

Strength

S1	Faint signals, barely perceptible.
S2	Very weak signals.
S3	Weak signals.
S4	Fair signals.
S5	Fairly good signals.
S6	Good signals.
S7	Moderately strong signals.
S8	Strong signals.
S9	Extremely strong signals.

Tone

T1	Extremely rough hissing note.
T2	Very rough a.c. note, no trace of musicality.
T3	Rough, low-pitched a.c. note, slightly musical.
T4	Rather rough a.c. note, moderately musical.
T5	Musically modulated note.
T6	Modulated note, slight trace of whistle.
T7	Near d.c. note, smooth ripple.
T8	Good d.c. note, just a trace of ripple.
T9	Purest d.c. note.

An 'X' is added after the appropriate T number if the transmission appears to be crystal controlled. The letter 'C' after the T number indicates the presence of chirp. The letter 'K' indicates key clicks.

TABLE 20.4

The SINPO Code

S	Signal Strength	**1**	Barely audible
		2	Poor
		3	Fair
		4	Good
		5	Excellent
I	Interference (QRM)	**1**	Extreme
		2	Severe
		3	Moderate
		4	Slight
		5	Nil
N	Noise (QRN)	**1**	Extreme
		2	Severe
		3	Moderate
		4	Slight
		5	Nil
P	Fading	**1**	Extreme
		2	Severe
		3	Moderate
		4	Slight
		5	Nil
O	Overall Rating	**1**	Unusable
		2	Poor
		3	Fair
		4	Good
		5	Excellent

TABLE 20.5

The Q Code

QAV	I am calling . . .
QCM	There seems to be a defect in your transmission.
QIF	. . . (station) is using . . . (frequency).
QRA	The name of my station is . . .
QRB	The distance between our stations is . . .
QRG	Your exact frequency in kc/s is . . .
QRH	Your frequency varies.
QRI	Your note varies.
QRJ	Your signals are very weak.
QRK	The readability of your signals is . . . (1 to 5).
QRL	I am busy.
QRM	There is interference.
QRN	I am being troubled by atmospheric noise.
QRO	Increase power.
QRP	Reduce power.
QRQ	Send faster (. . . . words per minute).
QRS	Send more slowly (. . . . words per minute)
QRT	Stop sending.
QRU	I have nothing for you.
QRV	I am ready.
QRW	Please tell . . . . . that I am calling him.
QRX	I will call you again.
QRZ	You are being called by . . . . .
QSK	I can hear between my signals (i.e., I am using break-in).
QSL	I give you acknowledgment of receipt.
QSM	Repeat the last message.
QSP	I will relay to . . . . .
QSV	Please send a series of V's.
QSW	I will transmit on . . . . kc/s.
QSY	Change to . . . . kc/s.
QSZ	Send each word or group twice.
QTH	My location is . . . .
QTR	The exact time is . . . .

The Z Code

ZAN	I am receiving absolutely nothing.
ZAP	Acknowledge please.
ZCK	Check your keying.
ZCL	Transmit your call letters intelligibly.
ZDF	Your frequency is drifting.
ZDM	Your dots are missing.
ZFO	Your signals have faded out.
ZGS	Your signals are getting stronger.
ZGW	Your signals are getting weaker.
ZOK	I am receiving O.K.
ZRN	You have a rough note.
ZSU	Your signals are unreadable.
ZWO	Send words once.
ZWT	Send words twice.

The most widely used code amongst amateurs is the ' Q ' code, Table 20.5. Where the ' Q ' group is followed by a question mark an answer to the question is required, where appropriate the answer should be qualified by the addition of a number according to the following classification: 1.—Very slight; 2.—Slight; 3.—Moderate; 4.—Severe; 5.—Extreme. For example " QRM5 " means " there is extremely severe interference." " QRM ? " means " Is there interference ? " These numbers also apply when the ' Z ' code is used.

The ' Z ' code, Table 20.5, is mainly used by commercial operators, but contains a number of useful phrases for amateur work not included in the ' Q ' code.

Log Keeping

The Post Office requires all amateurs to keep a log book containing full details of all transmissions. This must be a bound book, and all entries made in an indelible manner. The details to be recorded consist of the date, the time of commencement of every call (including test calls), the call-signs of all stations from which messages addressed to the station are received or to which messages are sent, times of commencing and ending communication with each such station (in GMT), the frequency (not frequency band) of transmission, the class of emission, and the time of closing down the station. When the station is operated from temporary or alternative premises, the address should be entered in the log. Entries must be made at the time of the operation, and no gaps should be left in the log. It should be noted that if the station is operated by anyone other than the licensee, the person doing so should sign the log entry with his full name and call-sign.

Suitable log books complying with all the necessary requirements are readily available, and may be purchased from RSGB Headquarters.

Power

The Post Office transmitting licence stipulates the maximum power input permissible on the various bands, and the maximum power output permissible in the case of single sideband transmitters. The maximum power that should be used is the lowest that will ensure satisfactory reception by the other station, and except where the transmitter is designed only for low power it is desirable to have some easy means of reducing power whenever conditions permit. Cross town contacts may be carried on with very low power indeed, and a great deal of interference with distant stations thereby avoided. Local contacts should be arranged wherever possible to take place on a band which is not in use for DX contacts.

STATION LAYOUT

Before commencing to build equipment it is desirable to have some idea of the ultimate layout of the completed station, for the physical size of the various units will, to some extent, depend on whether a table-top, bureau bookcase, cupboard, rack and panel, or console assembly will be used. This in turn will depend on the amount of space that can be devoted to the station. It is relatively easy to construct gear, even of high power capabilities, which occupies very little space, by employing miniaturized components. A complete transmitter–receiver may in fact be contained in one small box.

Choosing a Site

Amateur stations have been set up in many different places in and around the home, the location obviously depending on domestic circumstances, the accommodation available, and the operator's ambitions. Ideally the best arrangement is for an entire room to be set apart for the station, a small bedroom being an ideal choice. Not only does such accommodation provide the maximum comfort and quietness for operating, but also affords complete safety from danger to other members of the family. However there are other quite suitable places in most homes.

Very efficient installations have been set up in cupboards

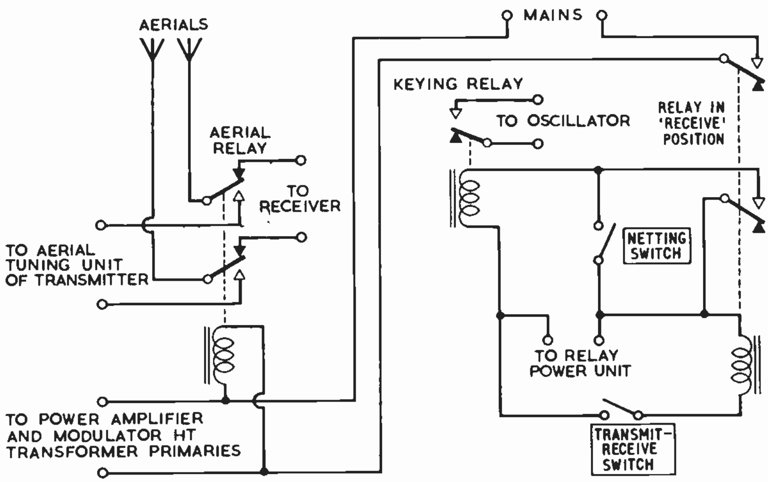


Fig. 20.1. A method of obtaining single switch control of a station. All relays are shown in the normal "receive" position. The driver stages may also be switched on and off with the power-amplifier and modulator h.t. supplies by using an additional pair of contacts on the receiver-controlled relay. Provided that suitable relays are chosen, the keying and aerial change-over relays may be energized from the same power supply, but where open-wire feeders are switched a mains-energized relay will probably be found more suitable.

under staircases, built into bureaux or cupboards in downstairs rooms or in sheds in gardens. All these places suffer from some drawback that does not exist for the lucky person who can devote a whole room to his station. The site under the staircase will inevitably be small, dark and difficult to ventilate, and the station in the downstairs room shared with the other members of the family will often be noisy, while the shed in the garden may be too cold and damp in winter and too hot in summer, besides being less accessible. The choice must be made by each individual, preferably in consultation with the rest of the family, after weighing up the pros and cons of each possible alternative. If at all possible, a part of the house which is nearest to the aerials should be chosen, at the same time remembering that aerials should be sited as far away from television aerials as possible. Another point to be borne in mind is the accessibility of power points, and earthing for the equipment. The aim should be to arrange the equipment in such a way that operating, even over long periods, is a pleasure so that maximum efficiency is obtained.

Station Wiring

The care that has gone into the construction of the individual items of equipment should be continued when linking them up. Connecting cables should be short and concealed wherever possible. It is a good idea to use connecting leads of different colours, since not only does this facilitate rapid servicing but it is also a useful safety precaution. Another point that should be watched where several plugs and sockets are used for carrying different voltages on one chassis is never to use identical components, because it is easy to make a mistake and put the wrong plug into the socket. Here again colour coding of the leads is helpful. In these days of imported equipment, which is designed for 110 volt mains, it is a good plan to feed the mains from the Variac or auto-transformer to a distribution board with three pin sockets of a type different from the 230 volt sockets, so that the expensive mistake of feeding 230 volts into gear designed for 110 volts cannot happen.

Circuit diagrams of all the circuit equipment should always be kept readily available, as it is an obvious aid to rapid and efficient servicing. Diagrams of all the interconnecting wiring will also be found invaluable.

Insurance

While every amateur will strive to construct and maintain his station so that it is completely safe for himself and any others who may visit it, there is always the possibility that an accident may occur. Owing to component failure a visitor may receive an electric shock, or an aerial or mast may fall and injure someone or damage property. Such an occurrence can result in a legal action, and the awarding of considerable damages against the person held responsible for the accident. This risk can be insured against, either by an extension to the existing Householder's Comprehensive policy or by taking out a separate public liability policy if only a fire insurance is held. The annual premium will be only a few shillings. Readers are strongly recommended to consult their insurance company or brokers.

Switching Systems

The minimum number of switches should be used to change the equipment from the *send* to the *receive* position, and this is best achieved by the use of relays so that at least single-switch operation is possible. Most communications receivers have a send-receive switch, and there are often terminals at the rear of the chassis, which have a d.c. voltage across them in the *receive* position, suitable for operating a relay. With the switch in the *receive* position and the relay wired as shown in **Fig. 20.1**, there will be a voltage on the relay; this will open-circuit the high-tension leads to the transmitter and modulator and to the aerial change-over relay (if this is wired so that in the *rest* position the aerial is connected to the receiver). When the switch is in the *send* position the relay closes the circuits switching on the transmitter and modulator and energizing the aerial relay so that the aerial is connected to the transmitter. A relay suitable for carrying out these functions will draw only a small current. Even if terminals are not provided on the receiver it is a very simple matter to provide a suitable voltage.

If the position where the transmitter is keyed is such that a high voltage is likely to be produced across the terminals of the key a relay should be used. In any case a keying relay is desirable where the leads from the keyed circuit to the key itself are of any appreciable length.

Break-in keying and voice-operated transmitter control

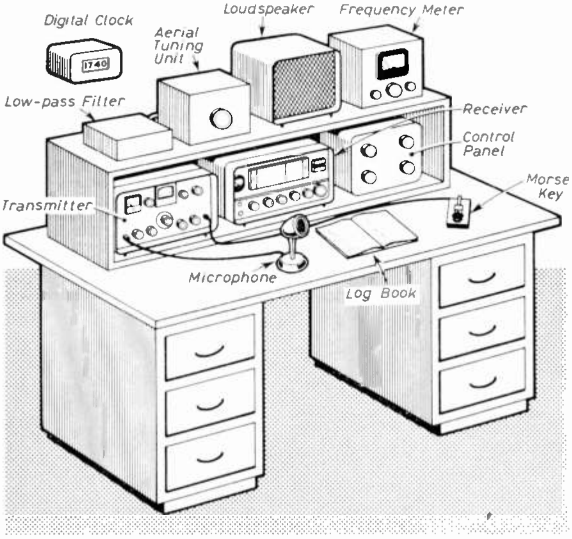


Fig. 20.2. Typical table-top arrangement of an amateur station. A suitable operating table may be made from a standard flush door polished and supported on two cupboards or two two-drawer filing cabinets.

are additional refinements (See Chapter 8—*Keying and Break-in* and Chapter 10—*Single Sideband Techniques*).

Arranging the Equipment

A typical arrangement of the basic equipment that goes to make up the average station is shown in **Fig. 20.2**, but there are, of course, many other ways of assembling the various items. It should be remembered that the focal point of the station will be the operator's chair, which should be as comfortable as possible. From the operating position, all the main controls should be within easy reach, and all meters clearly in view.

The transmitter should be sited so that the aerial feeders can be connected without too many bends having to be made in them, and so that in the case of tuned feeders they can be spaced well away from other objects. If it is possible to arrange for feeders to pass through a small window, the glass in which has been removed and replaced with Perspex, so much the better.

The Morse key is normally placed on the right side of the desk or table, but some operators have other preferences. Whatever position is chosen it should be the one which affords the greatest comfort during periods of sustained operation.

For phone operation the microphone can be located in any convenient position on the desk. A good flat area, with sufficient space to accommodate the key, microphone, log, scribbling pad, and a large ashtray is essential for comfortable operating.

SAFETY PRECAUTIONS

Safety is of paramount importance. Every precaution should be taken to ensure that the equipment is perfectly safe, not only for the operator himself but also for the other members of the household, or visitors. Double pole switches should be used for all mains circuits, and interconnected switches should be fitted so that no part of the equipment can have high voltage applied to it until the valve heaters and low power stages have first been switched on. This precaution may not only save the life of the operator, but it also protects the transmitter against damage.

The whole station should be controlled by one master switch, located in a prominent position, and all members of the household should know that in the event of an emergency it must be switched OFF *before* anything is touched.

All aerials should be safeguarded against lightning, either by manual switching to a good earth when the station is not being used, or by the use of lightning arresters. The construction of a suitable spark-gap arrester is shown in **Fig. 20.3.** Arresters for use in co-axial cable are also available. Great care should be exercised before touching feeders which have been disconnected during a thunderstorm. Further information on lightning protection is given in Chapter 13—*H.F. Aerials*.

Another often overlooked precaution concerns the rating of the mains supply to the station. A 150 watt amateur station fully equipped with ancillary apparatus can draw quite a heavy current from the mains, and before assembling or operating a station it is essential to calculate the current that will be drawn, and to make certain that the existing house wiring will carry this current without being overloaded. If any doubt exists new wiring should be installed.

It is very important that every amateur should develop a strict code of safety discipline for use when handling his equipment. It should be a rule never to work on a piece of gear which is plugged into the mains, even if all switches are turned off. Always be certain that filter capacitors in power packs are discharged to earth before putting a hand near them, as the larger ones are capable of retaining a considerable charge for quite a long time. On those few occasions when it is impossible to carry out necessary adjustments with the equipment turned off the operator should keep one hand behind his back or in his pocket, should never be wearing headphones, and should make certain that no part of his body is touching an object which is earthed. Any such adjustments should be carried out with well insulated tools.

The vast mast majority of shocks sustained from electrical equipment are derived from the 240 volt mains line lead. Every year there are 100 or more fatalities in the UK alone mostly from accidental contact with mains voltage. There is

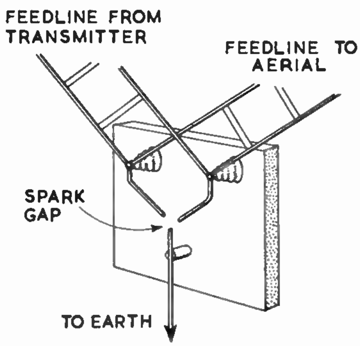


Fig. 20.3. Lightning arrester. This arrangement is suitable for all types of twin-feeder systems. The mounting board should be of fireproof insulating material provided that the spacing between the feeder and earth gaps is considerably smaller than the distance between the feeder gaps and the metal plate.

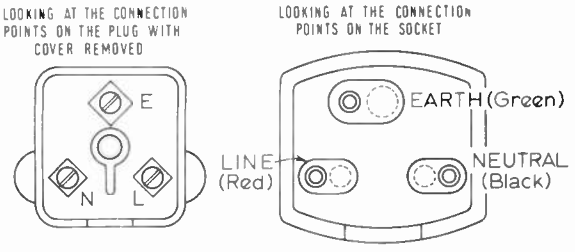


Fig. 20.4. The correct wiring for three-pin plugs and sockets. To test that a socket is correctly wired, a lamp should light if connected between " L " and " N " or " L " and " E " but not when connected between " N " and " E." A neon bulb will glow when touched against " L."

evidence to suggest that because of the different physiological effects, those who find themselves across voltages of more than 1000 volts have a better chance of survival than those subjected to severe medium voltage shocks. Voltages as low as 32 volts have been known to kill. As the jingle says " It's volts that jolts, but mils that kills."

The real danger from mains supplies is where either the skin resistance is lowered by dampness or perspiration, or where the victim grips an extensive area of " live " metal whilst in good contact with earth. It is against this second possibility that particular care is needed in amateur stations.

One hazard is equipment with a mains connected chassis being used under conditions for which it was not intended. All modern British television sets and a fair proportion of domestic broadcast receivers fall into this category; not only all the " a.c./d.c. " sets but also—and this is not always appreciated—a large proportion of " a.c. only " models. These sets are built to comply with the British Standards safety specification (B.S.415) which, among other precautions, lays down that it should be impossible for a little finger to touch any part of the chassis, and also insists on double pole on-off switches. Old radio sets, particularly pre-war models, often do not comply with B.S. 415, or may have damaged backs, or be used with exposed control spindles or grub screws. If there is any such equipment in *your* station, the exposed parts should be checked with a neon to make certain that they are not " live," with the on/off switch in both positions. Remember that a single pole switch in the neutral lead will leave the chassis " live " with the set off, even if the chassis is connected to " neutral " when the set is on. After checking, a *non-reversible* plug should be fitted. It is well worth while to ensure that all three pin mains sockets in the house are wired correctly: unfortunately many sockets in our homes *are* incorrectly connected. A three pin socket with the " earth " socket at the top should have the " neutral " socket on the bottom left, and the live " line " socket on the bottom right—these directions are when looking into the socket and must be reversed when looking at the back of the socket for wiring (See Fig. 20.4). Correct colour coding of leads in the UK is: '*Live*,' red; '*Neutral*,' black; '*Earth*,' green. It is very important to note that this code may not apply to imported equipment, and the manufacturer's instructions should be very carefully studied before plugging into the mains. The use of modern 13 amp fused plugs is recommended.

An even greater hazard, because it is seldom anticipated, can arise under fault conditions on equipment fitted with a double wound (i.e. " isolating ") transformer of the type used in the vast majority of amateur receivers and transmitters. It is by no means unknown for the primary winding to short circuit to the screening plate between the primary and secondary, the core, or to one of the secondary windings, so that the chassis of the equipment becomes " live." Such equipment will often continue to operate quite normally and can thus represent a real danger over a considerable period. The best safeguard against this is to ensure that the shield between the primary and the other windings, the core, and the chassis are all effectively earthed. The earth connection must be of sufficiently low resistance for the supply fuses to blow. These fuses should always be of the minimum practicable amperage. It is no use having a 50 ohm resistance to earth and a 10 amp fuse—if this is the case the size of the electricity bills may be surprising, but the hazard is likely to remain undetected.

Another hazard is the electric tool which has developed a fault and which has a " live " casing. This can happen, for example, with soldering irons and electric drills. The modern domestic electric drill is designed with this danger very much in mind, but even so it should be remembered that in industry it is recommended that such tools should be used only in "earth free" areas—which is far from being the case in the average amateur installation. A very careful eye should be kept on the leads to all such tools, and any " tingles " felt when they are in use should be investigated immediately.

Many amateurs fit extra power sockets in their stations and the control arrangements may call for quite a lot of semi-permanent a.c. wiring and switching. These should always conform with the high standards laid down in the IEE Wiring Regulations (14th Edition). These are rather formidable reading for the non-professional but a number of books giving sound advice on modern wiring practice, based on the IEE recommendations, have been published and can often be obtained from local libraries. Advice can also be obtained from the offices of local electricity boards.

It is a wise precaution to have a fire extinguisher of the type suitable for use on electrical equipment in the radio room. The best are those which direct a stream of carbon dioxide on to the burning area; the powder and carbon tetrachloride types may also be used but the latter is liable to cause further damage if the contents come into contact with radio equipment.

SAFETY RECOMMENDATIONS FOR THE AMATEUR RADIO STATION

1. All equipment should be controlled by one master switch, the position of which should be well known to others in the house or club.
2. All equipment should be properly connected to a good and permanent earth. (*Note A.*)
3. Wiring should be adequately insulated, especially where voltages greater than 500 V are used. Terminals should be suitably protected.
4. Transformers operating at more than 100 V R.M.S. should be fitted with an earthed screen between the primary and secondary windings.
5. Capacitors of more than 0.01 μF capacity operating in power packs, modulators, etc. (other than for r.f. bypass or coupling) should have a bleeder resistor connected directly across their terminals. The value of the bleeder resistor should be low enough to ensure rapid discharge. A value of 1/C Megohms (where C is in microfarads) is recommended. The use of earthed probe leads for discharging capacitors in case the bleeder resistor is defective is also recommended. (*Note B*). Low leakage capacitors, such as paper and oil filled types, should be stored with their terminals short-circuited to prevent static charging.
6. Indicator lamps should be installed showing that the equipment is live. These should be clearly visible at the operating and test position. Faulty indicator lamps should be replaced immediately. Gas filled (neon) lamps are more reliable than filament types.
7. Double-pole switches should be used for breaking mains circuits on equipment. Fuses of correct rating should be connected to the equipment side of each switch. (*Note C.*) Always switch off before changing a fuse. The use of a.c./d.c. equipment should be avoided.
8. In metal enclosed equipment install primary circuit breakers, such as micro-switches, which operate when the door or lid is opened. Check their operation frequently.
9. Test prods and test lamps should be of the insulated pattern.
10. A rubber mat should be used when the equipment is installed on a floor that is likely to become damp.
11. Switch off before making any adjustments. If adjustments must be made while the equipment is live, use one hand only and keep the other in your pocket. Never attempt two-handed work without switching off first. Use good quality insulated tools for adjustments.
12. Do not wear headphones whilst making internal adjustments on live equipment.
13. Ensure that the metal cases of microphones, Morse keys, etc., are properly connected to the chassis.
14. Do not use meters with metal zero adjusting screws in high voltage circuits. Beware of live shafts projecting through panels particularly when metal grub screws are used in control knobs.
15. Aerials should not, under any circumstances, be connected to the mains or other h.t. source. Where feeders are connected through a capacitor, which may have h.t. on the other side, a low resistance d.c. path to earth should be provided (r.f. choke).

★ ★ ★

Note A.—Owing to the common use of plastic water main and sections of plastic pipe in effecting repairs, it is no longer safe to assume that a mains water pipe is effectively connected to earth. Steps must be taken, therefore, to ensure that the earth connection is of sufficiently low resistance to provide safety in the event of a fault. Checks should be made whenever repairs are made to the mains water system in the building.

Note B.—A " wandering earth lead " or an " insulated earthed probe lead " is an insulated lead permanently connected at one end to the chassis of the equipment; at the other end a suitable length of bare wire is provided for touch contacting the high potential terminals to be discharged.

Note C.—Where necessary, surge-proof fuses can be used.

THE RSGB AND THE RADIO AMATEUR

THE Radio Society of Great Britain (RSGB) is the National Society of radio amateurs in the United Kingdom. More than 7500 of its 14,000 members hold amateur transmitting licences; the others either hope to do so later or are interested primarily in the receiving side of Amateur Radio. More than 1000 members live overseas.

H.R.H. The Prince Philip, Duke of Edinburgh, K.G., is the Society's Patron.

The Society acts as the spokesman for the radio amateur and Amateur Radio in the UK and is one of the Founder Members of the International Amateur Radio Union, the worldwide association of the various National Societies.

The Society was founded as the London Wireless Club more than 50 years ago in 1913 but soon attracted members throughout the country; the name of Radio Society of Great Britain was formally adopted in 1922. For many years its activities have been devoted almost entirely to the many aspects of Amateur Radio—that is, the transmission and reception of short wave and ultra-short wave radio signals as a hobby pursued for the pleasure to be derived from an interest in radio techniques and construction and for the ensuing friendships with like-minded persons throughout the world.

The Society helps Amateur Radio in many ways. Of particular importance is the provision of information on technical matters and on the various activities and events of concern to amateurs. Since 1925 it has published a monthly journal—RADIO COMMUNICATION—the oldest and largest magazine in this country devoted to Amateur Radio. All members receive this magazine by post, without payment other than their annual membership subscriptions.

The Society is administered by a Council elected by the Corporate membership. A full-time staffed Headquarters is maintained in London. There are also 17 elected Regional Representatives who with the help of Area Representatives arrange local meetings and other activities.

The Society is proud of its half-century role in the development of Amateur Radio and of the many eminent scientists who have been connected with it—including Marconi, Lodge and Fleming. It was the organized radio amateurs who originally demonstrated that wavelengths below 100 metres could provide world-wide communication on low power. But the Society looks to the future, rather than the past, and concentrates its efforts on the effective organization of amateur activity in order to provide the greatest opportunities for useful experimental work, and to encourage general interest in and enjoyment of the hobby of Amateur Radio.

Why You Should Join

The following are just a few of the many reasons why, if you are really interested in Amateur Radio, you should join the RSGB immediately:

You will receive every month a copy of RADIO COMMUNICATION, recognized as providing a complete and accurate survey of every phase of Amateur Radio activity. Containing at least 64 pages each month, RADIO COMMUNICATION is noted internationally for the high standard of its technical and constructional articles, written by Britain's leading radio amateurs, and for the wide scope of its news coverage. The needs of newcomers are not overlooked in the selection of technical articles.

You will be able to use the world's largest and most comprehensive free QSL Bureau, operated by the Society. Use of this efficient Bureau will save the active amateur and listener a great deal of trouble. QSL cards are sent to and received from the Bureau in batches. This eliminates the need for stamping, addressing and posting of individual cards. The Bureau distributes the cards via other National Societies' bureaux to amateurs throughout the world. Full details of how the QSL Bureau operates are sent to every member on election.

You will receive a certificate of membership and a lapel badge which identifies you as part of the Amateur Radio movement. Members who do not hold amateur transmitting licences are given special identification numbers for use in connection with the QSL Bureau, beginning BRS (British Receiving Station), ORS (Overseas Receiving Station) or A (Associate), followed by a number.

You will be encouraged to contribute, according to your interests, to the advancement of Amateur Radio. Many members serve as local representatives, or on local or national committees, or pass on to other members, through RADIO COMMUNICATION or by lectures, the results of their experiments and observations. The members in fact *are* the Society.

The Society is recognized as the representative of the Amateur Radio movement in all negotiations with the Post Office on matters affecting the issue of amateur transmitting licences.

The Society helps organize many special scientific studies and tests, and has set up a Radio Amateur Emergency Network in collaboration with the British Red Cross Society, the St. John Ambulance Brigade and the Police.

The Society maintains special committees dealing with such subjects as television interference, mobile operation, v.h.f., technical matters, licences and the annual RSGB exhibition.

The Society has been established for over half-a-century. It has a long record of sound and efficient administration, through an elected Council, aided by a full-time Secretariat.

If you are over 21 years of age or hold an amateur transmitting licence you are eligible to become a Corporate member of the Society. You do not have to be engaged professionally in radio—but equally this would not debar you from joining. Many members do in fact work in the electronics field, but for very many others radio is purely a spare-time hobby. Those under 21 who do not hold an amateur transmitting licence may become Associates. Associates have many of the privileges of full membership but do not vote in the annual Council election or on matters affecting the management of the Society. Licensed amateurs cannot be Associates.

How to Join the RSGB

Joining the RSGB is simple, but there are of course certain formalities to be observed. As explained earlier, anyone with an active and genuine interest in Amateur Radio is warmly welcome to apply for membership.

All applicants, for both Corporate and Associate membership, should be proposed by a Corporate Member of the Society to whom they are personally known. The member simply completes the proposal on the Application Form, and you will find that he will be glad to do this. Many newcomers to Amateur Radio—who are most welcome as members—may not know or be in touch with other members. In such cases a brief reference in writing should be submitted from a suitable person who can vouch for your interest in Amateur Radio.

In order to apply for Associate membership, you must be under 21 years of age and not be a holder of an amateur transmitting licence. Associates must apply for transfer to Corporate membership on reaching 21 years of age, or if under this age, immediately they obtain a transmitting licence. If you wish to apply for Associate membership you should ask for the appropriate Application Form. All applications are placed before the Council at its monthly meetings, generally held during the second week of the month.

The current annual subscription rates are: Corporate Members £2 10s. 0d.; Associates £1 5s. 0d. The first year's subscription should be sent with the Application Form. There is no entry fee. After having been a Corporate Member for five consecutive years, a member may, subject to the approval of Council, commute all future annual subscriptions by a payment of £45.

All correspondence should be sent to the Radio Society of Great Britain, 35 Doughty Street, London, W.C.1.

WHAT THE SOCIETY DOES

Some of the important activities of the Society have already been described, but there are many other ways in which the Society helps radio amateurs and all those interested in Amateur Radio. A few of these are outlined below:

Publications

The main publishing activity of the Society is the monthly Journal, RADIO COMMUNICATION. However, the Society also produces many books and other publications to help the amateur. Because the Society is anxious to disseminate sound technical information as widely as possible, many are issued at prices well below what a commercial publishing organization would have to charge. A notable example is this Handbook.

Some publications of the Society are specially written for newcomers to help them obtain transmitting licences; these include *A Guide to Amateur Radio*, *The Radio Amateur's Examination Manual* and *The Morse Code for Radio Amateurs*.

The Society also provides facilities for obtaining a selection of the many Amateur Radio publications issued in the United States where there are over a quarter-million radio amateurs. The most popular American publications are generally available immediately from Headquarters.

Meetings

Official Society meetings are held throughout the British Isles. Local meetings are held by RSGB Groups and by affiliated radio societies. Mobile rallies and specialized conventions are also held regularly. A full list of Forthcoming Events appears monthly in RADIO COMMUNICATION.

Frequencies

The Society maintains close liaison with the GPO on all matters affecting licence facilities and the frequencies assigned to Amateur Radio, and sends official representatives to the important World Radio Conferences of the International Telecommunication Union and other conferences where decisions vital to the future of Amateur Radio are taken. The RSGB is a member of the International Amateur Radio Union, the world-wide organization of national amateur radio societies.

Contests and Field Days

Many interesting Tests, Contests and Field Days—some of which are open to listeners as well as transmitting members —are held each year. Trophies or Certificates are awarded to leading entrants.

Achievement Certificates

A number of Certificates representing graded degrees of achievement in Amateur Radio operating (receiving as well as transmitting) are issued by the Society. These include the DX Listeners' Century Award, the Worked the British Commonwealth Award, Four Metres and Down Certificates and the Commonwealth DX Certificate. The Rules governing all awards are available from RSGB Headquarters.

News Bulletin Service

Every Sunday morning special news bulletins for radio amateurs are transmitted, under the call-sign GB2RS, from stations throughout the British Isles in the 3·6 and 144 Mc/s bands.

Slow Morse Transmissions

The Society sponsors the transmission by amateurs throughout the country of Morse practice lessons intended for beginners. Details appear periodically in RADIO COMMUNICATION.

Beacon Stations

By arrangement with the BBC the Society operates a beacon station on 144·5 Mc/s under the call-sign GB3VHF from Wrotham, Kent.

GB3ANG at Craigowl Hill, Dundee, is on 145·95 Mc/s.

By arrangement with the ITA, GB3GI operates on 145·990 Mc/s from Strabane, Northern Ireland. Details of other beacon stations appear in RADIO COMMUNICATION.

* * *

The Society supports and encourages all activities "For the Advancment of Amateur Radio." It welcomes within its ranks all those who share this view.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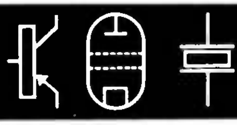

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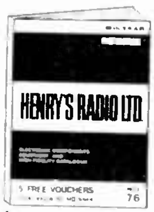

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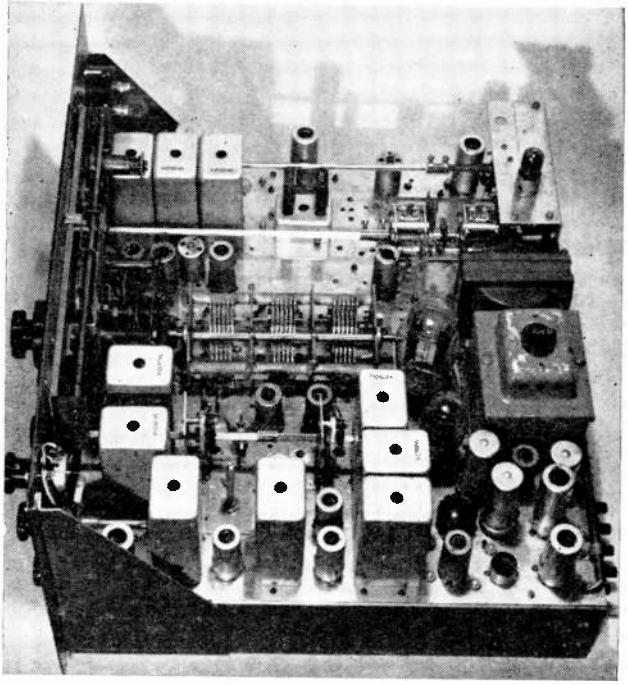

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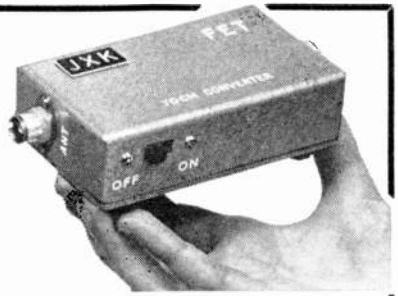

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