RADIO

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VOLUME II

THE ENGLISH UNIVERSITIES PRESS LTD

102 NEWGATE STREET

LONDON, E.C.1

First printed 1954
Reprinted 1957

Printed in Great Britain for the English Universities Press, Limited
by Richard Clay and Company, Ltd., Bungay, Suffolk

GENERAL TECHNICAL SERIES

RADIO

VOLUME II

GENERAL TECHNICAL SERIES

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INTRODUCTION

THIS second volume has been written to cover the syllabus of RADIO II of the City and Guilds of London Institute. The work is of a similar standard to that of the National Certificate (S.3).

A specimen answer has been given at the end of most of the chapters to illustrate more fully some of the subjects discussed therein. Acknowledgment is gratefully made to the examining bodies of the University of London, the City and Guilds of London Institute, the Institution of Electrical Engineers, the British Institution of Radio Engineers and E. T. A. Rapson, M.I.E.E., for their kind permission to publish their questions, and to the Training Branch of the Post Office Engineering Department for many of the specimen answers.

The authors wish to express their thanks to their colleague H. C. Chandler, B.Sc.(Eng.), A.M.I.E.E., A.C.G.I. who has given valuable technical advice, to D. E. L. Shorter, B.Sc.(Eng.), A.M.I.E.E. of the B.B.C. Engineering Division who has given much assistance in connexion with microphones, and to A. J. Robinson, B.Sc.(Eng.), A.M.I.E.E. of the G.P.O. External Telecommunications Executive for assistance with Chapter 12.

Acknowledgment is also due to: Wayne Kerr Laboratories for information on radio-frequency bridges; The British Broadcasting Corporation; J. H. Mitchell, B.Sc., Ph.D. and the British Institution of Radio Engineers; Mullard Limited; Philips Electrical Limited; R.C.A. Photophone Limited; Standard Telephones and Cables Limited (Radio Division); Stratton & Co., Ltd., F. W. O. Bauch and Iliffe & Sons Limited, for supplying illustrations and blocks, and to Mr. K. A. White who has again prepared the diagrams.

JOHN D. TUCKER
DONALD F. WILKINSON.

CONTENTS

SUMMARY OF SYMBOLS

A	ampere (amp.)—unit of current
A	area
c	cycle
C	capacitance
E	e.m.f. (volts)
f	frequency
F	farad—unit of capacitance
g	mutual conductance
h	height of aerial
H	henry—unit of inductance
I	current
k	amount of coupling
l, *L*	length or distance
L	self-inductance
m	metres—unit of length
m	magnification (of amplifier)
M	mutual inductance
P	power
Q	" goodness " of tuned circuit
Q	quantity of electricity
r, *R*	resistance
s	second—unit of time
t	time
v	velocity
V	potential difference (volts)
W	energy or work done
W	watt—unit of power
X	reactance
Z	impedance

Greek Symbols

η (eta)	efficiency
λ (lambda)	wavelength
ω (omega)	$2\pi \times$ frequency
π (pi)	ratio of circumference to diameter of circle
μ (mu)	amplification factor (of valves)
ρ (rho)	resistivity
Σ (sigma)	algebraic sum of

Prefixes

k	kilo (1,000)
m	milli $\left(\frac{1}{1,000}\right)$
M	meg(a) (1,000,000)
μ	micro $\left(\frac{1}{1,000,000}\right)$
$\mu\mu$ or p	micro-micro or pico $\left(\frac{1}{1,000,000,000,000}\right)$

CHAPTER

RADIO-FREQUENCY COILS AND CAPACITORS

THE design of coils and capacitors is a very important branch of radio engineering, for the efficiency of a transmitter and the characteristics of a receiver depend on the qualities of these components.

1.1. Coils

The coils employed in radio receivers differ greatly from those used in transmitters because of the difference in the power levels of the two units. Radio-frequency coils in receivers are rarely subjected to a p.d. of more than a few volts or required to carry a current of more than a few milliamperes. Inductors used in high-power transmitters, however, are required to withstand potentials of thousands of volts and currents of many amperes.

In a receiver resonant circuits containing capacitance and inductance are used to obtain high amplification at and near to the resonant frequency and low amplification at other frequencies. The term used to denote the ability of a receiver to amplify wanted signals at the expense of signals of other frequencies is "selectivity", which depends on the "goodness" or Q, of the tuned circuits, and has been discussed in Volume I.

The Q of an inductor is defined as the ratio of its reactance to its effective series resistance $\left(\frac{\omega L}{r}\right)$. A high Q coil is thus one whose magnetic field is confined within its windings, which themselves have a low value of resistance.

1.2. Inductance of Coils

The inductance of a coil depends on its physical size, shape and the number of turns in it, and increases with size and number of turns. It is thus possible to obtain the same inductance with widely different arrangements, but to obtain the highest Q, that shape which contains the shortest length of wire, and so has the lowest resistance, must be used. This occurs for a single-layer coil when the ratio of radius/length is 1·5.

In order to produce a coil having a high value of inductance, a large number of turns must be used, and to keep the dimensions reasonable the turns must be wound in layers on top of each other. In this case the ratio of radial depth of the winding to its mean radius becomes a factor determining the inductance. See Chapter 4 of Volume I for a description of typical radio-frequency coils for receivers.

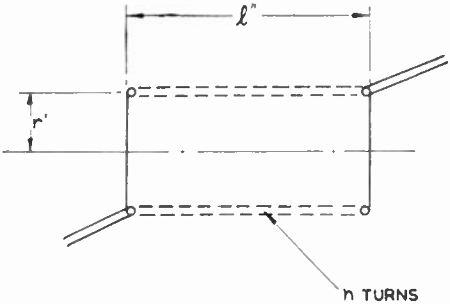


FIG. I.1.—Single-layer Coil

The method of design depends on the type of coil and the accuracy required. For instance, small inductors for receivers need not be calculated very accurately, especially as they are usually screened and so have to be finally adjusted on test.

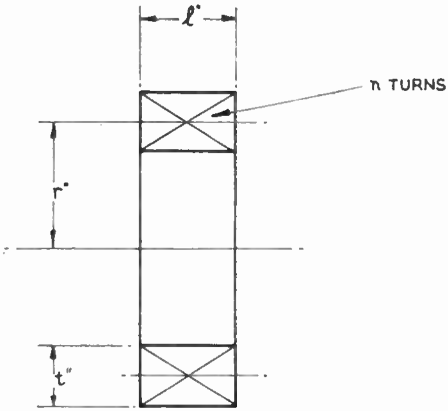


FIG. I.2.—Multi-layer Coil

A common formula for calculating inductance is

$$L = \frac{r^2 n^2}{9r + 10l} \text{ (due to H. A. Wheeler)}$$

where L is given in μH, r is radius in inches, l is length in inches, n is the number of turns (Fig. I.1), and this formula is accurate to about 1 per cent. if $l < 0{\cdot}8\ r$.

For multi-layer coils a useful formula is

$$L = \frac{0{\cdot}8\ r^2n^2}{6r + 9l + 10t} \text{ (also due to H. A. Wheeler)}$$

where L is given in μH, r is mean radius in inches, l is length in inches, t is radial thickness in inches, n is the number of turns (Fig. I.2).

1.3. Effective Resistance of Coils

When a high-frequency alternating current flows in a conductor the current is not distributed uniformly over the cross-section but concentrates near the surface. The centre of the conductor carries proportionally less current as the frequency increases. This phenomenon, known as " skin effect ", causes a higher loss in the conductor with high-frequency current flowing than when a direct or low-frequency alternating current of the same value flows. Thus the effective resistance of a conductor is greater for high-frequency alternating current than for direct current.

There are two methods of overcoming " skin effect ". One is to use a wire of large diameter so that the current-carrying area near the surface is increased, and the other is to divide the wire into a number of strands, all insulated from each other. A form of multi-strand wire known as " Litzendraht " is so formed that all conductors occupy all parts of the cross-section in a given length to ensure equal currents in all strands, but is very expensive to manufacture and is seldom used. " Litz-type " wire, having insulated strands but not so carefully woven, is commonly used up to frequencies of about 2–3 Mc/s, above which solid wire is more satisfactory. Care must be taken when using stranded wire to make connexion to every strand, so that careful cleaning and soldering of the terminations is essential to obtain the advantages of this type of wire.

It is simple to determine the d.c. resistance of a coil of known dimensions and wire size, but " skin effect " raises the effective value many times, depending on the working frequency. An estimate of effective resistance is usually made with charts, examples of these being based mainly on original work by Butterworth. These details are beyond the scope of this work.

When an alternating current flows in an inductor an alternating magnetic field is set up. This field is appreciable, except in the case of toroids, within a distance equal to about one diameter of the coil, and any conductors in this space will have e.m.f.s induced in them. These e.m.f.s cause currents to flow in the conductors, representing a power loss which has to be supplied from the circuit of which the inductor is part. Due to this effect the losses within the coil increase, and so its Q value decreases.

1.4. Proximity Effect

Associated with the eddy-current losses is a change in the inductance of the coil caused by the coupling between it and the surrounding conductor. A badly situated coil thus has both inductance and Q value reduced compared with the values otherwise obtainable. This is sometimes known as " Proximity Effect ".

1.5. Self-capacitance

As an inductor comprises turns of wire close to each other, there is a capacitance between turns. The p.d. applied to it is that in

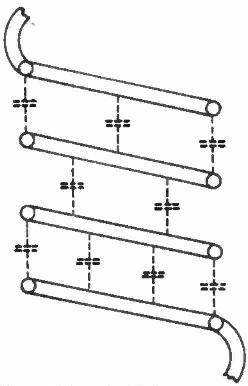


FIG. 1.3.—Self-Capacitance of Single-layer Coil

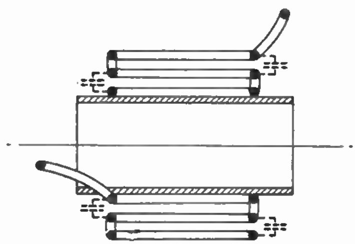


FIG. 1.4.—Self-capacitance of Multi-layer Coil

one turn in a single-layer coil (Fig. 1.3). The total effect in such a coil is a small capacitance, termed the " self-capacitance " of the coil. This self-capacitance resonates with the inductance of the coil at the " self-resonant " frequency, and the inductor then has the dynamic resistance of a parallel-tuned circuit at resonance. Above this frequency the inductor behaves as a capacitance and is no longer useful as a coil. The self-capacitance increases with the amount of wire in a coil, so that high inductances in a single unit are not obtainable for use at high frequencies, for a large-inductance coil will have high capacitance and hence a low self-resonant frequency.

The self-capacitance of a multi-layer coil will be greater than that of a single-layer coil, for the capacitance between turns includes that between turns in different layers (Fig. I.4). For this reason universal or wave winding is employed to separate the turns (see Volume I, Chapter 4).

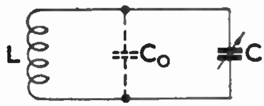


FIG. I.5.—Effect of Self-capacitance of Coil

When a coil is tuned by a variable capacitor, the self-capacitance is effectively in parallel with the variable capacitance (Fig. I.5), and the ratio of maximum to minimum capacitance of the combination is less than that of the variable component $\left(\frac{C_{\text{max.}} + C_0}{C_{\text{min.}} + C_0} < \frac{C_{\text{max.}}}{C_{\text{min.}}}\right)$. This is another reason for keeping the self-capacitance as small as possible.

1.6. Screening

The external field of an inductor will link with other nearby coils and induce e.m.f.s in them. This effect is usually undesirable, and can be minimised in two ways. The first method is to reduce the external field by using a toroidal winding (Fig. I.6). The magnetic

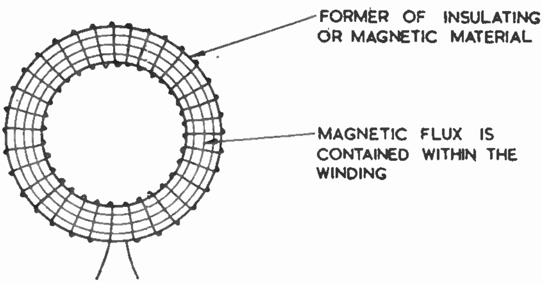


FIG. I.6.—Toroidal Coil

flux occupies the loop in this arrangement, and very little is manifested externally.

Another method is to " screen " the inductor by a metal can enclosing the coil. The can restricts the field of the inductor, preventing it from spreading outside the can by eddy-current action.

The most satisfactory method of obtaining inductors which do not mutually couple is to use screened, toroidal coils, for the small external field of the toroid is easily restricted by screening.

At high frequencies the inductance required is usually too small for the toroidal form to be used, and the screen must be relied on to restrict the external field of a cylindrical coil. A screen made of a good conductor is very effective in preventing the spread of a high-frequency magnetic field, because the field induces eddy currents in the surface of the screen which set up an opposing field. The field therefore penetrates only a fraction of the screen thickness, so that no field is manifested outside the screen.

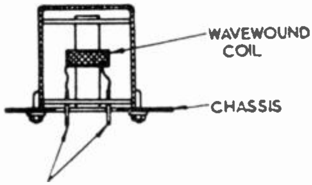


FIG. I.7.—Screened Coil

At the same time the presence of the screen reduces the inductance of the coil by reducing the volume of its magnetic field, and it also increases the effective resistance of the coil because of the eddy-current loss in the screen. This loss is small if the conductivity of the screen is high, and if the screen diameter is twice the coil diameter the inductance is reduced to about 90 per cent. of the value obtained without the screen.

Screening cans are usually pressed from aluminium and have one end open. The can is attached to the chassis, which closes the end and is pierced for the connexions to be brought out (Fig. I.7).

1.7. The Construction of Inductors for Transmitters

The frequency of a transmitter has to be kept very closely to its allotted value in order to avoid interference with other stations on nearby frequencies (see Chapter 7).

The tuned circuits of high-powered transmitters are consequently not continuously variable, but are designed for operation on a few selected frequencies.

The tuned circuits in the earliest stages are built so as to obtain great stability of value by ensuring that the dimensions are maintained at their initial values.

The inductors for the medium-power stages and for the final stages of low-power transmitters are often wound in grooves on porcelain strips or formers, the ends of the coil being secured to terminals. There is little solid material within the coil, so that the minimum of solid dielectric is subjected to the resultant field. The conductor is solid wire of 12 S.W.G. (0·104 in. dia.) or 14 S.W.G. (0·080 in. dia.), often silver plated to resist corrosion and to provide a surface of material of lower resistivity where the current density is highest. Figs. I.8 and I.9 show such coils.

In the final stages of a transmitter signal p.d.s of the order of thousands of volts are developed, resulting in large circulating currents in parallel-tuned circuits. These currents cause losses in the conductors (*i.e.*, I^2R losses), and to minimise the losses the effective resistance must be kept low.

By constructing the inductors for the final stages with copper tube or strip, heavily silver plated, and having a large diameter for

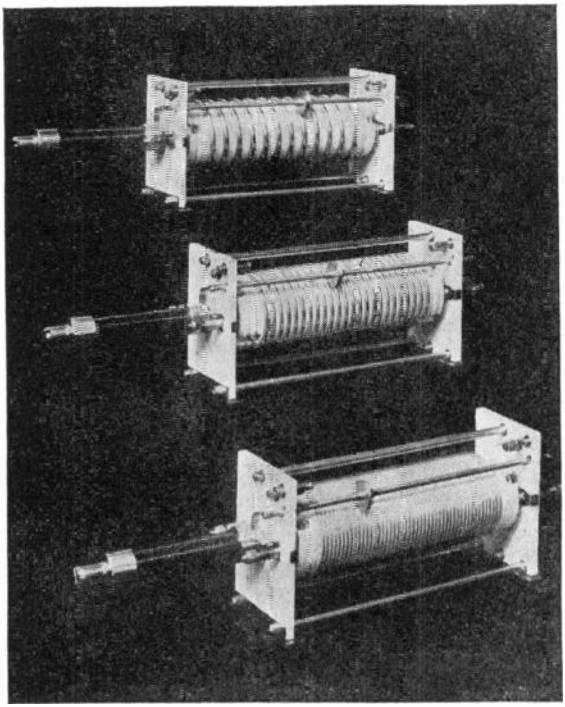


FIG. I.8.—Adjustable Short-wave Coils

the coil, the required inductance is obtained with a few turns. Copper tube enables material to be saved, because the inner layers of the conductor if solid would carry negligible high-frequency current. The coils are mounted clear of other conductors to prevent eddy current losses from occurring.

Coils of this size are either " wound " on a frame, or in the case of coils with few turns are self-supporting and carried on special insulators. The turns of these coils are spaced apart because of

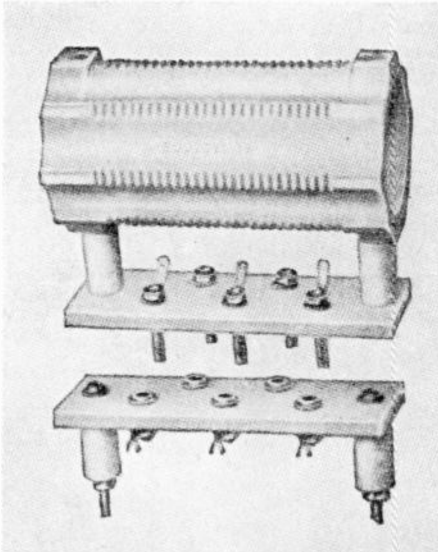


FIG. I.9.—Plug-in Ceramic Former for Radio-frequency Coil

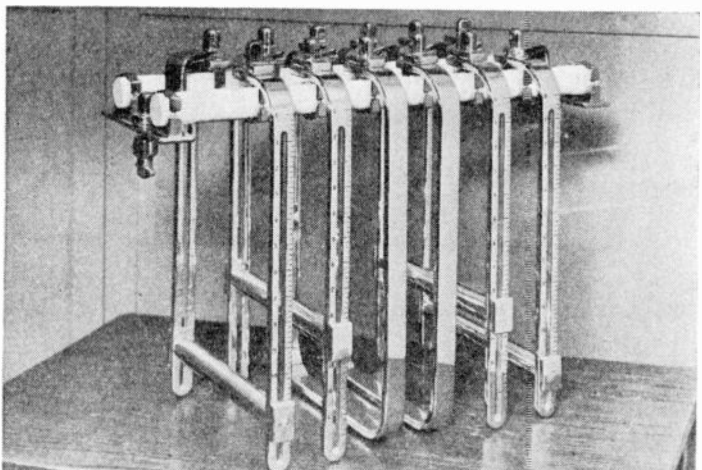


FIG. I.10.—Radio-frequency Anode Coil, Set up for 28 Metres the high potentials existing between turns. Adjustment of inductance is effected by movable links (for initial alignment purposes)

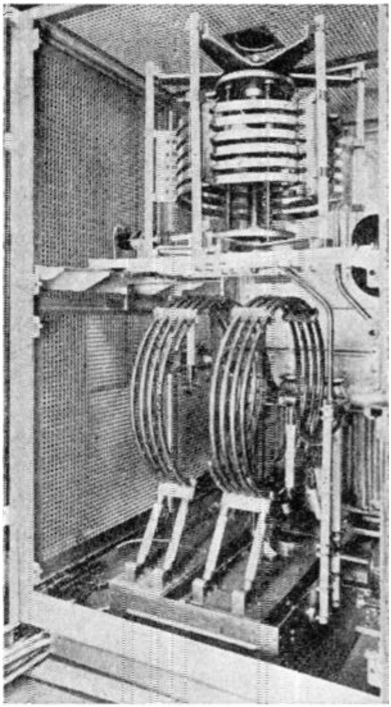


FIG. I.11.—Output Circuit for 100-kW Broadcast Transmitter.

Upper coil has movable contact rollers remotely adjustable by means of the small motor on the left.

or by sliding contacts (for adjustment during operation). Figs. I.10 and I.11 show typical examples.

1.8. Capacitors

The types of capacitors used in receiver circuits were discussed in Volume I of this book, and they are no doubt familiar to the student of Radio.

The energy stored in the components of a tuned circuit is $\frac{1}{2}LI^2$ for the inductor, where L is the inductance in henrys and I is the current in amperes, and is $\frac{1}{2}CV^2$ for the capacitor, where C is the capacitance in farads and V is the p.d. in volts. By having a low $\frac{L}{C}$ ratio for the tuned circuit (whose resonant frequency is determined by $f = \frac{1}{2\pi\sqrt{LC}}$) the same energy is stored with C large in $\frac{1}{2}CV^2$ so that V is smaller, and I large in $\frac{1}{2}LI^2$. Although the circulating current is increased for a given amount of energy, this is preferable to designing for a very large p.d. The construction of the inductor is also simplified by keeping its value down. This arrangement is adopted in transmitters, but in receivers $\frac{L}{C}$ is generally made larger because a high dynamic resistance is required, energy considerations being unimportant.

The early stages of high-power transmitters and the final stages

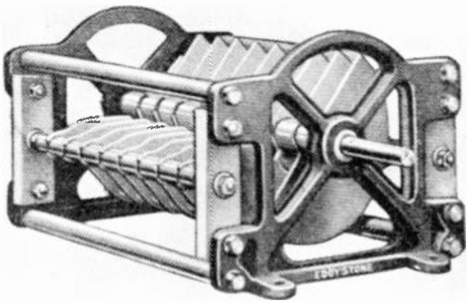


FIG. 1.12. Small Transmitting Capacitor

of low-power transmitters make use of air dielectric capacitors. These are similar in principle to the variable capacitors used in receivers, but they are constructed with thicker plates and larger air gaps to withstand the higher potentials. The insulation necessary to support the fixed plates is located outside the main electric field to minimise the losses (Figs. 1.12 and 1.13).

In small transmitters intended for operation on a few selected frequencies, the value of capacitance is sometimes changed by providing a number of spring-loaded cams, each selecting a certain position of the rotor and hence a value of capacitance corresponding to the desired frequency.

The capacitors used in the tuned circuits of large radio transmitters do not need to be continuously variable as already mentioned (p. 6). Because of the high potentials the dimensions are large to accommodate the necessary insulation; the space between the plates is usually filled with glass, mica or gas, and the whole is contained in a steel case with porcelain or glass insulators to carry the connexions (Fig. I.14). The types using glass or mica insulation are often filled with oil to reduce the possibility of flash-over between the edges of the plates and to conduct away the heat.

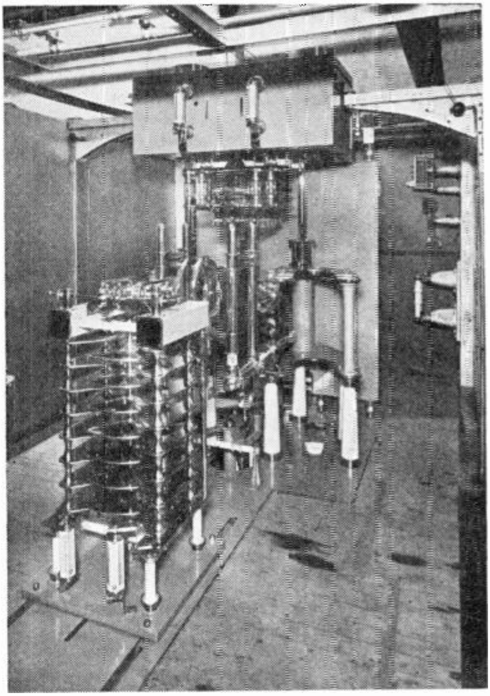


FIG. I.13.—Short-wave Transmitter Amplifier Showing Air-capacitor and Tubular Inductor

Sometimes there is no solid dielectric, the plates being spaced by their fixings and the oil forming the dielectric.

The total capacitance is made up by connecting several units in series, parallel or series-parallel, depending on the capacitance required and the p.d. across it.

Capacitors for working at high potentials have all edges of plates, etc., rounded off to prevent the concentration of electric stress, which occurs at sharp corners and points. High electric stress causes a discharge, visible in the dark (known as "Corona" discharge), and a power loss occurs, while if the stress is further increased there is a possibility of a complete breakdown and consequent damage to other parts of the equipment.

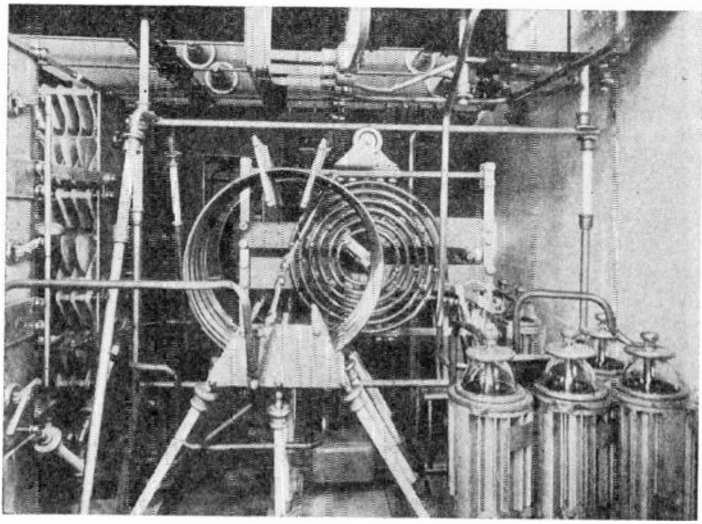


FIG. I.14.—Medium-wave Radio-frequency Stage Showing (*left*) Air-capacitor, (*centre*) Inductors with Arrangement for Altering Coupling and (*right*) Gas-filled Capacitors

The design of capacitors is based on the formula for the capacitance between two parallel plates :

$$C = 0{\cdot}224\ K\frac{A}{d}\ \mu\mu\text{F}$$

where K = relative permittivity, A = area of plates (sq. in.), d = distance apart (sq. in.).

By allowing for the number of working sides of plate the capacitance of a multi-plate unit is easily found. Small capacitors for receivers have plate spacings of the order of 0·012 in., but for transmitters, when potentials are high, a spacing of 0·05 in. per 1,000-volt peak working is usual in air.

Questions

1. Sketch the construction and layout of an inductor and variable capacitor forming the final stage of a tuned circuit in a high-power short-wave transmitter. State the precautions adopted to minimise circuit losses and to prevent breakdown due to high voltages and currents. (C. & G., 1953.)

2. What do you understand by the self-capacitance of a tuning coil? Why is it undesirable, and what steps can be taken to minimise it?

Describe an experiment which you would perform in order to determine the self-capacitance of a typical radio tuning coil.

3. Describe the constructional features of the coils used in the output stages of: (*a*) a 1-kW transmitter operating at 100 kc/s; (*b*) a 1-kW transmitter operating at 10 Mc/s. (C. & G., 1948.)

4. The inductance of an air-cored solenoid of negligible self-capacitance is given approximately by the formula:

$$L = \frac{r^2 n^2}{10(r + l)} \ \mu\text{H}$$

where n is the number of turns and r and l the radius and length of the solenoid in inches.

If $r = l = 1$ in., how many turns are required on the coil if it is to tune to 1 Mc/s when connected in parallel with a 400-$\mu\mu$F capacitor? (C. & G., 1949.)

Answer: 36 turns.

5. Describe with the aid of a sketch, the construction of a 500-$\mu\mu$F fixed capacitor with air-spaced plates, suitable for use in the tuned circuits of a high-power transmitter.

6. State the precautions necessary to avoid breakdown due to large radio-frequency potentials. (C. & G., 1951.)

7. Discuss the chief causes of energy loss in the components used in radio transmitters. (C. & G., 1945.)

8. State the methods adopted in a radio receiver to effect electrostatic and electro-magnetic screening. (C. & G., 1943.)

9. Describe the effect of eddy currents in conductors carrying high-frequency currents. State what steps are taken in practice to minimise these effects. (C. & G., 1929.)

10. Why is the high-frequency resistance of an inductor higher than the d.c. resistance? State the methods that are adopted to reduce the high-frequency resistance in coils. (C. & G., 1931.)

11. Show roughly by means of a sketch the current density in a solid cylindrical conductor when carrying a high-frequency current, explain the reason for any variation in density. What steps are taken in practice to avoid excessive losses and uneconomical use of material in high-frequency conductors? (C. & G., 1933.)

12. Two solid copper wires, one very thin and the other many times thicker, are of such length that their resistance is the same to low-frequency currents. Would you expect their resistances to change to the same extent at very high frequencies? Give reasons for your answer. (C. & G., 1940.)

SPECIMEN ANSWER I

Q. What are the functions of the screens that surround:

(*a*) intermediate-frequency transformers;
(*b*) audio-frequency transformers; and
(*c*) loop aerials?

State the materials used for such screens and explain why the material for (*b*) may differ from that for (*a*) and (*c*). Why is a gap provided in the screen used for (*c*)? (C. & G., 1952.)

A. Intermediate-frequency transformers are screened to prevent coupling between transformers and to prevent direct pick-up at intermediate-frequency. Screening at such frequencies is by eddy-current action, and is most effective when the screen is made of material of low resistivity; copper or aluminium are very suitable materials.

Audio-frequency transformers are screened to prevent the magnetic fields of power transformers or smoothing chokes from inducing e.m.f.s at power of ripple frequency in the windings of the audio-frequency transformer. In this case shielding is best effected by a magnetic material of high permeability; a nickel–iron alloy is generally used. The effectiveness of eddy-current shielding increases with frequency, and is poor with screens of practical thickness at low audio-frequencies.

The screening of a loop aerial has to prevent the capacitance from the aerial to its surroundings from varying with position. At the same time it must permit electromagnetic waves to induce e.m.f.s in the loop, and the screen is therefore gapped to prevent it from acting as a short-circuited turn. Being for use at radio-frequency, eddy-current screening is effective, and aluminium or brass, suitable treated to withstand exposure, may be used.

Specimen Answer II

Q. What do you understand to be the self-capacitance of a tuning coil? Why is it undesirable, and by what means can it be minimised?

Describe an experiment by which the self-capacitance of a coil may be found.

A. The self-capacitance of an inductor is the total effect of the small capacitances between the turns of wire forming the coil, and it may be represented by a capacitance in parallel with the inductance of the coil. The combination therefore forms a parallel resonant circuit whose impedance is inductive below resonance

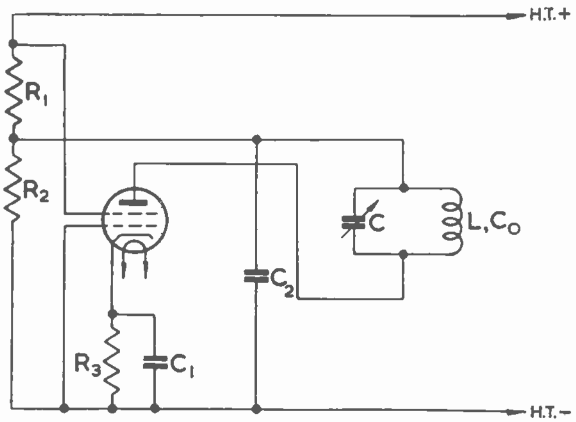


Fig. 1.Q.1

and capacitive above, so that the coil is useful as an inductor only below its " self-resonant " frequency. The effect of the self-capacitance is also to restrict the tuning range when the inductor is tuned with a variable capacitor.

The method of winding the inductor affects the self-capacitance, which in single-layer coils may be reduced by spacing the turns. In multi-layer radio-frequency coils universal—or wave—winding is employed to introduce air-gaps between turns, and in multi-layer audio-frequency coils the thickness of paper between layers (which are also kept as short as possible by splitting the coil into several sections) is increased.

The self-capacitance of an inductor may be calculated from the results of the following experiment:

Connect the inductor in parallel with a calibrated variable capacitor and a dynatron circuit to form an oscillator (Fig. I.Q.1). With a wavemeter measure the frequency of oscillation at several settings of the variable capacitor.

Let L = inductance of coil (henrys);
C = capacitance of variable capacitor (farads);
C_0 = self-capacitance of coil (farads);
f = frequency of oscillation (cycles per second).

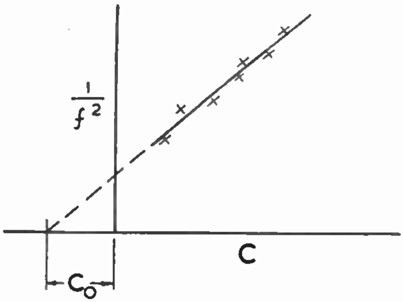


FIG. I.Q.2

Then $$f = \frac{1}{2\pi\sqrt{L(C + C_0)}}$$

or $$f^2 = \frac{1}{4\pi^2 L(C + C_0)} = \frac{1}{K(C + C_0)}$$

where $$K = 4\pi^2 L,$$

$$\therefore \quad C + C_0 = \frac{1}{Kf^2}$$

Plot $\frac{1}{f^2}$ against C, giving a graph as shown in Fig. I.Q.2.

When $$\frac{1}{f^2} = 0, \quad C + C_0 = 0,$$

$$\therefore \quad C_0 = -C,$$

and the self-capacitance is given by the intercept of the line with the x-axis.

CHAPTER 2

ELECTRONIC TUBES

In this chapter the construction and characteristics of electronic tubes are discussed. Some reference is also made to cathode-ray-type tuning indicators, which are special types of electronic valves and to gas discharge tubes. In Volume I the basic principles of the thermionic valve were discussed, and here we shall deal with their construction in more detail. The first important item n a thermionic valve is the source of electrons—the cathode.

2.1. Electron Emission

The emission of electrons from metals may be :

(1) Thermionic emission (primary emission).
(2) Secondary emission.
(3) Photo-electric emission.

Thermionic Emission

In Volume I the electrical conductivity of materials was discussed, and it was found that this electrical conductivity depended on the number of electrons not attached to any particular molecule, known as " free electrons ". The greater the number of free electrons there are, the better is the electrical conductivity of any material. Although these free electrons move about in the material at normal temperatures, the movement is not sufficient for the electrons to escape into the neighbouring atmosphere because of a " surface attraction " which tends to keep them within the material. With increase in temperature the speed of the electrons is increased, and at high temperatures (different for various materials) an appreciable number of electrons is able to overcome the surface attraction of the material. This is known as thermionic emission. For thermionic emission to take place, energy is required, and the amount of energy imparted to an electron to cause it to leave the surface of the material is known as the " work function ".

Secondary Emission owes its importance to the possibility of more than one electron being ejected for each incident electron, enabling " multiplier " tubes to be made, and also causing difficulties in the operation of normal valves.

Photo-electric Emission of electrons takes place at the surface of certain materials when light (a form of energy) falls on them.

2.2. Thermionic Emitters

The types of emitter used in this case fall into three groups :

(1) Pure metals.
(2) Alloys.
(3) Oxide-coated metals.

Thermionic emission of electrons does not take place satisfactorily until temperatures between 700° and 2,500° K. are reached, depending on the type of emitter (° K. = ° C. + 273). At these high temperatures very few substances are suitable.

Pure Metals

Pure tungsten (melting point approx. 3,500° K.) is the only example of a pure metal, and is operated as a thermionic electron emitter at about 2,500° K. Whilst it is a relatively poor emitter having a work function of 3·6 units, it is used for cathodes of vacuum valves of the highest power because of the great durability required.

Alloys

A combination of tungsten with a small proportion of thorium is known as thoriated tungsten (Langmuir, 1914). It has an electron emission many thousand times that of pure tungsten when operated at the same temperature, but when operated some 600° K. lower than for pure tungsten, appreciable electron emission still takes place. The greater electron emission is caused by the formation of a layer of thorium only one molecule deep on the surface of the thoriated tungsten, which considerably reduces the work function.

The basic processes for the construction of a thoriated-tungsten filament are as follows. A small quantity of thorium oxide is mixed with the tungsten before drawing, and after the valve is assembled the filament is heated electrically (to about 2,500° K.) for a short period ; this is known as " flashing ", and is followed by subjecting the filament to an activating temperature of about 2,200° K. for a slightly longer period known as " glowing ". When " flashing " takes place some of the thorium oxide is reduced to metallic thorium, and the subsequent " glowing " of the filament at the lower activating temperature causes this thorium to form a film one molecule deep over the filament surface. It is this film which provides high electron emission.

During the operation of thoriated tungsten filaments at about 1,800–2,000° K., thorium is continuously being evaporated from the filament surface, and is replaced by diffusion from the interior of the filament as described above.

A further improvement in this type of emitter is obtained by "glowing" it in an atmosphere of hydrocarbon; free carbon is produced, which combines with the tungsten to form tungsten carbide. It is found that the surface of tungsten carbide thus formed allows the metallic thorium to cling much more firmly than to pure tungsten.

This permits working the filament at a higher temperature without destroying the thin surface of thorium and increases the electron emission. It also increases the rate at which metallic thorium may be diffused to the surface to maintain the thin coating required for satisfactory emission, and permits this type to be used in high-power valves.

The main differences between pure metal and thoriated-tungsten filaments are:

(1) Working temperature of thoriated tungsten is lower, therefore life may be longer and less heating power is required.

(2) The work function of thoriated tungsten is 2·5 as compared with that of tungsten, which is 3·6.

Oxide-coated Metals

Oxide-coated emitters consist of a coating of barium and/or strontium on a metal core, the metal being usually tungsten, nickel, molybdenum or sometimes an alloy known as "Konel" metal.

The coating is sprayed on the surface of the metal in the form of carbonates. Activation of the emitter is necessary, and it is heated to a temperature of about 1,500° K. for a short period to reduce the carbonates to the required oxide coating. The normal operating temperature is about 1,000° K. Metallic barium or strontium forms on the surface of the filament, and is evaporated during emission. The optimum depth of the metallic coating for maximum electron emission is again one molecule, and (as in the thoriated tungsten filament) the metallic surface is replaced by diffusion of the oxide coating from the interior.

2.3. Electron Emitter Characteristics

When employed as valve cathodes, the various emitters described above may be heated either: (*a*) directly, or (*b*) indirectly

Directly heated cathodes have the emitting material coated on a tungsten filament whose temperature is raised to the correct value by the passage of an electric current (Fig. II.1).

Indirectly heated cathodes consist of a cylinder of emitting material surrounding a tungsten filament. The tungsten filament is heated to incandescence by an electric current, and the cylindrical cathode is heated by radiation (Fig. II.2).

Reference to Table I shows that the operating temperature

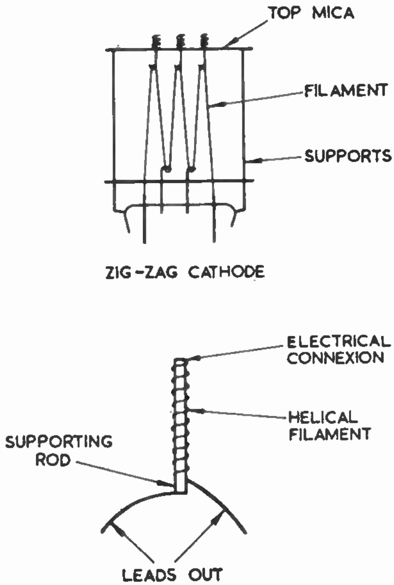


FIG. II.1.—Directly-heated Cathodes

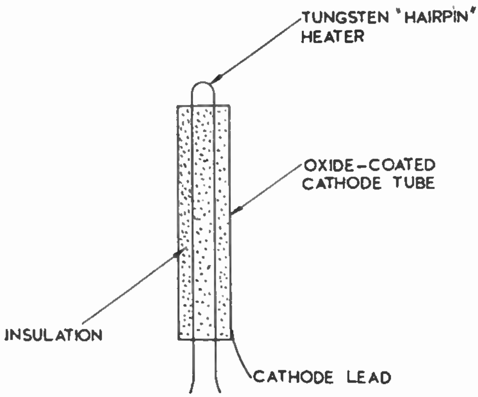


FIG. II.2.—Indirectly-heated Cathode

of tungsten and thoriated-tungsten emitters is much higher than for other types. For this reason, indirectly heated cathodes are always oxide coated because of the difficulty of obtaining the high temperature necessary to operate the other two types. Directly heated cathodes, however, may be of tungsten, thoriated tungsten or oxide-coated metal.

TABLE I

Material.	Operating temperature, ° K.	Emission in mA. per watt of heater power.
Tungsten	2,500	5–10
Thoriated tungsten . .	1,500	40–100
Oxide-coated :		
Directly heated . .	1,000	150–1,000
Indirectly heated . .	1,000	10–250

In a high vacuum such as a valve the heat radiation from an object at a temperature many times that of surrounding objects is proportional to the fourth power of the temperature of the object. This is why emitters like tungsten operating at high temperatures require considerably more power for a given electron emission than do oxide-coated emitters operating at a lower temperature, because the power required increases rapidly with operating temperature (Table I).

2.4. Use and Life of Various Emitters

In large transmitting valves requiring anode potentials exceeding 5,000 volts, a pure tungsten filament is often used. This is the only material that will stand up to the heavy ion bombardment of the cathode caused by the high anode potential and the residual gases in the envelope. As previously stated, their emission is poor compared with other types of emitter (Table I), so that they are used only where conditions are such that the use of other emitters is impossible. Their operating temperature is chosen according to the life and emission required, for as the temperature is increased the filament diameter decreases more rapidly during use, and the emission is insufficient when the diameter is too small.

Thoriated-tungsten emitters are used in valves requiring anode potentials of between 1,000 and 5,000 volts and at higher voltages (up to about 10 kV) when a very high vacuum can be ensured. The life of this type of filament depends upon the thorium content of the tungsten and the rate at which it is used from the surface of the emitter. The correct operating temperature gives an expected life of between 5,000 and 10,000 hours.

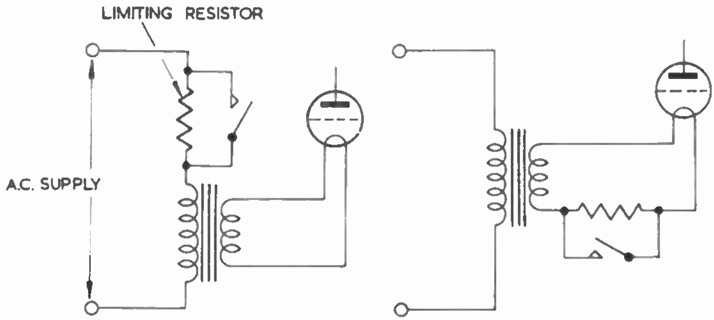


FIG. II.3—Current-limiting Resistors

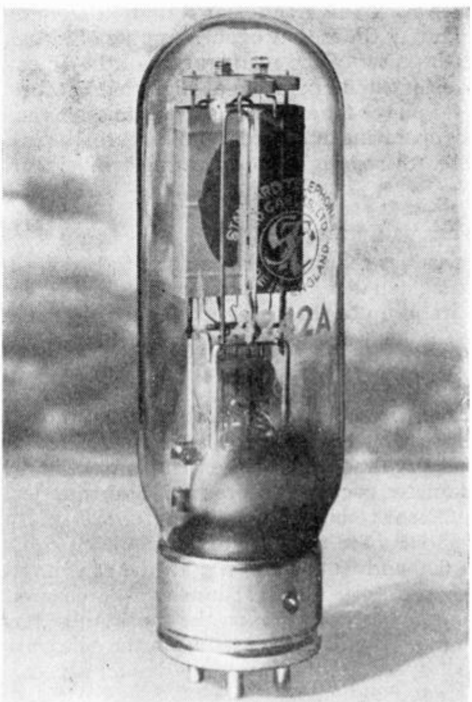


FIG. II.4.—STC 4242A Triode

Anode dissipation, up to 85 W; H.T. supply, up to 1250 V.

Oxide-coated emitters, as can be seen by reference to Table I, are most efficient, and are therefore used whenever possible. Most valves having anode potentials of less than 1,000–2,000 volts employ oxide-coated cathodes, and this category includes practically all radio and television receiving valves in use today. The operating temperature of these filaments is not so critical as with tungsten and thoriated-tungsten emitters, and their life is probably about 10,000 hours.

Practically all high-power valves having an anode potential of more than a few thousand volts employ pure tungsten filaments for the reasons stated above. Reference to valve data shows that the filament power may be as much as 10 kW in the largest valves in order to obtain the required emission.

It will be appreciated that the cold resistance of such filaments is very low, and it is usual to include a resistor in series with the filament circuit when switching on to limit the initial current (Fig. II.3).

Oxide-coated cathodes are never used for continuous operation in high-power valves, because high anode potentials cause ionic bombardment of the cathode, whose coating would be destroyed very quickly.

2.5. Other Electrodes

When a valve is operating, power is supplied to the anode circuit. Some of this power is dissipated in the anode load, and the remainder is dissipated at the anode of the valve itself. We shall be dealing with the ratio of these powers in Chapters 3 and 7, but here it is sufficient to note that in high-power circuits the anode dissipation is never less than about $\frac{2}{11}$ of the power developed in the load. A circuit delivering 11 kW therefore involves an anode dissipation of at least 2 kW, probably $2\frac{1}{2}$ kW in the associated valve. This dissipation takes the form of heat which may be removed by radiation, conduction or convection, and the method of removal depends on the power involved.

A metal or carbon (graphite) anode in a glass envelope may be used up to about 0·5–1 kW, cooling being by radiation. The use of silica for the envelope enables radiation cooling to be used up to about 4 kW; formerly such valves were common up to about 10 kW.

Valves up to this rating follow the construction of the well-known receiving-type valves, and commonly have all their connexions brought out to pins in the base. Figs. II.4 and II.5 show two radiation-cooled valves in this category having glass envelopes. The anodes are graphite or molybdenum. Small receiving valves, having anodes of nickel or molybdenum capable of dissipating up

B

FIG. II.5.—Mullard QY4–250A Tetrode

Anode dissipation, up to 250 W; H.T. supply, up to 4 kV; note screen over grid leads.

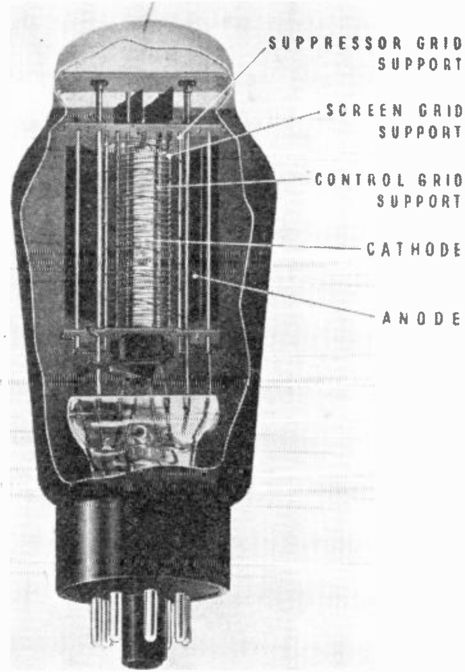


FIG. II.6.—Mullard EL37 Pentode

Anode dissipation, up to 25 W. Half anode removed.

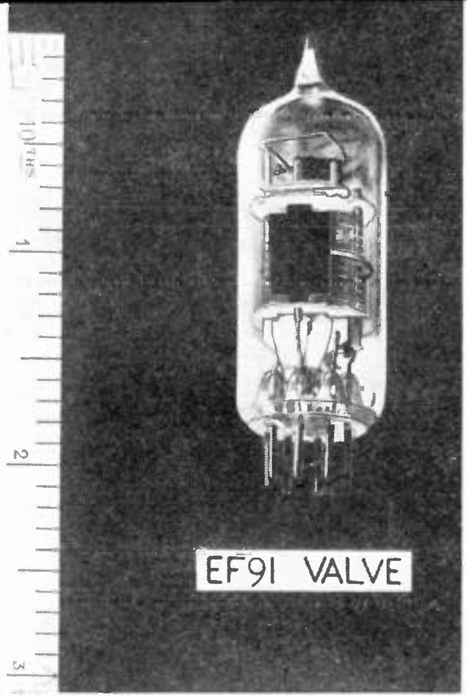


FIG. II.7. Mullard Radio-frequency Pentode
Anode dissipation, up to 2·5 W.

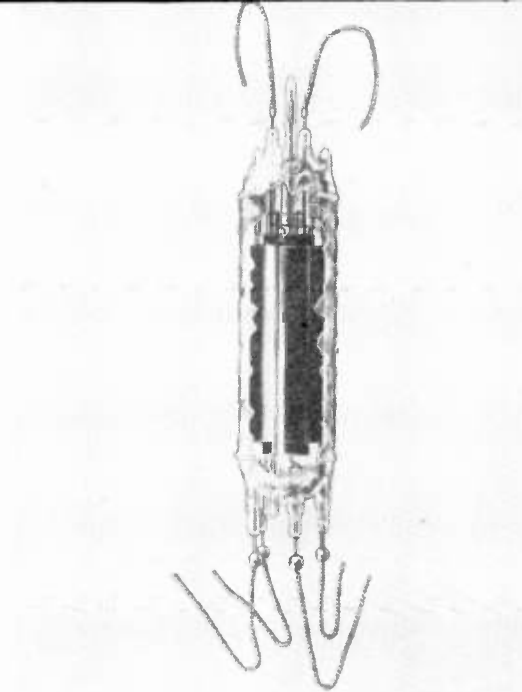


FIG. II.8. Mullard TYS5–3000 Triode in Silica Envelope

Anode dissipation, up to 3·5 kW; H.T. supply, up to 6 kV; thoriated-tungsten filament, 26 A at 20·5 V; length about 20 in.

to 25 watts, are shown in Figs. II.6 and II.7. Fig. II.8 shows a 3·5-kW valve in a silica envelope.

With the exception of silica types, the anode forms part of the envelope of the valve when the dissipation exceeds 1 kW. The

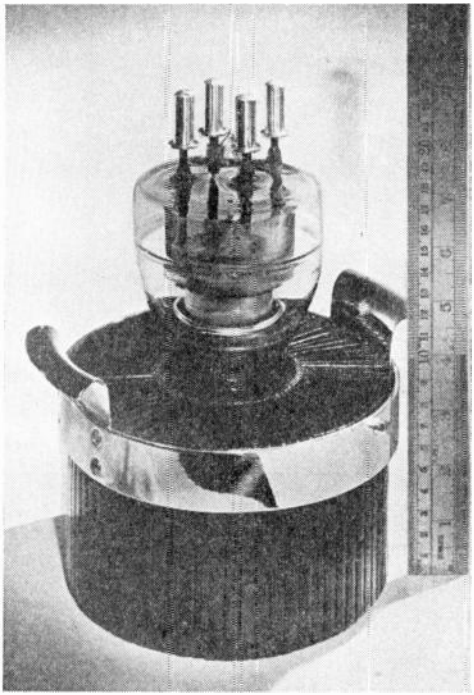


FIG. II.9.—STC 3J/192E Triode

Max. anode dissipation, 4·5 kW; max. H.T. supply, 7 kV; thoriated-tungsten filament, supply 66 A at 5 V. Suitable for full-power working up to 22 Mc/s. Cooling air, 350 cu. ft. per minute.

anode then consists of a cylinder of copper with one end sealed, the other being " welded " to the glass envelope. An assembly of fins, in close metallic contact with the anode, is used with air-cooling up to about 20 kW, or a jacket, into which the anode is placed, is used with water cooling for valves from 5 kW upwards. Figs. II.9 and II.10 show valves of these types.

In high-power valves the grid will be subjected to high potentials, and also grid current will flow. The grid must be capable of dissipating the heat generated. Of the few possible materials, tungsten and molybdenum are the most suitable. The heat is conducted away via the grid support, through the grid seal to the terminal, whence it is removed by air blast.

2.6. Other Features

The evacuation of valves has been described in Volume I, Chapter 4. In the case of high-power valves, where the mass of the electrodes is much greater than in receiving valves, a long

FIG. II.10.—STC 3Q/331E Triode

Max. anode dissipation, 160 kW; max. H.T. supply, 17·5 kV; pure tungsten filament, supply 600 A at 27·5 V; cooling water, 50 gallons per minute; seal-cooling air, 5 cu. ft. per minute; overall length, 40 in.

period of pumping is necessary. At the same time the valve is operated so that the electrodes are as hot as possible in order to drive off occluded gases before sealing off. Tantalum and zirconium when hot are capable of absorbing large quantities of gas, and so anodes are sometimes made of tantalum or coated with zirconium in order to maintain a high vacuum in service.

Large valves (above about 250 watts dissipation) require cooling artificially. To cool the anodes of water-cooled valves distilled water (which has a very high resistivity) is circulated by a pump and is cooled by passing it through a radiator. The radiator and pump are insulated from the anode by passing the water through rubber hoses or porcelain tubes. The grid and filament seals often require air blast for cooling.

Valves having finned anodes are surrounded by a cylinder through which air is drawn by a fan. This air stream also cools the grid and filament seals.

Because of the large potentials used in high-power valves, special precautions must be taken in insulating the electrodes from one another. In the largest valves where the anode supply may be 20,000 volts, the anode connector is spaced as far as possible from the grid and cathode leads.

Where valves are to operate at high frequencies the inter-electrode capacitances and the inductance of the connexions must be as low as possible. This requires a small electrode area and large spacing between the electrodes. Obviously a compromise must be made between power dissipation, physical size and inter-electrode capacitance.

2.7. Demountable Valves

The tungsten filaments of high-power valves become thinner during operation, and eventually the filament breaks at the weakest point. Because these failures were at one time a frequent occurrence, it was thought an advantage to make valves in two parts with a bolted joint, so that their filaments could be replaced when necessary and the valve then re-exhausted on site. Such valves were called " demountable valves ". The disadvantages of these valves are that elaborate pumping facilities are required for exhausting the valves, scrupulous cleanliness must be observed during handling and assembly and unless adequate spares are available, the replacement time is considerable. These disadvantages, and the improvements in high-power valve processing and filament design giving longer life, have rendered these valves obsolete.

2.8. Pentodes

The development of the pentode valve has already been described in Volume I, and therefore only a brief description is given here.

A pentode valve has two more electrodes than a triode (Fig. II.11, also Figs. II.6 and II.7).

One of these extra electrodes (known as the screen) is normally at a higher potential than the cathode, and the other extra electrode (known as the suppressor) is normally at cathode potential.

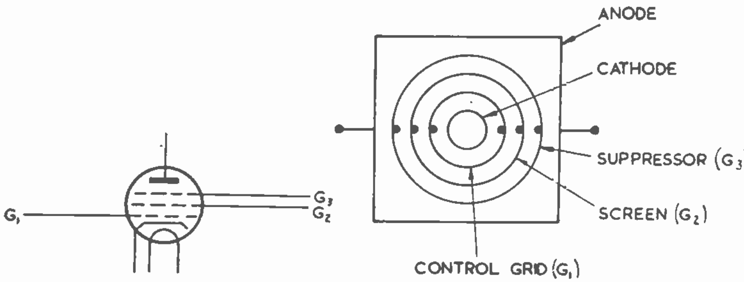


FIG. II.11.—Diagram of Pentode Valve

This latter electrode has the effect of repelling secondary electrons emitted from the anode.

The electrode assembly consists of the cathode as the core, with the other four electrodes arranged around it in the following order :

(1) Control grid (corresponding to grid of triode).

(2) Screen grid (normally at some potential higher than the cathode).

(3) Suppressor grid (normally at cathode potential).

(4) Anode (corresponding to anode of triode).

These differences in electrode assembly and electrode potentials give a set of valve characteristics entirely different from those obtained with a triode valve.

These characteristics of the pentode valve are desirable for many circuits, and indeed many circuits in use today would not be possible without the pentode valve. Radio-frequency pentodes are readily available in which the mutual conductance is controlled by the grid bias. The pitch of the control grid is varied over its length, so that the wires are close at the ends and open at the centre. As the bias is made more negative the close-pitch parts cut off the electron stream, which then passes only through the wider spaces. The rate of decrease of anode current and the mutual conductance fall as the grid bias is increased. Such valves are known as " variable-mu " or more correctly as " variable-g_m " valves. The operation of the pentode valve with its characteristics under various conditions will be considered in Chapter 3.

2.9. Frequency-changer Valves (Hexode, Heptode and Pentagrid)

All frequency-changer valves are required to produce in their anode circuits the sum and difference frequencies of the local

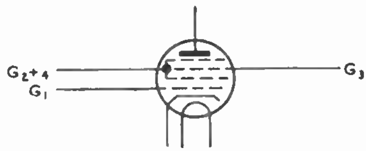


FIG. II.12.—Diagram of Indirectly-heated Hexode

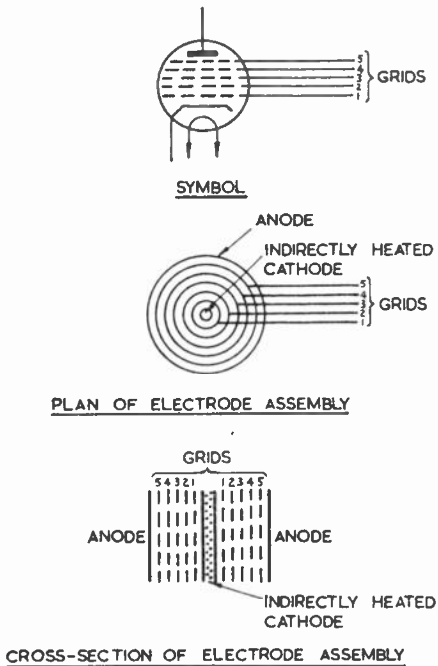


FIG. II.13.—Diagram of Heptode Valve.

oscillator and incoming signal. The difference frequency is then selected and amplified at this new frequency, termed the intermediate frequency. These special valves are used for mixing incoming signals with a local oscillation and have extra grids for the purpose. The hexode and heptode valves are similar in principle.

The hexode valve is a valve having four grids (six electrodes) (Fig. II.12). The heptode valve contains five grids, between anode and cathode (Fig. II.13). Grid 1 is the normal control grid, and is often provided with a variable-mu characteristic. This grid is nearest to the cathode, and has applied to it the incoming signal frequency. Grid 2 is a screen grid, and grid 3 is another control grid. Grid 4 is a second screen, and in the heptode only grid 5 is a suppressor grid, as in an ordinary pentode valve. Grid 3 has the local oscillator output fed to it, and the electron stream produced by grids 1 and 2 is modulated by grid 3.

The pentagrid converter consists of an oscillator and modulator in one envelope. The electrode assembly is similar to that of the heptode, in that there are five grids between cathode and anode, but the connexions and functions are somewhat different.

Triode-hexodes and triode-heptodes are also available, and comprise a hexode (or heptode) and a triode in one envelope, the triode being used as local oscillator and having its grid internally connected to grid 3 of the hexode or heptode section.

Circuits employing frequency-changer valves are described in Chapter 8.

2.10. The Hot-cathode Mercury-vapour Rectifier

The rectifiers we have so far considered (Volume I) have been of the " hard-valve " type, *i.e.*, " high vacuum " having a very small amount of gas left inside the envelope. They suffer from the disadvantage of a high internal resistance, which means a power loss in the valve when delivering current. For most radio receivers and low-power equipment the associated loss may be tolerated, and hard valves are used. Where, however, the power loss is important, it is essential that the internal resistance of the rectifier is kept as low as possible. By introducing a small quantity of mercury into the glass envelope it is found that the full anode current flows for an anode–cathode potential of about 15 volts. The reason for this is as follows: the oxide-coated cathode is heated in the ordinary way, and emits electrons which are attracted to the anode. As these electrons pass through the valve towards the anode, they collide with molecules of mercury vapour, and if their velocity is high enough ionisation takes place. Ionisation is the process of charging molecules by forcibly knocking electrons away from them, and it occurs when the moving electrons have sufficient speed to do this and also avoid being trapped by the molecule's attraction. The ionisation potential of mercury vapour is 10·4 volts, and at this potential the velocity of some of the electrons will be great enough to produce partial ionisation. At 15–18 volts between cathode and anode the velocity of the electrons is sufficiently great for complete ionisation of the mercury vapour to take place, and a blue glow is observed around the electrodes.

When a cathode is emitting electrons in a hard valve, the rate of

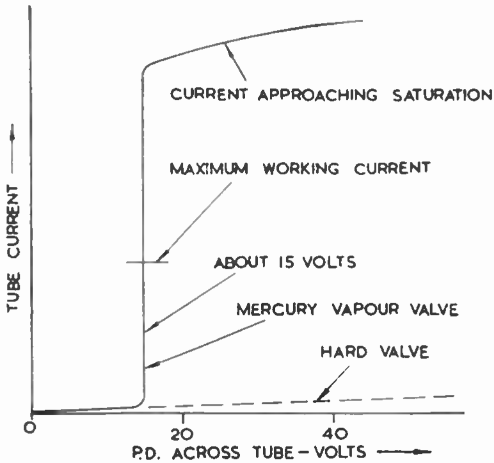


FIG. II.14.—Characteristic of Mercury-vapour Diode

emission is usually higher than that at which the anode can attract them away, unless a high p.d. exists between anode and cathode. As a result a " space charge " or cloud of electrons gathers round the cathode and slows down the rate of emission by repelling electrons tending to be emitted. In the mercury-vapour tube the positive ions of mercury vapour neutralise the electrons in the space charge, which is dispersed when complete ionisation occurs, so that the full emission of the cathode is obtained with a p.d. of only 15–18 volts, instead of perhaps 100 volts. Before this potential (15–18 volts) is reached, the internal resistance of the valve is high, but at and after this potential the internal resistance suddenly decreases to a very low value (Fig. II.14).

If the anode potential is further increased, the current rises above the safe emission of the cathode, the velocity of the positive ions will increase and bombardment of the cathode will result, causing damage to the emitting surface.

It is possible to use an oxide-coated cathode in this type of valve because of the low anode–cathode p.d. when conducting, but care in operation is essential. At a bulb temperature below about 15° C. the ionisation potential rises to a value (about 22 volts) at which cathode bombardment may be injurious, so that the valves must be allowed to warm up thoroughly before switching the power on to the anode circuit. The use of these valves will be discussed in Chapter 6.

2.11. Neon (Discharge) Tubes

Neon tubes, as their name implies, consist (in their simplest form) of two electrodes in a sealed glass container filled with neon

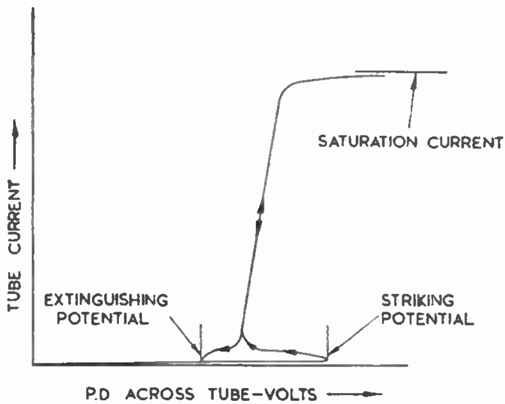


FIG. II.15.—Characteristic of Gas Discharge Tube

gas. When a small direct potential is applied across the electrodes the current passed by the tube is extremely small. This current remains small as the p.d. across the tube is increased until the ionisation potential of the gas is reached, when the current suddenly increases to many times its original value, and unless limited by a resistor external to the tube will result in destruction of the tube. This point is known as the " striking potential " of the tube, and is accompanied by a glow around the electrodes. It will be noted that the action is comparable with the ionisation of mercury in a hot-cathode mercury-vapour valve described above (Fig. II.15).

Further increase in current through the tube will produce very little change in p.d. across the tube. After " striking ", the discharge through the tube may be maintained at a few volts below the " striking potential " (known as the " running potential "). Further reduction in p.d. will cause the discharge to cease suddenly, at a point just a few volts below the " running potential " ; this point is termed the " extinction potential ". From Fig. II.15 it may be seen how close these points are to one another. Because of these properties the neon tube may be used as an H.T. stabiliser or potential reference tube (see Chapter 6). As with the mercury-vapour valve, the ionising potential of the neon tube is fixed by the gas pressure within it. This pressure falls during the life of the tube, gas becoming trapped with electrode material which is deposited on the glass envelope.

2.12. Tuning Indicators

With receivers employing automatic gain control (A.G.C.), it is possible to tune the receiver so that it is not exactly at resonance,

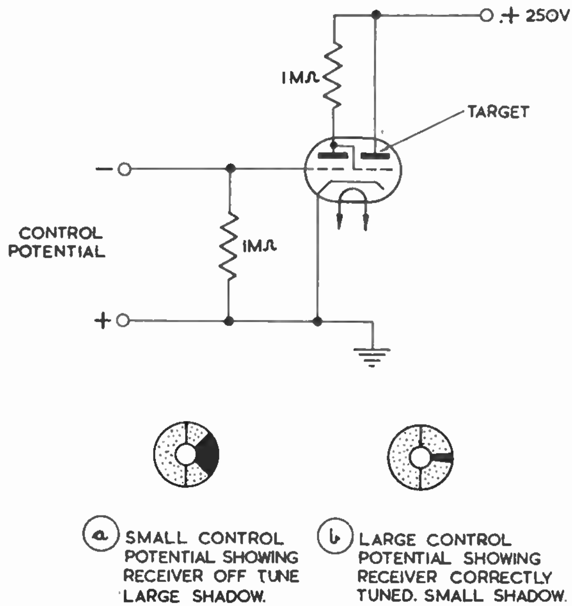


FIG. II.16.—The Cathode-ray Tuning Indicator

because the action of the A.G.C. is to maintain the output constant. This de-tuning results in distortion, and some means of accurately tuning such receivers is desirable. A simple method is to use a milliammeter in the anode circuit of one of the valves employing A.G.C. When the circuit is tuned to resonance the negative A.G.C. bias will be a maximum and the anode current a minimum. Thus by obtaining a minimum deflexion on the milliammeter the receiver is tuned to resonance. Another method is to place the milliammeter so that it records the rectified output of the detector circuits, thus when the receiver is at resonance there will be a maximum rectified output from the detector. Therefore in this case tuning is accomplished by observing maximum deflexion of the milliammeter.

Most modern receivers employ what is known as a cathode- or electron-ray-type tuning indicator. This simply consists of a valve that has a fluorescent portion (known as the target) which glows when struck by electrons. The electrons are emitted from an indirectly heated cathode, and by having another electrode internally connected to the anode of the triode portion of the valve, the area of glow is controlled. By providing the triode portion of the valve with variable-mu characteristics, overloading is prevented on strong signals, whilst retaining good angular movement with weak signals. A typical circuit using a cathode-ray-type tuning indicator together with sketches showing the area of glow obtained with varying values of grid bias is shown in Fig. II.16.

The condition of zero grid bias is the maximum positive A.G.C. potential possible (all other values will be more negative), and it is this zero grid-bias value that will produce the largest shadow angle. On tuning through resonance for the desired transmitter the maximum value of grid bias will produce a minimum shadow angle; this value is obtained when the receiver is correctly tuned.

2.13. Cathode-ray Tubes

The cathode-ray tube is a special type of valve in which a stream of electrons is caused to impinge on a fluorescent screen, so producing a luminous spot. The electron beam is arranged to be deflected, generally in two mutually perpendicular directions, so that the luminous spot may be made to appear at any point on the screen. A deflexion in a horizontal direction is known as the "x deflexion", and a deflexion in a vertical direction is known as the "y deflexion", which is similar to the labelling of axes for graphical representation. In fact, the cathode-ray tube can be made to "draw" a graph. The construction of a typical tube is shown in Fig. II.17, and comprises a cathode, grid and anode assembly known as the "electron gun" and two pairs of plates for deflecting the electron beam, contained in a flask-like envelope. The fluorescent coating is applied to the inside of the flat end, while the anodes (commonly three in number) form an electrostatic lens which focuses the electron stream to a spot at the plane of the screen. The electron-beam density, and hence the intensity of the spot, is controlled by the grid-cathode p.d.

Cathode-ray tubes are used in conjunction with associated apparatus to examine waveforms and potentials of all kinds, and their use is further described in Chapter 9. For this " oscilloscopic " work the additional apparatus is conveniently housed

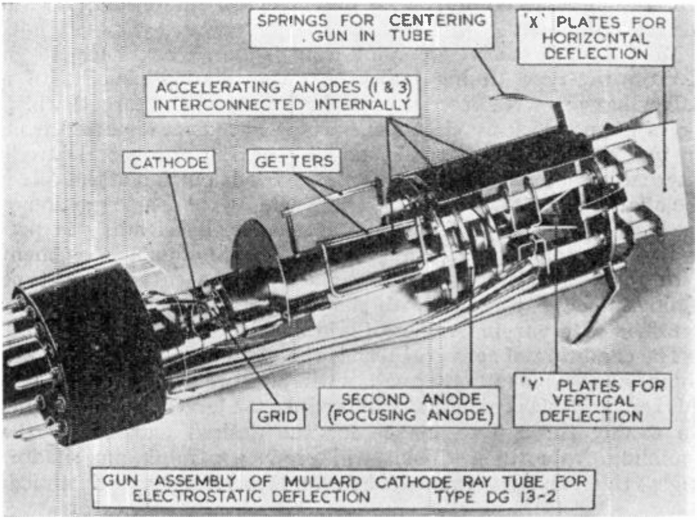


FIG. II.17.—Electrodes of High-vacuum Cathode-ray Tube

with the cathode-ray tube into one unit, such a unit being termed a Cathode-ray Oscilloscope.

Early tubes for oscilloscopes were " soft ", focusing being obtained by ionisation of the gas in the tube, but this type has a limited " writing speed " because of the mass of the ions. The modern oscilloscope tube is highly evacuated and has electrostatic focusing and deflexion.

Cathode-ray tubes are used in television, focusing and deflexion usually being by electro-magnets external to the tube. The electron beam constitutes an electronic current, and is subject to a force at right angles to its direction of flow when passing through a magnetic field.

The fluorescent screen consists of a material known as a "phosphor", and is a compound of cadmium, calcium or zinc. The term fluorescent implies that light is produced only when the material is excited and ceases when the excitation stops. Some phosphors cease glowing more slowly than others, for example, zinc sulphide glows for some seconds after the removal of excitation. The colour of the light also varies with the material; for oscilloscopes a material giving a green light is usual, while if the trace is to be photographed, a material giving blue light is chosen. For television tubes a white light is required.

Questions

1. Draw and describe the main constructional features of a water-cooled triode valve for use in a high-power transmitter. (C. & G., 1949.)

2. Describe with a sketch the main constructional features of an indirectly heated pentode valve for use in the output stage of a broadcast receiver. (C. & G., 1949.)

3. In what respect does a pentode differ from a triode thermionic valve in physical construction and electrical characteristics?

State with reasons which type you would use for a loudspeaker amplifier. (C. & G., 1940.)

4. Trace the development of screen grid and pentode vacuum tubes as amplifying valves, giving reasons for the introduction of the auxiliary grids and an explanation of their functions. (C. & G., 1939.)

5. Discuss the types of filament used in high-power transmitting valves.

6. Explain the method of operation of the cathode-ray tuning indicator (magic eye), and state in what part of the circuit of a receiver it should be connected. (C. & G., 1945.)

7. Explain why it is necessary to cool high-powered transmitting valves.

Describe the methods adopted in practice to cool these valves.

Sketch and describe an air-cooled tetrode valve suitable for use in a 1-kW output stage of a short-wave transmitter. Indicate the approximate potentials applied to the various electrodes and the currents that flow. (C. & G., 1952.)

8. State the materials used, and the reasons for their use, in: (*a*) the filament of a low-consumption battery-operated valve, and (*b*) the heater and cathode assembly of a mains-operated indirectly-heated valve.

Include in your answer sketches showing the forms of construction commonly used. (C. & G., 1953.)

9. Describe with sketches the principles of: (*a*) air-blast cooling, and (*b*) water cooling of high-power transmitting valves.

What are the relative advantages and disadvantages of these two techniques? (C. & G., 1953.)

Specimen Answer

Q. Describe the construction of a beam tetrode valve, and state for what purposes this type of valve is specially suited, giving reasons for its suitability. (C. & G.)

A. The construction of a beam tetrode valve is shown in Fig. II.Q.1.

The electrode system comprises cathode, control-grid, screening grid, anode and special beam-forming plates arranged as shown.

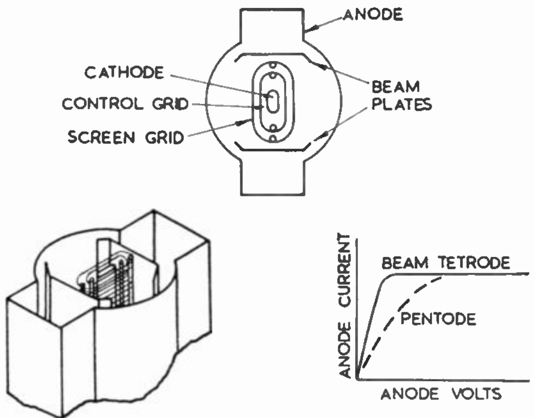


Fig. II.Q.1

The beam-forming plates are internally connected to the cathode, and their function is to concentrate the electrons into a beam and so produce a space-charge effect, which, together with the large screen-anode spacing, results in a potential minimum in the inter-electrode space between screen-grid and anode. This potential minimum forms an opposing field for secondary electrons emitted from the anode, and so causes them to return to the anode, thus performing the same function as a suppressor grid. The control-grid and screen-grid wires are wound to the same pitch and optically aligned, so that the electron stream passes through in the form of flat sheets or horizontal beams. This reduces to a minimum the proportion of the total space current intercepted by the screen grid.

The anode characteristic of a typical beam tetrode is shown in Fig. II.Q.1. It is similar to the pentode characteristic, but the constant-current portion begins more abruptly and at a lower anode potential, due to the absence of the variable-mu effect caused by the suppressor-grid mesh of the pentode. This abrupt transition characteristic gives the beam tetrode an advantage over the pentode as a power amplifier, for which purpose it is chiefly used. Apart from this, the characteristics are similar to those of a pentode. (P.O. Eng. Dept.)

CHAPTER 3

AUDIO-FREQUENCY AMPLIFIERS

THE amplifiers discussed in Volume I are known as voltage amplifiers, which are used to increase the amplitude of signals without consideration of signal power, the load resistance into which the output is delivered generally being high. Power amplifiers were mentioned in Volume I, and we shall now discuss some of their properties.

In this volume we shall assume that audio-frequency power amplifiers work into a load which appears resistive over the frequency range concerned. All the necessary facts are derived from valve characteristics, familiarity with which is essential to a proper understanding of amplifier design.

3.1. Classes of Power Amplifiers

There are three classes of operation of amplifiers, as follows :

A Class " A " amplifier is one in which anode current flows for the whole of the grid-input cycle.

In a Class " B " amplifier anode current flows for approximately one-half the input cycle (180°), the anode current being cut-off during the remainder of the input cycle.

In a Class " C " amplifier anode current flows for less than one-half of the input cycle. This type is used only in radio-frequency amplifiers, and is discussed in Chapter 4.

There are further subdivisions according to whether grid current flows or not, and this is commonly denoted by subscripts. For example, in a Class " A_1 " amplifier anode current flows for the whole of the input cycle, but no grid current flows. A_2 denotes that grid current does flow, but is a condition rarely used in practice. Class " B_2 " amplifiers are much used, however.

To cover the case where anode current flows for more than one-half of the input cycle, the term " Class AB " is used, and such amplifiers are frequently run into grid current at full power. The subscripts, however, are often omitted, as the valve-operating conditions very soon reveal whether or not grid current flows.

3.2. Class " A " Power Amplifiers

The Class " A " condition is used for small (say, up to 15 watts) output stages, where faithful reproduction is important and anode efficiency is not.

We shall derive the value of the load resistance for maximum undistorted output, the anode efficiency and the valve-operating conditions. To avoid a difficult and unnecessary mathematical analysis, we assume that valve characteristics are straight lines apart from the bottom bend, and we choose our conditions to avoid the most curved part of the characteristic. The signal input and grid bias are also chosen so that anode current always flows during the input cycle, and the grid never becomes positive with respect to the cathode, so that grid current does not flow.

Fig. III.1 shows the anode current/grid volts (i_a/v_g) characteristics of a triode. With no anode load the variations in anode current caused by the signal on the grid will follow the i_a/v_g curve

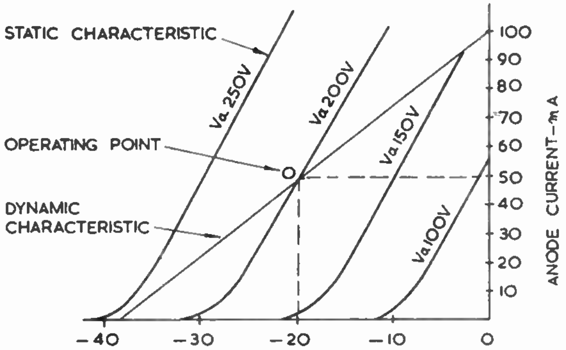


FIG. III.1.—Static and Dynamic Characteristics of a Triode

or static characteristic. Consider now the case where there is an impedance in the anode circuit. When the signal input is a maximum in the positive-going direction, the anode current will increase to a maximum value and the potential of the anode will fall below the static value. The opposite effect occurs when the signal input is a maximum in the negative-going direction; the anode current will be reduced to its minimum value, and the potential at the anode will rise to a value above that of the static condition. These changes in the anode potential are caused by the potential drop across the load resistor in the anode circuit. The effect of these changes is to cause the valve to follow a working or "dynamic" characteristic, having a lower value of slope than that of the static characteristic (Fig. III.1).

We cannot therefore determine the circuit conditions from the static characteristics. It is convenient while working from the dynamic characteristics to allow for the curvature at the lower values of anode current.

Now we have assumed that the load is resistive at the working frequencies, but the load in a power amplifier is usually coupled by a transformer which we shall consider to be perfect. (In practice, transformers may be very efficient or nearly perfect.) The circuit is therefore as shown in Fig. III.2.

The resistance of the primary winding is negligible, so the static anode potential is equal to the supply potential. Increase of

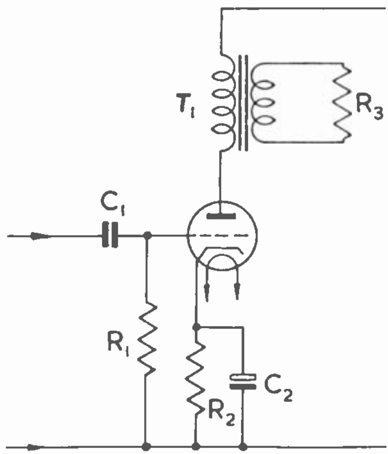


FIG. III.2.—Circuit of a Triode Power Amplifier

anode current caused by a more positive grid potential (above the static value) causes the anode potential to fall below the supply potential, while decrease of anode current causes the anode potential to rise above the supply potential. The difference between instantaneous and static values is given by the induced e.m.f. in the primary inductance, energy being available from the magnetic field set up by the static current in the primary.

3.3. Value of Load Resistance

In order to determine the conditions after allowing for the curvature of the lower part of the characteristic we shall use the i_a/v_a curves. Fig. III.3 shows these characteristics, which are drawn straight above the lower dotted line. The figure shows three i_a/v_a curves, one for zero grid potential, one for the grid at E_g (the grid bias potential) and one at twice E_g. The working point is A, at which the anode potential and current are E_s and I_s respectively.

If the grid potential is zero, the increased current causes a greater drop across the anode load, and point B is reached, while if the grid potential is twice E_g point C is reached. The amplitudes of the current and p.d. across the load resistor are i_m and e_m respectively, and it is required to find the slope of the line BAC, and

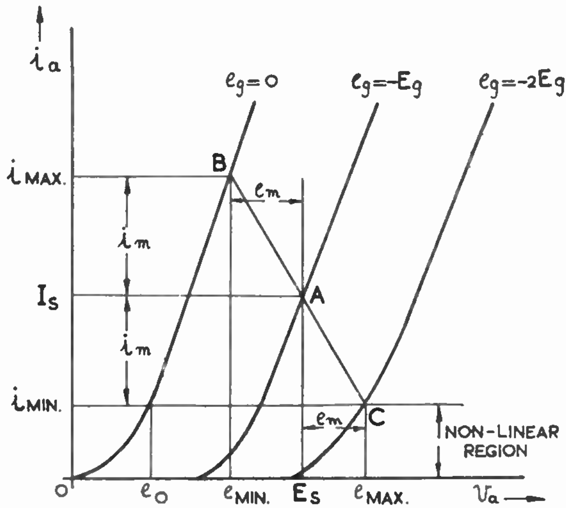


FIG. III.3.—Determination of Operating Conditions of Triode

hence the value of load resistance R_L, which makes the product $e_m \times i_m$ (proportional to power output), a maximum.

The conditions of Fig. III.3 satisfy our requirements of no grid current, and of anode current flowing throughout the input cycle.

We can pick three fixed parameters from the curves and base our calculations upon them :

1. $i_{min.}$, below which the characteristics become curved.
2. e_0, the anode potential causing an anode current $i_{min.}$ at zero grid volts.
3. E_s, the fixed supply potential, also equal to the anode potential in the absence of signal.

We wish to determine the best ("optimum") value of load resistance, R_L.

From the definitions of valve parameters (Volume I, Chapter 5),

$$r_a = \frac{\text{change of anode volts}}{\text{change of anode current}} \text{ (at constant grid potential)}$$

$$\text{we have } r_a = \frac{e_{\text{min.}} - e_0}{2i_m} \left(\frac{\text{total potential change}}{\text{total current change}}\right)$$

$$\therefore \quad e_{\text{min.}} - e_0 = 2r_a i_m \quad . \quad . \quad . \quad . \quad . \quad (1)$$

We can write from inspection of Fig. III.3 :

$$E_s - e_0 = (E_s - e_{\text{min.}}) + (e_{\text{min.}} - e_0)$$

$$\therefore \quad \text{from (1) } E_s - e_0 = E_s - e_{\text{min.}} + 2r_a i_m \quad . \quad (2)$$

Also from inspection $E_s - e_{\text{min.}} = i_m R_L$ . . . . . . (3)

(Peak signal output = peak current × load resistance)

From (2) and (3) $E_s - e_0 = i_m R_L + 2r_a i_m$ . . . . . (4)

Now E_s and e_0 have been fixed; we are finding R_L in terms of r_a for a given set of conditions.

$$\therefore \quad E_s - e_0 = \text{a constant} = E$$

$$\therefore \quad E = i_m(R_L + 2r_a)$$

$$\therefore \quad i_m = \frac{E}{R_L + 2r_a} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

Now the output power is given by

$$P_0 = \left(\frac{i_m}{\sqrt{2}}\right)^2 R_L$$

$$= \frac{1}{2}\left(\frac{E}{R_L + 2r_a}\right)^2 R_L$$

$$P_0 = \frac{E^2}{2} \frac{R_L}{(R_L + 2r_a)^2} \quad . \quad . \quad . \quad . \quad . \quad (6)$$

To find R_L for maximum P_0 we differentiate P_0 with respect to R_L and equate the differential coefficient to zero.

$$\frac{dP_0}{dR_L} = \frac{E^2}{2}\left[\frac{(R_L + 2r_a)^2 - R_L 2(R_L + 2r_a)}{(R_L + 2r_a)^4}\right]$$

$$= \frac{E^2}{2(R_L + 2r_a)^4}\left[(R_L + 2r_a)^2 - R_L 2(R_L + 2r_a)\right]$$

and if $\dfrac{dP_0}{dR_L} = 0$ then $(R_L + 2r_a)^2 - R_L 2(R_L + 2r_a) = 0$

$$(R_L + 2r_a)^2 = 2R_L(R_L + 2r_a)$$

$$R_L + 2r_a = 2R_L$$

$$\therefore \quad 2r_a = R_L \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

The output power can also be written in another form. The power is given by the product of r.m.s. current in and r.m.s. p.d. across the load.

$$\therefore \quad P_0 = \frac{1}{2}\frac{(e_{\text{max.}} - e_{\text{min.}})}{\sqrt{2}} \times \frac{1}{2}\frac{(i_{\text{max.}} - i_{\text{min.}})}{\sqrt{2}} \text{ (from Fig. III.3)}$$

$$= \frac{(e_{\text{max.}} - e_{\text{min.}})(i_{\text{max.}} - i_{\text{min.}})}{2 \times 2 \times \sqrt{2} \times \sqrt{2}}$$

$$P_0 = \frac{(e_{\text{max.}} - e_{\text{min.}})(i_{\text{max.}} - i_{\text{min.}})}{8}$$

3.4. Value of Grid Bias

Fig. III.3 enables us to determine the grid bias for the above conditions.

From the fact that gain $= \dfrac{\mu R_L}{R_L + r_a}$

we have
$$\frac{E_s - e_{\text{min.}}}{E_g} = \frac{\mu R_L}{R_L + r_a}$$

also,
$$E_s - e_{\text{min.}} = i_m R_L$$

$$\therefore \quad i_m R_L = \frac{\mu E_g R_L}{R_L + r_a}$$

$$\therefore \quad i_m = \frac{\mu E_g}{R_L + r_a} \quad . \quad . \quad . \quad . \quad (8)$$

From (5) and (8) $i_m = \dfrac{E_s - e_0}{R_L + 2r_a} = \dfrac{\mu E_g}{R_L + r_a}$

$$\therefore \quad \mu E_g = E_s - e_0 \times \frac{R_L + r_a}{R_L + 2r_a}$$

If $R_L = 2r_a$, $\quad E_g = \dfrac{E_s - e_0}{\mu} \times \dfrac{3r_a}{4r_a}$

$$= \frac{3}{4}\frac{E_s - e_0}{\mu}$$

3.5. Maximum Possible Anode Efficiency of Class " A " Amplifier when Delivering Maximum Undistorted Power

Power Output $= \left(\dfrac{i_m}{\sqrt{2}}\right)^2 R_L$

and since from equation (4) $i_m = \dfrac{E_s - e_0}{R_L + 2r_a}$

then Power Output $(P_0) = \dfrac{1}{2}\left(\dfrac{E_s - e_0}{R_L + 2r_a}\right)^2 R_L.$

Now if the valve characteristics were straight, $i_{\text{min.}}$ and e_0 could be made zero.

The power input (P_i) to the anode circuit would then be

$$E_s \times I_s = E_s \times i_m = E_s \times \left(\frac{E_s - e_0}{R_L + 2r_a}\right)$$

$$\therefore \quad P_i = \frac{E_s^2}{R_L + 2r_a} \text{ and } P_0 = \tfrac{1}{2}R_L\left(\frac{E_s}{R_L + 2r_a}\right)^2$$

$$\therefore \quad \text{Anode efficiency } \eta = \frac{P_0}{P_i} = \frac{\tfrac{1}{2}R_L\left(\frac{E_s}{R_L + 2r_a}\right)^2}{\frac{E_s^2}{R_L + 2r_a}} = \frac{\tfrac{1}{2}R_L}{R_L + 2r_a}$$

This is equal to $\tfrac{1}{2}\dfrac{2r_a}{2r_a + 2r_a} = \tfrac{1}{4}$ or 25 per cent.

Practical anode efficiency is less than 25 per cent. because $i_{min.}$ and e_0 must be greater than zero, and so triode anode efficiencies vary from 15 to 20 per cent., the higher values being obtained with large valves working at high values of supply potential. This must not be confused with the maximum power sensitivity of a valve without regard to correct operating conditions for minimum distortion. If distortion is unimportant and may be tolerated (it very rarely can be in audio-frequency work), then the maximum power output for a given grid input is when the load impedance is equal to the internal impedance of the source; that is when $r_a = R_L$

The anode efficiency may be increased by increasing R_L for $\eta = \tfrac{1}{2}\dfrac{R_L}{R_L + 2r_a}$ and if R_L is large compared with $2r_a$, n approaches 50 per cent. However, the power output for a given drive and the maximum output are decreased.

3.6. Triodes in Parallel and Push–Pull

The output obtained from a given valve is limited by its permissible anode dissipation, for example, a Class "A" triode capable of dissipating 15 watts can deliver $0{\cdot}2 \times 15 = 3$ watts at an efficiency of 20 per cent. The cost of valves increases considerably as the dissipation increases, and it may be desirable to use two smaller valves rather than one large one. Two valves may be connected in parallel (Fig. III.4) or "push–pull" (Fig. III.5). In the parallel arrangement the anode load must obviously be one-half of the optimum load of a single valve; the anode current and power output for a given input are both double the values for a single valve. The primary winding of the output transformer has to carry a heavy direct current.

In the push–pull arrangement two great advantages are obtained. First, the static anode currents of the two valves flowing in opposite directions through the two halves of the primary winding produce no resultant direct magnetisation of the transformer core, and the primary inductance is a maximum. Sec-

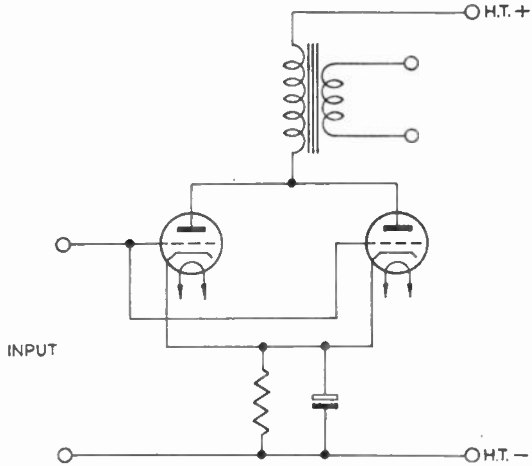


FIG. III.4.—Power Valves in Parallel

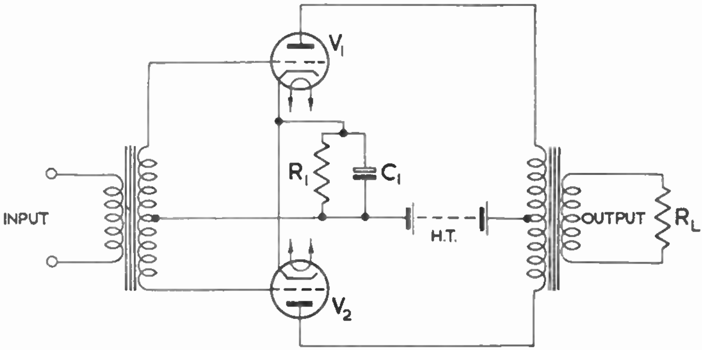


FIG. III.5.—Power Valves in Push-Pull

ondly, the distortion caused by the curvature of the valve characteristics is cancelled out. To produce an alternating p.d. across the primary of the output transformer, the two valves must be driven in antiphase (for example, by an input transformer having a centre-tapped secondary, as shown) so that the current in one valve increases, while that in the other decreases when a signal is applied to the primary of the input transformer. Therefore, when one valve is on the upper part of its dynamic i_a/v_g characteristic, the other is the lower part, thus the effect of the curvature of the characteristic is largely cancelled. The distortion produced by this curvature is equivalent to the introduction of even harmonics —particularly the second harmonic—of the applied signal, and

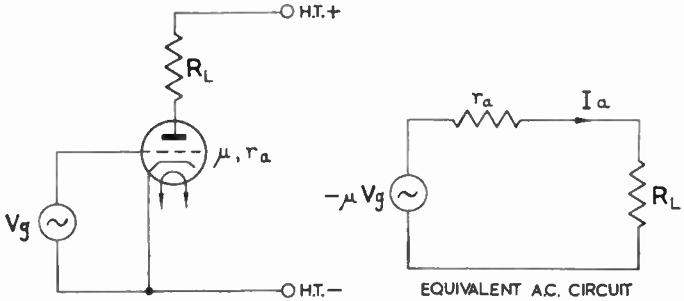


FIG. III.6.—Valve with Resistive Load

therefore it is said that, in a push–pull amplifier, even harmonics are cancelled out. As the two valves are effectively in series with the output-transformer primary, the load resistance presented from anode to anode must be twice that for a single valve.

Push–pull Class " A " triodes are commonly used when outputs between about 4 and 15 watts are required, especially when fidelity of reproduction is important.

3.7. Power Output of Power Amplifiers

The a.c. power delivered to the load is an important factor in all power amplifiers, and it is this " a.c. power " that is referred to when the term " power output " is mentioned.

Consider the case of a valve amplifier operating into a purely resistive load (R_L) (Fig. III.6).

The alternating component of anode current caused by a grid signal V_g (r.m.s.), $i_a = \dfrac{\mu V_g}{r_a + R_L}$

$$\therefore \quad \text{the a.c. power in } R_L = i_a^2 R_L = \left(\frac{\mu V_g}{r_a + R_L}\right)^2 \times R_L$$

Since $\dfrac{\mu}{r_a + R_L}$ is the dynamic conductance (g'_m), the a.c. power in the load (P_0) $= V_g^2\, g'^2_m R_L$.

If the anode load of the valve is a transformer, then the calculation of the a.c. power in the secondary may be more complex. In this volume, however, we are concerned only with " perfect transformers ", that is to say, transformers that reflect purely resistive loads from secondary to primary at all frequencies with which we are concerned. Consider the case illustrated in Fig. III.7, where

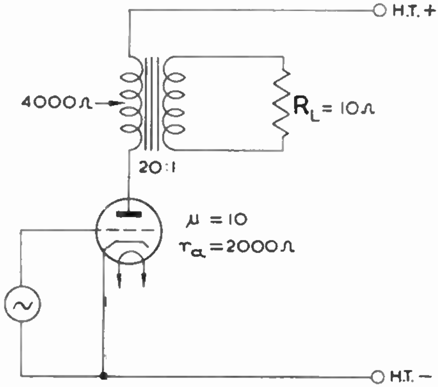


FIG. III.7—Numerical Example

the anode load of the valve is a transformer having a turns ratio of 20 : 1, the secondary of which is terminated in a resistive load of 10 ohms.

The calculation of the a.c. power delivered to the 10 ohms load terminating the secondary of the transformer in Fig. III.7 is as follows:

> The resistance reflected to the primary (*i.e.*, the anode load) is in the ratio of 400 : 1 (square of turns ratio). Therefore the 10 ohms appears as 4,000 ohms reflected from secondary to primary (twice the r_a of the valve, which is a typical value for operating triode valves for maximum undistorted output).
>
> Now a.c. power in load $P_0 = V_g{}^2 g'_m{}^2 R_L$, where g'_m is the dynamic mutual conductance and V_g is the a.c. input to the stage.
>
> If V_g is 10 volts r.m.s., $\mu = 10$, $r_a = 2{,}000$ ohms and $R_L = 4{,}000$ ohms, then

$$P_0 = \left(\frac{10 \times 10}{2{,}000 + 4{,}000}\right)^2 . \ 4{,}000$$

$$= \frac{40}{36} = 1{\cdot}11 \text{ watts}$$

3.8. Class " B " " Push–Pull " Audio-frequency Power Amplifiers Employing Triode Valves

The low efficiency of Class " A " working makes it necessary to use a more efficient system of power amplification, especially when the power output required is appreciable. This is obtained by changing the bias point from mid-way along the i_a/v_g characteristic

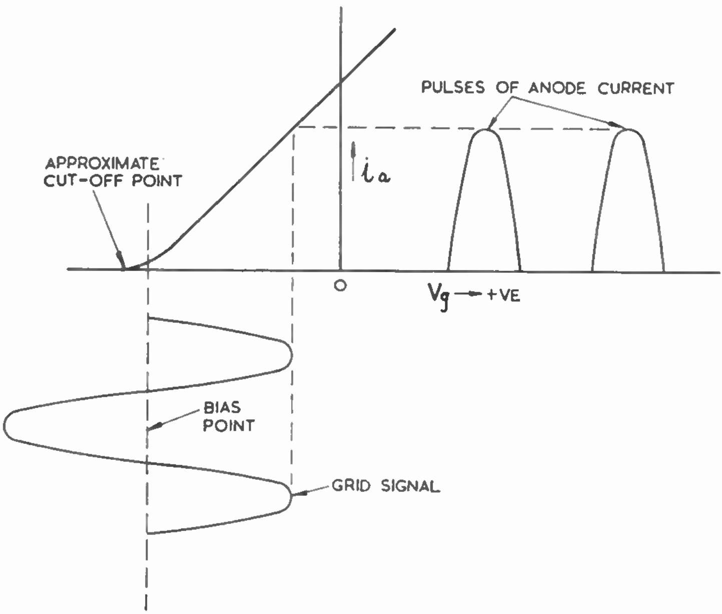


FIG. III.8.—Class " B " Operation

for Class " A " working to approximately anode-current cut-off for Class " B " working (Fig. III.8).

Because of the " tail " of the i_a/v_g characteristic, the bias point for Class " B " operation is referred to as mid-way along the lower bend.

By setting the bias point at or near the anode current cut-off point on the i_a/v_g characteristic, as shown in Fig. III.8, the drain from the power supply under static conditions is made small—so that the " no-load " losses are reduced. With a sine-wave input applied to a Class " B " amplifier, anode current will flow for approximately one half-cycle only. The wave shape produced is similar to that obtained from a half-wave rectifier. The use of two identical valves, whose grids are fed in anti-phase so that one valve amplifies one half-cycle while the second valve amplifies the other half-cycle, is known as Class " B " " push–pull " operation. Consider the amplifier shown in Fig. III.9. The input is via a transformer having a centre-tapped secondary producing e.m.f.s of equal amplitude but opposite in phase with respect to the centre tap. Both valves are biased to the same point by means of a common grid-bias battery. With no input, the static anode current in both valves will be equal and will flow in opposite directions through each half of the centre-tapped output transformer; there is therefore no resultant magnetic flux in the core. Now let a

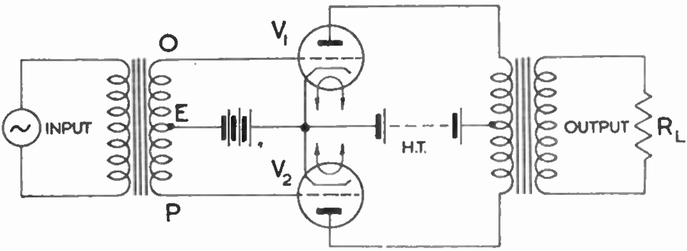


FIG. III.9.—Class " B " Push–pull Amplifier

sinusoidal signal be applied. When V_1 conducts, V_2 is non-conducting, and in the anode circuit of V_1 current will flow proportional to the input signal. During the next half-cycle V_2 will conduct and V_1 will be inoperative. Therefore in each half of the primary of the output transformer currents will flow equal in magnitude but opposite in phase, and across the whole primary winding there will be an alternating p.d., which is an amplified version of the p.d. across the secondary of the input transformer.

3.9. Determination of Anode Load Resistance, Power Output and Anode Efficiency for a Push–Pull Class " B " Stage

Let the load resistance referred to the whole primary of the output transformer be R_L ohms. Then the load on each valve is $\frac{R_L}{4}$ because of the centre-tap. It is correct to consider each valve separately, as only one is conducting at a given instant. Let the peak anode current in each valve be $i_{max.}$

The peak p.d. across a half-primary is then

$$i_{max.} \times \frac{R_L}{4}$$

The minimum instantaneous anode potential of the valve $e_{\min.}$ is given by the supply potential E_s minus the peak p.d. across the half-primary, and therefore $e_{\min.} = E_s - \left(\frac{i_{\max.}R_L}{4}\right)$

$$\therefore \quad \frac{i_{\max.}R_L}{4} = E_s - e_{\min.}$$

$$R_L = 4\left(\frac{E_s - e_{\min.}}{i_{\max.}}\right)$$

The power output (P_0) for the case of a sinusoidal signal is given by the product of r.m.s. current and r.m.s. volts

$$\therefore \quad P_0 = \frac{i_{\max.}}{\sqrt{2}} \times \frac{i_{\max.}R_L}{4\sqrt{2}}$$

$$= \frac{i^2_{\max.}R_L}{8}$$

But

$$R_L = 4\left(\frac{E_s - e_{\min.}}{i_{\max.}}\right)$$

$$\therefore \quad P_0 = \frac{i^2_{\max.}}{8} \times 4\,\frac{E_s - e_{\min.}}{i_{\max.}}$$

$$= \tfrac{1}{2}(E_s - e_{\min.}) \times i_{\max.}$$

The power input is equal to the product of the mean value of the anode current and the supply potential E_s.

The mean value of the anode current is given by integrating the anode current of one valve over half a cycle.

$$i_a = \frac{\omega}{\pi}\int_0^{\frac{\pi}{\omega}} i_{\max.} \sin \omega t dt$$

$$= -\frac{\omega i_{\max.}}{\pi}\left[\frac{\cos \omega t}{\omega}\right]_0^{\frac{\pi}{\omega}}$$

$$= i_{\max.} \times \frac{2}{\pi}$$

If $e_{\min.}$ is 0, the output is equal to

$$\tfrac{1}{2}E_s i_{\max.}$$

$$\therefore \quad \text{Anode efficiency} = P_0/P_i = \frac{\frac{1}{2}E_s i_{\max.}}{E_s I_a}$$

$$= \frac{\frac{1}{2}E_s i_{\max.}}{E_s i_{\max.} \times \frac{\pi}{2}} = \frac{\pi}{4} \text{ or } 0{\cdot}785 \text{ or } 78{\cdot}5 \text{ per cent.}$$

In practice ($E_s - e_{\min.}$) must be less than E_s, and the efficiency is less than 78·5 per cent. by the ratio of $\frac{E_s - e_{\min.}}{E}$, which may reach

0·8 so that practical efficiencies of the order of 60 per cent. in Class " B " may be attained, compared with 20 per cent. for Class " A ".

Some valves designed for use as Class " B " amplifiers (particularly for battery operation, where the reduced no-signal current increases the life of the battery) have i_a/v_g characteristics which pass close to the origin, that is, the anode current at zero grid potential is very small, thus dispensing with bias arrangements. Class " B " amplifiers are commonly driven into grid current at full power, and this means that the previous stage must be capable of delivering the power represented by the product of peak grid volts and peak grid current of the output stage.

3.10. Relative Output from Class " A " and Class " B " Amplifiers

It is interesting at this stage to consider the power outputs obtainable from identical valves operating under various conditions.

Power amplifiers are usually operated under such conditions that the maximum permissible anode dissipation of the valve is reached. In Class " A " amplifiers this occurs at zero output (the mean direct input being constant, the anode dissipation is given by input minus output). Other conditions being disregarded, the output of two valves in Class " A ", whether in parallel or in push-pull, will be twice that of one valve. If the permissible anode dissipation per valve is P_a and the efficiency in Class " A " is 20 per cent., the maximum output of two valves is $2 \times 0{\cdot}2 \times P_a = 0{\cdot}4\ P_a$.

The same two valves in Class " B ", push-pull driven so that the dissipation is P_a per valve, will have an efficiency of about 60 per cent. and the maximum anode dissipation occurs at full output.

Now
$$\eta = \frac{P_0}{P_i} = \frac{P_0}{P_0 + P_a}$$

$$\therefore \quad P_0 = \eta(P_0 + P_a)$$

$$P_0(1 - \eta) = \eta P_a$$

$$\therefore \quad P_0 = \frac{\eta}{1 - \eta} P_a \text{ (per valve)}$$

In this case
$$P_0 = \frac{0{\cdot}6}{1 - 0{\cdot}6} 2P_a$$

$$= 3P_a$$

This is over seven times the output from the same two valves in Class " A ".

3.11. The Pentode as an Audio-frequency Power Amplifier

It was stated in Volume I that the amplification factor of pentodes is higher than that of triodes.

It will also be remembered that for pentodes the anode current is almost independent of anode potential above a certain value. This means that the static and dynamic i_a/v_g characteristics of the pentode for various values of anode load are very similar, and the effect of the screen is to enable a much lower value of $e_{a\,\mathrm{min.}}$ to be reached than in small triodes. The dynamic characteristic of a pentode, however, is more curved than that of a triode, having upper and lower bends and giving rise to greater distortion. Odd harmonics are predominant, and increase with increase of anode-load resistance. Since push–pull operation will reduce only even harmonic distortion, very little reduction of harmonic distortion is achieved by using pentodes in this way. Therefore unless power output with the strictest economy regardless of distortion is the prime consideration, the pentode as a power amplifier compares unfavourably with triodes.

It is possible by special circuit arrangements to reduce the distortion produced by pentodes, and hence take advantage of their higher anode efficiency.

3.12. Tone Control of Audio-frequency Amplifier

For fidelity of reproduction it is usually considered necessary for the output from the loudspeaker at all audible frequencies to be of the same relative volume as the sounds being reproduced. The most satisfactory way of achieving this is to make all the components in the chain—microphone, amplifiers, etc., and loudspeaker—have "flat" response/frequency characteristics. However, other considerations, for example, surface noise on gramophone records, may necessitate departures from linearity of characteristics. The "surface noise" or "record hiss" produced when a record is played is in the higher audio-frequency range from about 5 kc/s upwards, and may be objectionable. It may be reduced by the use of a "scratch filter", which usually precedes the amplifier and greatly attenuates all frequencies above, say, 5 kc/s. This results in the loss of a useful portion of the audio-frequency range, and it is often better to introduce a circuit in the amplifier which causes increasing attenuation with frequency above about 5 kc/s.

The various characteristics used are of one of the following types:

(*a*) low-frequency attenuation;
(*b*) low-frequency intensification;
(*c*) high-frequency attenuation;
(*d*) high-frequency intensification.

In this volume we shall discuss qualitatively the circuits and the effects that they produce.

The simplest way of introducing low-frequency attenuation is to reduce the value of the coupling capacitor between stages. This results in a greater p.d. across the coupling capacitor at low frequencies, and therefore less signal is passed on to the succeeding stage. This type of low-frequency attenuation is commonly used to prevent overloading by low-frequency signals of large amplitude.

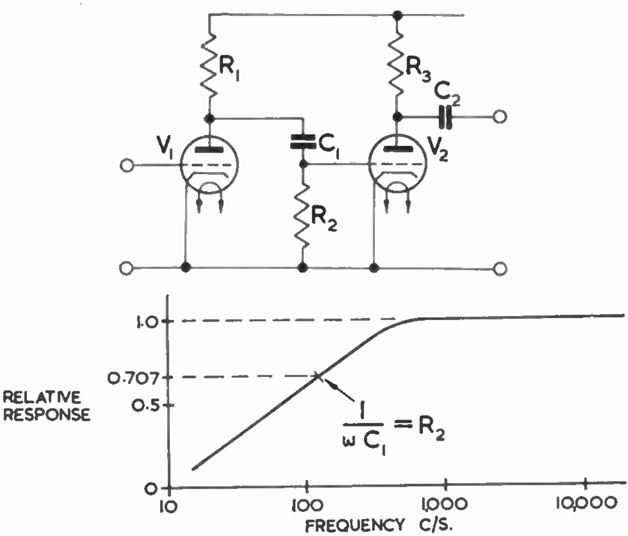


FIG. III.10.—Characteristic and Circuit for Low-frequency Attenuation

The type of characteristic obtained is shown in Fig. III.10. At the frequency at which the reactance of the coupling capacitor C_1 equals the resistance of the grid resistor R_2, the input to V_2 is $\frac{1}{\sqrt{2}}$ times the output of V_1, and the frequency at which this occurs is controlled by choice of C_1 and R_2.

Fig. III.11 shows a circuit for increasing the response as the frequency falls. The coupling capacitor C_1 is chosen so that its reactance is negligible over the frequency range concerned. Resistor R_5 is the grid resistor of V_2. At high frequencies the reactance of C_2 is negligible, so that the output of V_1 is multiplied by $\frac{R_3}{R_3 + R_2}$ and applied to V_2. At low frequencies, however, C_2 offers an appreciable reactance so that the impedance of R_3 and C_2 in series increases, and an increasing proportion of the output of

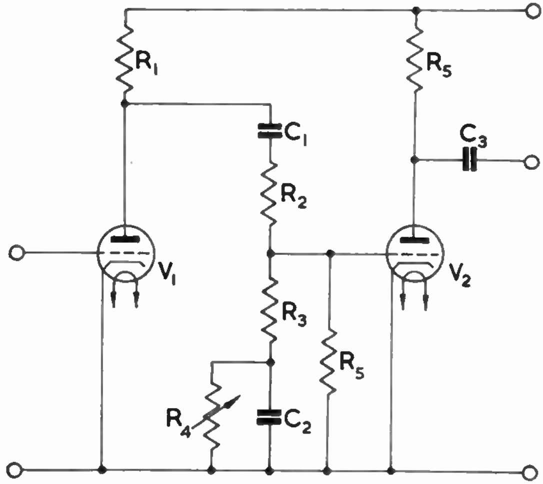


FIG. III.11.—Circuit for Low-frequency Intensification

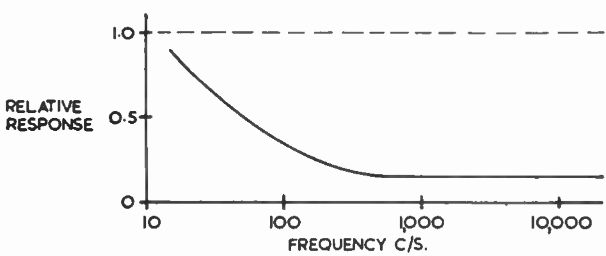


FIG. III.12.—Characteristic for Circuit of Fig. III.11

V_1 is applied to V_2. This proportion may be controlled by adjustment of R_4. It is necessary for R_5 to be high compared with R_3 for the circuit to be effective. Fig. III.12 shows a typical response curve.

The most common method of introducing attenuation at high frequencies is to shunt the anode resistor of one stage with a capacitor (Fig. III.13). The gain of the stage depends upon the load Z_L; at the higher frequencies the total impedance Z_L of the resistor R_1 and capacitor C_1 in parallel will be less than at the low and medium frequencies, therefore less high-frequency signal will be passed on to the grid circuit of V_2, and attenuation of the high

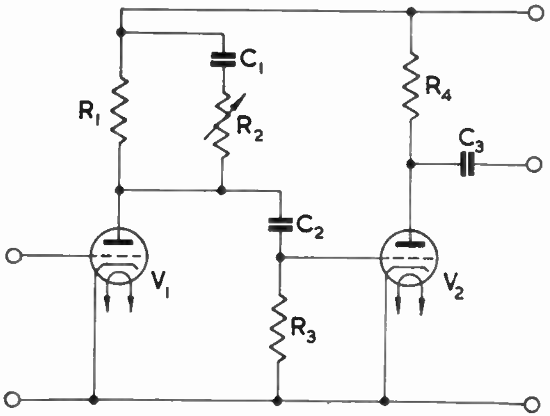


FIG. III.13.—Circuit for High-frequency Attenuation

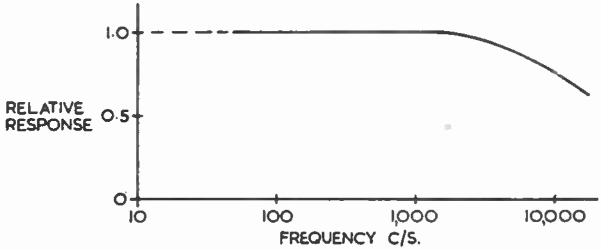


FIG. III.14.—Characteristic for Circuit of Fig. III.13

frequencies will result. By varying the value of C_1 with respect to R_1 the point at which high-frequency attenuation begins is determined (Fig. III.14). R_2 enables the amount of attenuation to be adjusted.

A simple circuit to increase the gain at high frequencies relative to low and medium frequencies is shown in Fig. III.15. The coupling capacitor C_1 is chosen so that the reactance is negligible to all the frequencies concerned. The operation of this circuit is very similar to that for low-frequency intensification; here the parallel combination of C_2 and R_2 offers a low impedance at high frequencies, and a large fraction of the output of V_1 is applied to V_2. At low frequencies the reactance of C_2 increases and a decreasing

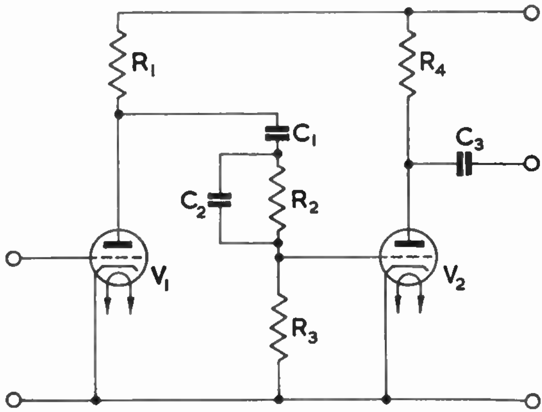


FIG. III.15.—Circuit for High-frequency Intensification

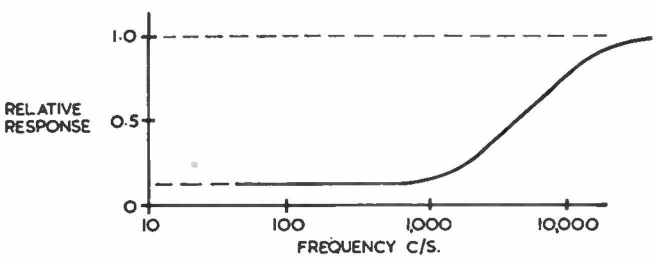


FIG. III.16.—Characteristic for Circuit of Fig. III.15

proportion of the output of V_1 is developed across R_3, as shown in Fig. III.16.

All circuits for intensification must introduce a loss at frequencies not being intensified, and this loss is prevented at low or high frequencies, as in Figs. III.11 and III.15 respectively.

By suitable combinations and adjustment of component values, a variety of overall characteristics may be obtained from circuits of the types shown.

Questions

1. A triode output valve has ideal linear characteristics, for which $\mu = 10$, $g_m = 5$ mA/V. Derive from first principles what should be the load resistance for maximum output power if the resistor is :

(*a*) connected between anode and the H.T. supply ;
(*b*) coupled into the anode circuit by a transformer with step-down turns ratio 20 : 1.

Assume that no grid current flows and the valve is never cut off.
(Brit. I.R.E., Radio Reception, November 1950.)
Answer : (*a*) 2,000 ohms ; (*b*) 5 ohms.

2. Show how you would calculate graphically the power output of a Class " A " amplifier, given a complete set of characteristic curves of the valve. What is the approximate overall efficiency you would expect, under ideal conditions, assuming that the signal input to the grid was the maximum allowable without distortion?
(Brit. I.R.E., Radio Technology, May 1946.)

3. Compare the relative merits of resistance-capacitance, choke and transformer coupling for audio-frequency amplifiers. The answer should include details of stage gain, frequency response, limiting values, etc. (Brit. I.R.E., Radio Technology, May 1946.)

4. A triode valve has the static characteristics given in the accompanying table. Sketch the anode current, anode potential curves using a scale of 0–250 volts for anode potential. The valve is operated as a low-frequency amplifier with a pure resistance anode load of 50,000 ohms, the static working point being $V_g = -2{\cdot}0$ volts, $I_a = 2{\cdot}8$ mA.
Draw a load line and determine the voltage amplification.

V	I_a (mA)		
	$V_g = -1$ volt.	$V_g = -2$ volts.	$V_g = -3$ volts.
70	2·0	1·1	0·6
110	4·7	2·8	1·6
150	7·9	5·5	3·4
190	—	8·9	6·2

5. Distinguish carefully between the amplification factor of a triode valve and the voltage amplification of an amplifier stage

employing the valve. State the relations connecting mutual conductance, anode slope, resistance and amplification factor. At low frequencies a triode with a resistive anode load of 5,000 ohms gives a stage gain of 7·5 times. With a resistive load of 30,000 ohms the gain is 20 times. Determine the static mutual conductance of the valve.

Answer : 2 mA/volt.

6. Explain graphically and with the aid of a circuit diagram, the operation of a Class " B " audio power amplifier.

Assuming identical valves and perfect balance, what are the chief advantages of this system as compared with Class " A " operation for the same total power output?

7. The triode output stage of a radio receiver has an output transformer with a number of tappings on the secondary winding. Describe how you would determine by experiment the tappings for optimum power output into a specified load resistance and state the theoretical relation between the valve anode resistance, the transformer primary/secondary turns ratio and the load resistance. (C. & G., 1948.)

8. Give the circuit of a push–pull output stage for an audio-frequency amplifier and state the advantages as compared with a single-valve stage. (C. & G., 1949.)

9. Explain the principles of operation of either a pentode or tetrode valve and discuss the relative advantages of the one selected and a triode valve as :

(*a*) a high-frequency voltage amplifier ;
(*b*) an audio-frequency output valve.

(C. & G., 1946.)

10. A gramophone pick-up gives 0·5 volts. Give an outline design of an amplifying circuit suitable for delivering 10 watts to a loudspeaker having an impedance of 1,000 ohms. Include details of the valves used. (L.U., 1931.)

11. Describe the push–pull method of amplification, and enumerate and explain its advantages over the single-valve method. (I.E.E., May 1935.)

12. Give a diagram of a low-frequency transformer-coupled amplifier. What tends to limit the amplification of such an amplifier : (*a*) at low frequencies ; and (*b*) at high frequencies? (C. & G., 1936.)

13. The primary of a transformer of 50 : 1 turns ratio is energised from a source having an e.m.f. of 10 volts (r.m.s.) and a resistance of 10,000 ohms.

Calculate :

(*a*) the power delivered by the secondary of the transformer to a load of 10 ohms resistance ;

(*b*) the load resistance for maximum power output ; and

(*c*) the maximum power output.

(C. & G., 1952.)

Answers : (*a*) 2·05 mW. ; (*b*) 4 ohms ; (*c*) 2·5 mW.

14. Draw the diagram of a single-valve power-amplifier output stage suitable for a domestic radio receiver, and mark on your diagram approximate values of bias resistor and by-pass capacitor.

How is such a stage matched to the load? If the optimum load for an output valve is 8,100 ohms, how can it be matched to a resistive load of 100 ohms?

(Brit. I.R.E. Principles of Radio Engineering, November 1950.)

Answer : By means of a 9 : 1 turns ratio transformer.

15. Give the circuit diagram of a Class " B " audio-frequency amplifier stage and, with the aid of sketches showing current and potential waveforms, explain the principles of operation.

What are the advantages and disadvantages of Class " B " compared with Class " A " operation of audio-frequency amplifiers? (C. & G., 1953.)

Specimen Answer

Q. Give the circuit diagram of an audio-frequency amplifier suitable for a public-address system, capable of an output of about 10 watts. The circuit of the H.T. and L.T. power-supply unit may be omitted.

Explain how (*a*) the level and (*b*) the tone quality (frequency response) of the audio output could be controlled. (C. & G., 1951.)

A. The circuit diagram of a 10-watt audio-frequency amplifier suitable for a public-address system is shown in Fig. III.Q.1.

The first stage consists of a pentode voltage-amplifier with a maximum gain of about 150–200. At the input a moving-coil microphone is connected via screened balanced lead to the primary winding of a high-ratio step-up transformer. The closing impedance for the secondary winding is provided by a potentiometer, the slider of which is connected to the control grid of the pentode.

The pentode stage is R.C. coupled to the next, which is a triode amplifier, and this in turn is coupled by a transformer to the push–pull power-amplifier stage. The triode, having a medium anode impedance, provides sufficient current variation in the primary winding of the transformer to produce the necessary drive for the grids of the push–pull power valves. The gain of the triode stage is about 10.

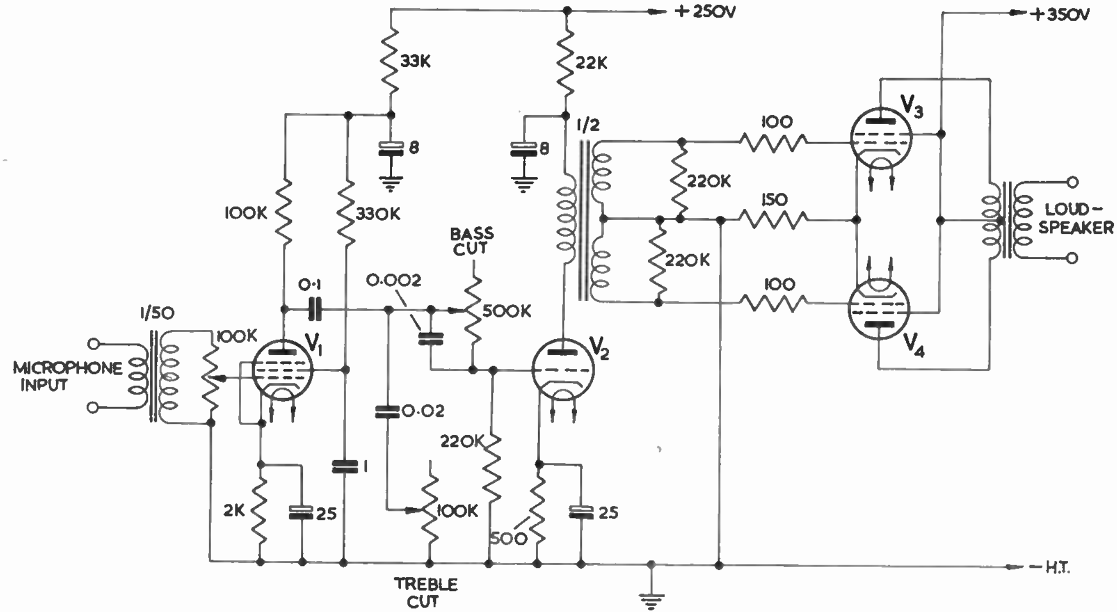

+250V
+350V
33K
22K
8
8
1/2
100
V3
220K
150
LOUD-
SPEAKER
100K
330K
BASS
CUT
220K
100
0·1
0.002
500K
V4
1/50
100K
V1
V2
MICROPHONE
INPUT
0·02
220K
1
2K
25
100K
500
25
-H.T.
TREBLE
CUT

Fig. III.Q.1

In the anode circuit of the power-amplifier stage is an output transformer which matches the impedance of the loudspeaker to the optimum load impedance for the valves.

The potentiometer connected across the secondary winding of the microphone transformer controls the level of the audio output from the amplifier. If this has a logarithmic law, then equal rotations of the control knob will produce approximately equal changes in the level of the audio output.

The tone quality or frequency response of the amplifier can be varied by resistance–capacitance networks incorporated in the coupling between the first and second valves. The first R.C. network, consisting of a 0·02-μF capacitor and a 100,000-ohm variable resistor, provides increasing treble cut as the resistance is decreased. The second R.C. network, consisting of a 0·002-μF capacitor shunted by a 500,000-ohm variable resistor and connected in the series path of the intervalve coupling, provides increasing bass cut as the resistance is increased.

(P.O. Eng. Dept.)

CHAPTER 4

RADIO-FREQUENCY AMPLIFIERS

In Volume I we discussed the elementary principles of radio-frequency amplifiers, showing the use of tuned circuits and of pentode or tetrode valves. The idea of more than one stage of amplification was also introduced, making use of a tuning capacitor with two or more sections mechanically coupled and using switched coils to cover more than one band of frequencies. The

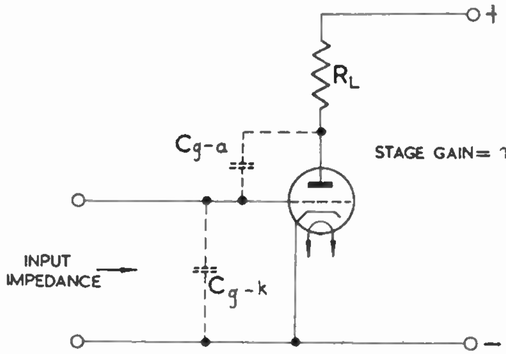


FIG. IV.1.—Triode Amplifier

increased selectivity obtained with two stages compared with one was also noted. In high-power amplifiers for transmitters it may be necessary to use triode valves because of the difficulty of constructing tetrodes and pentodes for use at high power. We must therefore consider the effect of the anode-grid capacitance in more detail than before.

4.1. Miller Effect

The anode-grid capacitance of a triode valve is a few pF, and when a triode valve is built into equipment this capacitance is increased by the valve-socket and wiring.

Consider a triode valve (Fig. IV.1) having a grid–anode capacitance C_{g-a}. If the input impedance of a valve is measured between grid and cathode when the valve is not operating, the result is a capacitance equal to C_{g-a} together with another capacitance C_{g-k}. C_{g-a} will be modified by the presence of the anode load resistor R_L.

Note that this input impedance is measured, in effect, by applying an alternating p.d. between grid and cathode, and observing the resultant current.

Let the valve now be switched on, the stage gain being m, and the measurement repeated. The application of an alternating p.d. V_s between grid and cathode (earth) causes an amplified version equal to $-mV_s$ to be produced at the anode. The negative sign is used because the alternating anode potential is opposite in phase to the grid potential with respect to earth (more positive grid potential increases the anode current and lowers the positive anode potential). The p.d. across the capacitance C_{g-a} is therefore equal to $V_s - (-mV_s) = V_s(1 + m)$. This is shown in the vector diagram in Fig. IV.2. The current flowing through C_{g-a} is

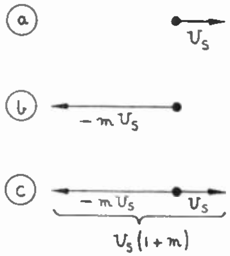


FIG. IV.2.—Vector Diagram for Fig. IV.1

also increased in the same proportion, being given by $\frac{\text{p.d.}}{\text{reactance}}$.

$$i = \frac{V_s(1 + m)}{1/\omega C_{g-a}} = V_s(1 + m)\omega C_{g-a}$$

where $\omega = 2\pi \times$ frequency of V_s.

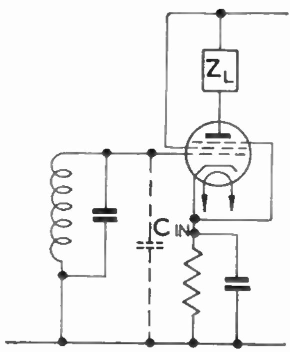


FIG. IV.3.—Pentode Radio-frequency Amplifier with Tuned Input Circuit

Now it has been assumed that the applied alternating p.d. is V_s between grid and cathode and that the input impedance is given by $\frac{V_s}{i}$.

Ignoring the effect of C_{g-k}, the impedance is

$$\frac{V_s}{V_s(1 + m)\omega C_{g-a}} = \frac{1}{(1 + m)\omega C_{g-a}}$$

It will be seen that the impedance has been multiplied by $\frac{1}{1 + m}$

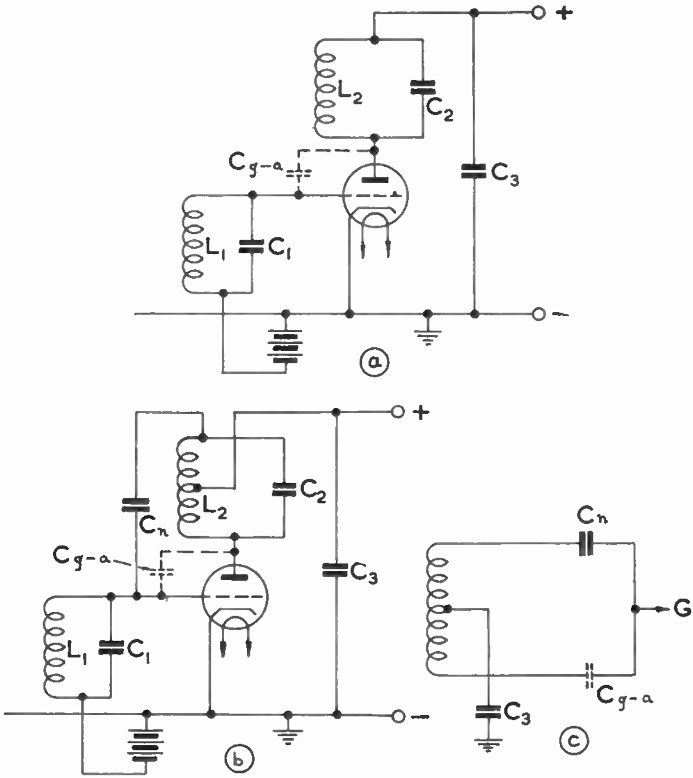


FIG. IV.4.—Neutralisation of Triode Radio-frequency Amplifier

compared with that measured when the valve is not operating (again ignoring the effect of C_{g-k}, which is not affected in this way). This means that the effect of the capacitance C_{g-a} has been multiplied by $(1 + m)$, and this is known as the " Miller effect ".

This high effective input capacitance is characteristic of a triode, and it led to the introduction of the tetrode (and later the pentode) valve, which contains a screen of open mesh, between grid and anode. This screen is held at a steady potential, between those of control grid and anode, and reduces the grid–anode capacitance to a very small value. The gain given by these valves is higher than with triodes, and the input capacitance when working is little greater than the cold value. The capacitance of the grid and the circuit wiring to earth remains, and to obtain a high input impedance the grid circuit is always tuned to resonance at the working frequency (Fig. IV.3).

4.2. Neutralisation

The input capacitance in triode radio-frequency amplifiers may be reduced by " Neutralisation ". In Volume I the effect of anode–grid capacitance was found to be a cause of trouble

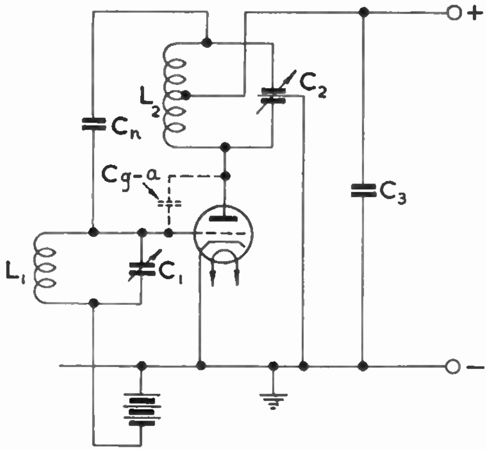


FIG. IV.5.—Neutralised Amplifier with Split Tuning Capacitor

because energy fed back from the anode circuit to the grid circuit through C_{g-a} may cause self-oscillation (see Chapter 5). Neutralisation is a method of reducing the effect of the anode–grid capacitance C_{g-a} by feeding energy back from anode to grid in opposite phase to that fed through C_{g-a}. Fig. IV.4(*a*) shows the capacitance C_{g-a} in a triode amplifier having tuned anode and tuned grid circuits, and Fig. IV.4(*b*) shows a simple way of providing an anti-phase feed through a neutralising capacitor C_n.

The anode inductor is centre-tapped, the tap being at fixed potential so that the alternating potential at the upper end of the

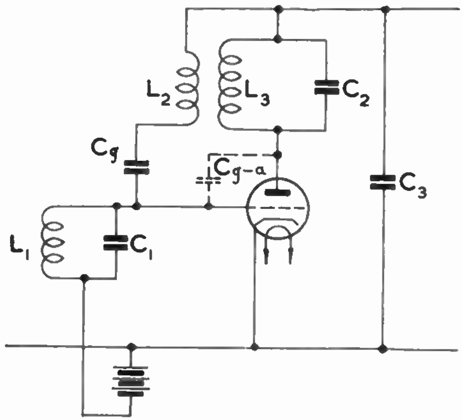


FIG. IV.6.—The Neutrodyne Circuit

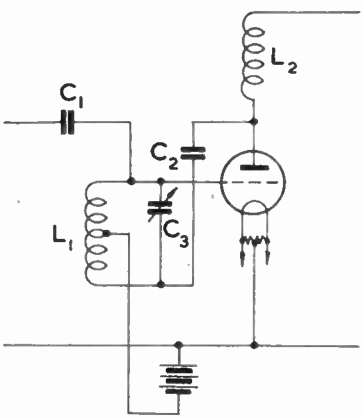


FIG. IV.7.—The Rice Circuit

coil is in opposite phase to that at the lower end. Fig. IV.4(*c*) shows the equivalent circuit of Fig. IV.4(*b*), and it is seen that if $C_n = C_{g-a}$ the grid of the valve is unaffected by change of anode potential. This arrangement is known as the " Modified Neutrodyne " circuit, and although it gives good neutralisation, it has the disadvantage that the load presented to the anode of the valve is reduced to $\frac{1}{4}$ of the value in Fig. IV.4(*a*) because of the tap on the anode coil. Note that the tuning capacitor C_2 has both sides at high radio-frequency potential, and a special type is sometimes used (Fig. IV.5). The original Neutrodyne circuit (Fig. IV.6) used a separate winding L_2 closely coupled to the anode inductor L_3 and connected so as to provide the required anti-phase

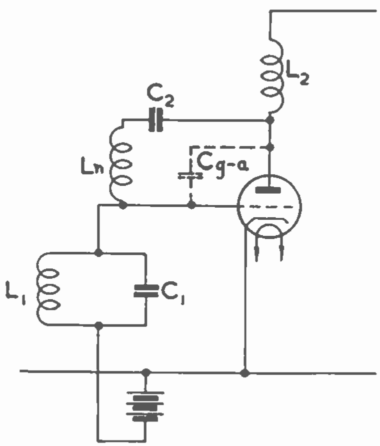


FIG. IV.8.—Coil Neutralisation

potential. This arrangement is seldom used, because it is impossible to have perfect coupling between L_2 and L_3, so that the neutralisation is not exact over a range of frequencies.

The Rice method of neutralisation uses a tapped grid coil (Fig. IV.7) and has similar disadvantages to the modified neutrodyne circuit; here only half the p.d. across the grid coil L_1 is applied to the grid of the valve, and the tuning capacitor again has both sides at high radio-frequency potential.

When an amplifier has to work on one frequency (or a narrow band of frequencies) coil neutralisation may be used (Fig. IV.8). This method operates by making C_{g-a} resonant with L_n at the operating frequency so that the feedback is effectively through a very high resistance. The blocking capacitor C_2 is necessary to isolate the grid circuit from the direct anode supply potential, and the value of C_2 will be large compared with that of C_{g-a}.

A push–pull radio-frequency amplifier is very simply neutralised, for anti-phase points already exist to which the neutralising capacitors may be connected (Fig. IV.9).

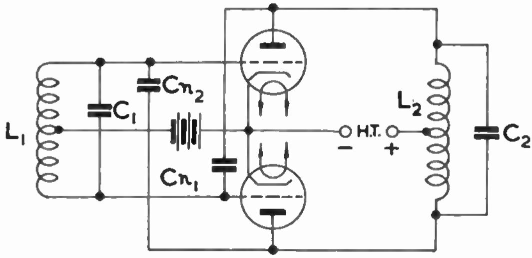


FIG. IV.9.—Neutralised Push–Pull Radio-frequency Amplifier

4.3. Use of Tuned Circuits as Anode Loads

Fig. IV.10 shows a conventional tuned circuit in which the capacitor C is assumed to be perfect while the inductor L has

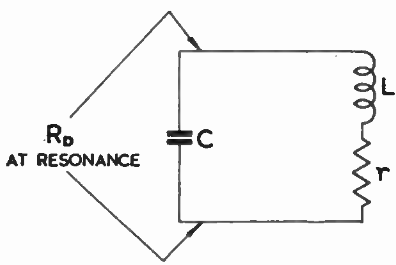


FIG. IV.10

losses which may be represented by a series resistance r. We saw in Volume I that at the resonant frequency $f = \frac{1}{2\pi\sqrt{LC}}$ c/s (L being in henrys and C in farads) the combination appears to be a resistance $R_D = \frac{L}{Cr}$. Also Q, which determines the shape of the response curve, is defined as $Q = \frac{\omega L}{r}\left(\simeq \frac{R_D}{\omega L} = \omega C R_D \text{ at resonance}\right)$, so that we also have $R_D = Q\omega L = \frac{Q}{\omega C}$ (neglecting losses in C).

These equations show that the dynamic resistance for a given Q (and hence a given response curve) is proportional to L. In receivers we usually wish to have high gain, so that R_D should also be high, and for a given frequency L/C should therefore be as high as other considerations permit. Typical values of R_D in receivers are between 50,000 and 250,000 ohms.

In radio-frequency power amplifiers the dynamic resistance of the tuned circuit, together with any parallel resistive load, must be matched to the valve anode resistance to obtain the maximum efficiency. R_D will then be of the order of 2,000–20,000 ohms. The ratio L/C is therefore lower for a given frequency in radio-frequency power amplifiers than in receivers.

4.4. Equivalent Circuit of Pentode Amplifier

The high anode resistance of pentodes means that the anode current is little affected by large changes of anode potential, and it

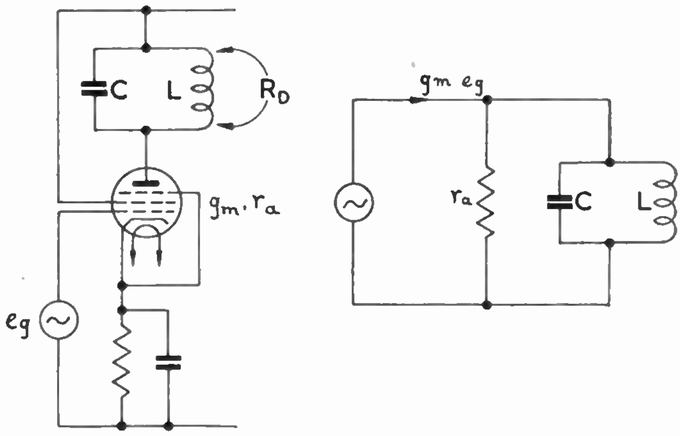


FIG. IV.11.—Pentode Radio-frequency Amplifier and Equivalent Circuit

is mainly controlled by the grid potential. Whereas we use the " series " equivalent circuit for a triode valve, the " shunt " type is more convenient for a pentode (Fig. IV.11). The valve is shown as a generator of a current $g_m \times e_g$ mA, where g_m is the mutual conductance in mA/volt for constant screen and anode potentials and e_g is the signal applied to the grid. The anode resistance is shown in parallel with the load. The equivalence to the series circuit can be shown mathematically (see Volume I, Chapter 5), and the gain is given by

$$m = g_m \frac{r_a R_L}{r_a + R_L} = g_m \frac{R_L}{1 + R_L/r_a}$$

Now if R_L is small compared with r_a (as is usually the case) $m \triangleq g_m R_L$.

The use of a control grid having differing grid wire spacing along its length has already been mentioned in Chapter 2. Fig. IV.12

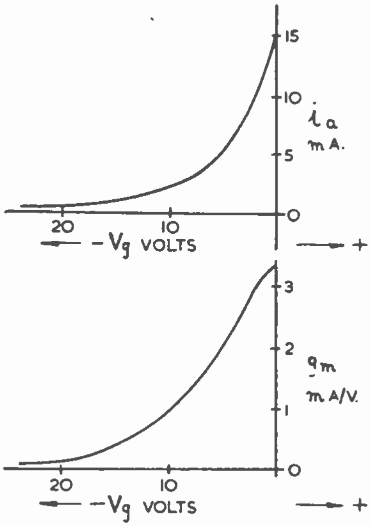


FIG. IV.12.—Characteristics of Variable-mu Pentode

shows i_a/v_g and g_m/v_g characteristics for a typical "variable-mu" radio-frequency pentode, which has a grid of this type. Variable-mu valves are used where it is desired to control the stage gain by altering the grid bias. This is very convenient for manual or automatic control of gain in radio-frequency stages, because the variable-resistor arrangement used in audio-frequency work is not satisfactory at radio-frequencies.

Manual control of gain may be effected by providing a variable cathode resistor or a variable bias supply from a separate battery and "volume control". Automatic control is discussed in Chapter 8.

4.5. Methods of Coupling

The simplest tuned amplifier has the resonant circuit L_1C_1 connected to form the anode load of the valve (Fig. IV.13) and a small

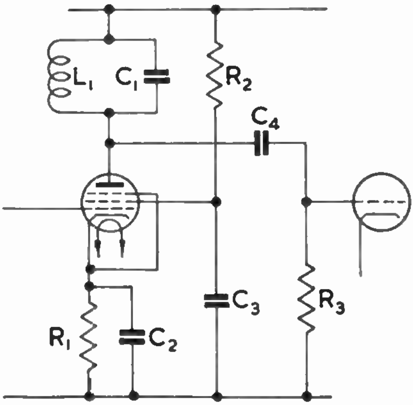


FIG. IV.13.—Tuned-anode Radio-frequency Amplifier

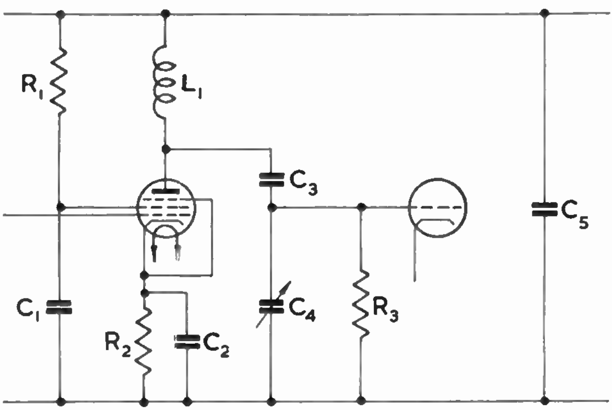


FIG. IV.14.—Modified Tuned-anode Circuit

capacitor C_4 is then used to feed the radio-frequency signal to the following stage, which must have a grid resistor or inductor to apply the necessary bias. This arrangement is much used in power amplifiers, but is not convenient when variable tuning is required, because C_1 is at H.T. potential. A variation of this method is shown in Fig. IV.14, which is common in television and other high radio-frequency amplifiers, while Fig. IV.15 shows a

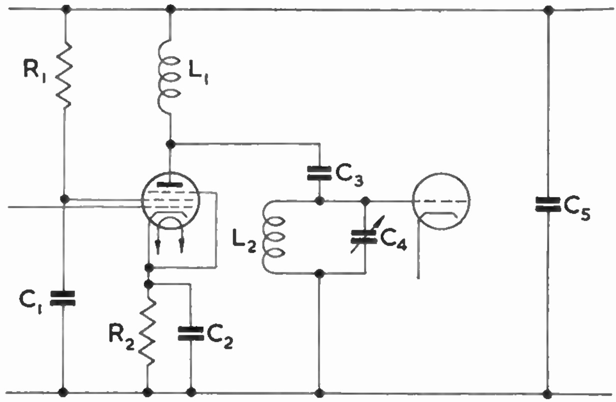


FIG. IV.15.—Tuned-grid Circuit

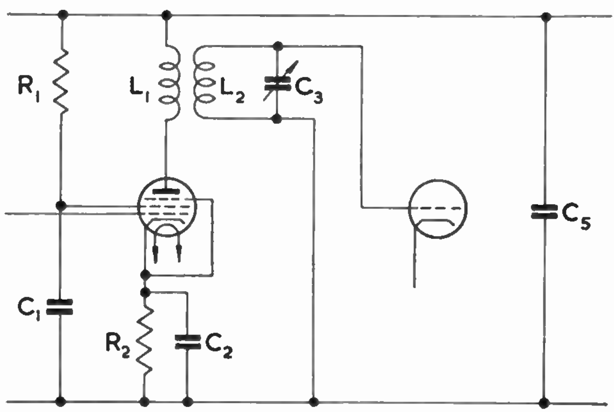


FIG. IV.16.—Transformer-coupled Radio-frequency Amplifier

tuned-grid circuit which is sometimes used in radio-frequency power amplifiers.

In Fig. IV.14 the tuning inductor L_1 and capacitor C_4 are " joined " by the capacitor C_3, so that the whole is effectively a parallel-tuned circuit completed by C_5. In Fig. IV.15, L_1 is a radio-frequency choke which presents a high impedance at the working frequency.

In receivers it is most common to use transformer coupling, as shown in Fig. IV.16. The untuned primary L_1 in the anode circuit presents a high impedance at the resonant frequency of the tuned secondary L_2 C_3, so that the amplification of the valve is then high and the two windings give the required isolation of direct potential. This method avoids damping of the tuned circuit, as no grid resistor is necessary; there is no possibility of secondary effects due to a grid inductor or resistor.

The stage gain is less with this arrangement than when a similar tuned circuit is directly connected in the valve anode, but at frequencies below about 10 Mc/s it provides as much gain as can be used without causing instability.

4.6. Bandwidth

A radio-frequency amplifier handling a modulated signal, whether telegraphy, telephony or television, has to pass a band of

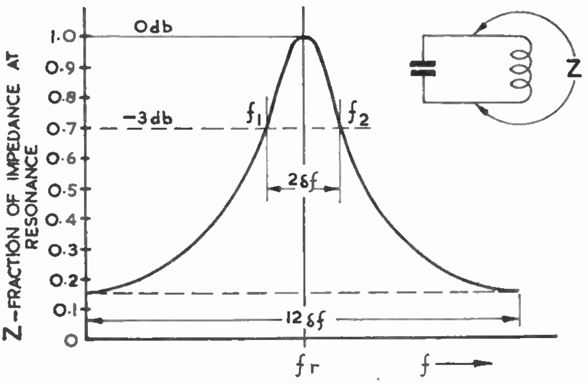


FIG. IV.17.—Part of Resonance Curve

frequencies varying in width from less than 100 c/s for some forms of telegraphy to several Mc/s for television.

For high-quality speech and music it would perhaps be ideal to reproduce all frequencies up to 20 kc/s, necessitating a bandwidth of 40 kc/s (side frequencies equal to carrier frequency + 20 kc/s and carrier frequency − 20 kc/s). While this is possible in a receiver, other considerations limit practical values of bandwidth to 9–20 kc/s, depending on circumstances. Again, it would be ideal to have the same response over the entire bandwidth, but owing to the characteristics of the ear, variations in power output of 2 to 1 are acceptable. Such a power change occurs with a change of p.d. of $\sqrt{2}$ to 1.

The bandwidth of a tuned circuit is defined as the frequency range between the points at which the response is $\frac{1}{\sqrt{2}}$ of the maximum. Fig. IV.17 shows a portion of the response curve of a single tuned circuit, covering about six times the bandwidth. The frequency range to the points f_1 and f_2 where the response is $\frac{1}{\sqrt{2}}$ of the maximum is given by

$$2\delta f = \frac{f_r}{Q}, \text{ where } \delta f = f_2 - f_r = f_r - f_1$$

Note that the output falls with the impedance of the tuned circuit because it is being fed from a " constant-current " source.

This shows that the bandwidth is proportional to the resonant frequency f_r if Q is constant, or if the bandwidth is to be kept constant Q must be proportional to the resonant frequency.

The selectivity of a radio-frequency amplifier is important not only in determining the bandwidth or pass-band but also in the rejection of frequencies further removed from resonance. Fig. IV. 17 shows that at six times the bandwidth the response has fallen to about a sixth of that at resonance.

If an improvement is sought by using two tuned circuits with a valve amplifier between them, each circuit causes a fall $\frac{1}{\sqrt{2}}$ at frequencies f_1 and f_2, and the overall effect is a drop to $\frac{1}{2}$ at these frequencies (the response is down by $\left(\frac{1}{\sqrt{2}}\right)^2 = \frac{1}{2}$ in volts), so that the effective bandwidth is reduced. The Q of each circuit must therefore be reduced to obtain the same overall bandwidth to the $\frac{1}{\sqrt{2}}$ points; the response at $f_r \pm 6\delta f$ is then about $\frac{1}{18}$, so that a considerable improvement over a single tuned circuit is obtained. This improvement can be continued by adding further tuned circuits, but practical difficulties arise if variable tuning is required.

Special methods are required to limit the pass-band to very low values for telegraphy, while in television the wide pass-band is obtained by using a high carrier frequency and tuned circuits having low Q in the amplifier, interference problems being few in this case.

4.7. Coupled-tuned Circuits

An approach to the " ideal " response curve with a flat top and steep sides to reduce the response to unwanted signals near to the desired one is given by coupled circuits. The use of coupled-tuned circuits is most common where tuning is pre-set because of the number of tuning capacitors required. A coupled-tuned circuit is obtained by having two tuned circuits, resonant at the same frequency and weakly coupling them by some impedance common to both resonant circuits—mutual inductance, self-inductance or capacitance.

Fig. IV.18 shows these methods, and Fig. IV.19 shows the type of response curve obtained. The flat-topped response is obtained with " critical coupling ", more coupling producing the " double-humped " curve and less than critical coupling producing a single-peaked curve.

If Q be the value appropriate to the circuits, and k the coupling coefficient, critical coupling is obtained when $k = \frac{1}{Q}$.

In Fig. IV.18(a) $$k = \frac{M}{\sqrt{L_1 L_2}}$$

In Fig. IV.18(b) $$k = \frac{C_c}{\sqrt{C_1 C_2}}$$

In Fig. IV.18(c) $$k = \frac{\sqrt{C_1 C_2}}{C_c}$$

In Fig. IV.18(d) $$k = \frac{L_c}{\sqrt{L_1 L_2}}$$

A variation of mutual-inductance coupling is known as link coupling. This is convenient when mutual-inductance-type coupling is required but the inductors are screened or widely separated. The circuit is shown in Fig. IV.20.

With coupled-tuned circuits it is reasonable to obtain a bandwidth to the $\frac{1}{\sqrt{2}}$ points of 16 kc/s and to have an output less than $\frac{1}{30}$ at 48 kc/s off tune. The addition of a second similar coupled-tuned circuit will reduce the overall bandwidth to the $\frac{1}{\sqrt{2}}$ points to 14 kc/s, while the output at six times this frequency off tune will be $\frac{1}{1000}$.

Such an arrangement is practical only at a fixed mid-frequency, and is widely used in the intermediate-frequency amplifier of superheterodyne receivers.

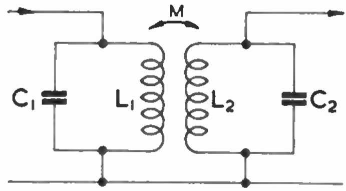


FIG. IV.18(a)

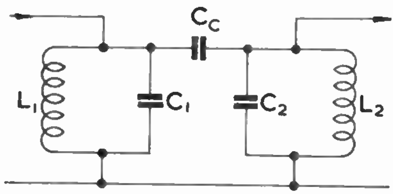


FIG. IV.18(b)

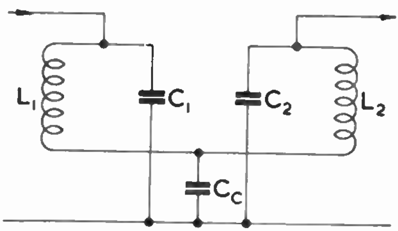


FIG. IV.18(c)

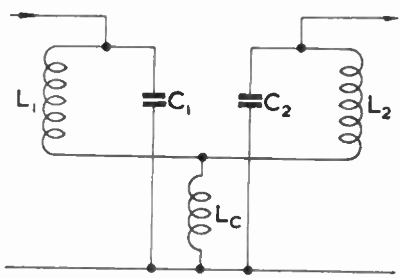


FIG. IV.18(d)

FIG. IV.18.—Band-pass Circuits

The circuit of a typical intermediate-frequency amplifier is shown in Fig. IV.21; these amplifiers commonly work on 465 kc/s, and the usual values of inductance and capacitance are 700 μH and 180 pF respectively. The Q value of the inductor varies from 60

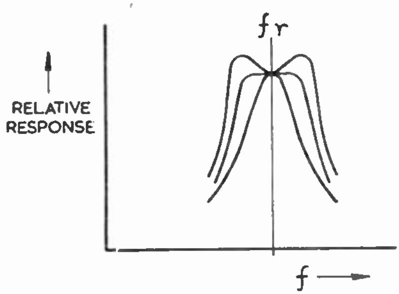


FIG. IV.19.—Band-pass Characteristics

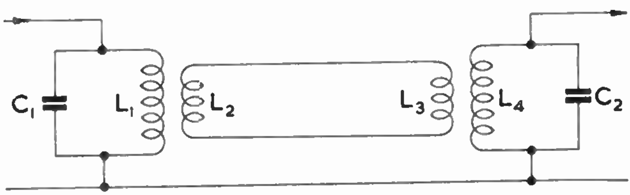


FIG. IV.20.—Link Coupling

to 120, according to the bandwidth desired, the lower value giving a bandwidth of about 16 kc/s, and the higher about 8 kc/s.

Receivers designed for individual requirements may have a bandwidth greater than 16 kc/s, but in general this figure is not exceeded because of the increased interference with the greater bandwidth.

4.8. Power Amplifiers for Transmitters

An important factor in power amplifiers is overall efficiency, that is, the ratio of power output at radio-frequency to power input (direct). The main losses in power amplifiers occur in the valves, other losses in circuit components being comparatively small. The valve "anode efficiency" is defined as the ratio of r.m.s. power output to the tuned circuit to the mean direct power supplied (the product of direct H.T. volts and mean anode current).

The anode efficiency depends on the operating conditions of the valve, and has certain maximum theoretical values in each case.

The maximum theoretical efficiency in a Class "A" amplifier (anode current flowing throughout the cycle) is 50 per cent. (when

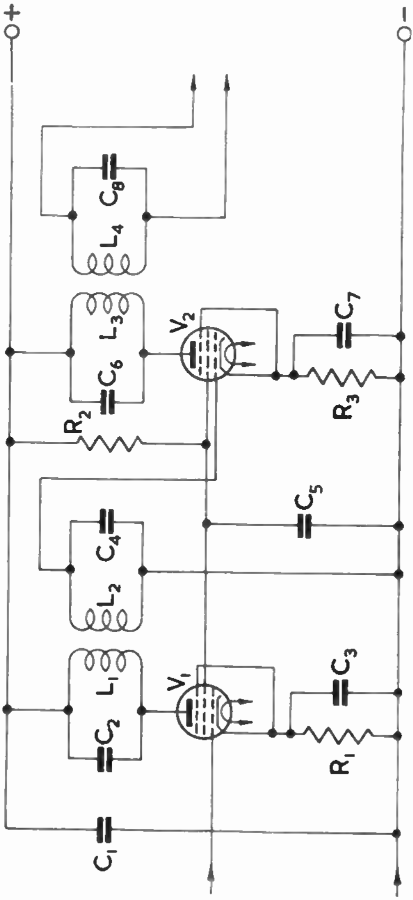


FIG. IV.21.—Fixed-frequency Band-pass Radio-frequency Amplifier

the anode load equals the anode resistance of the valve). It has already been stated in Chapter 3 that this condition of operation introduces amplitude distortion, but when working into a tuned circuit this is of little importance.

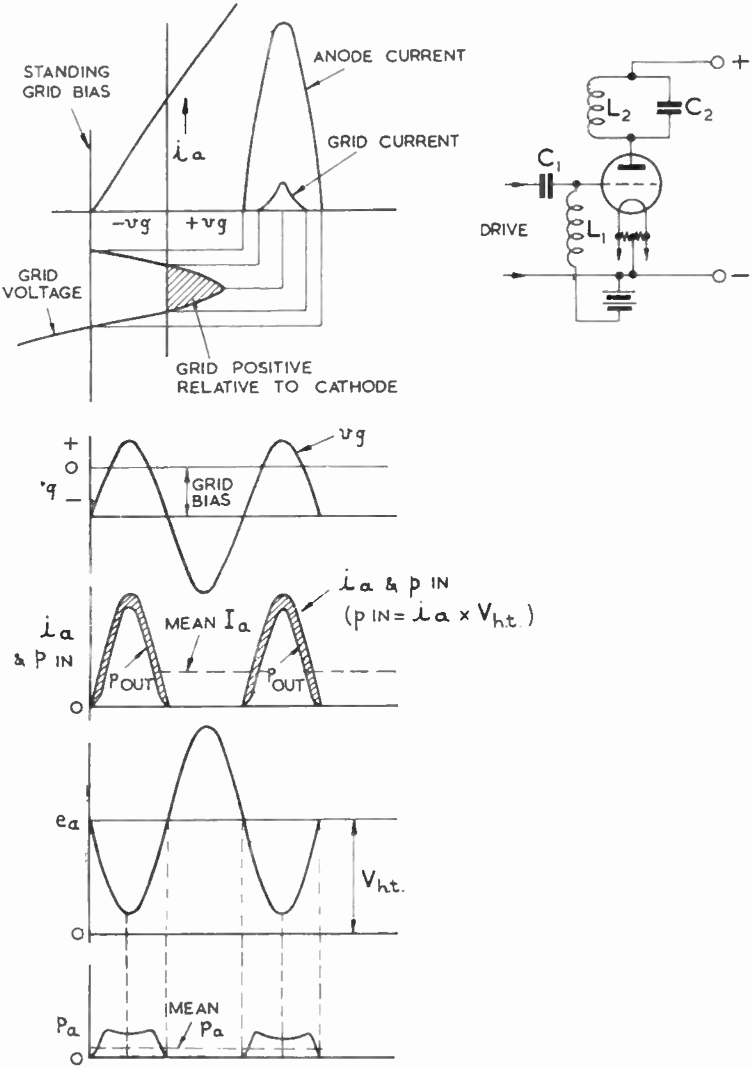


Fig. IV.22.—Class "B" Tuned-anode Amplifier and Waveforms

In a Class " B " amplifier (anode current flowing for approximately half a cycle) the maximum theoretical efficiency is $\pi/4$ or 78·5 per cent. Again, in Chapter 3 it was stated that considerable distortion occurs with a single Class " B " output stage, but when working into a tuned circuit this is of little consequence. As shown in Fig. IV.22, although anode current flows only in half-cycle pulses, the oscillatory circuit causes a complete voltage wave to be developed, and if the Q of the circuit exceeds about 5 the distortion from a sine wave is small. The theoretical anode efficiency is obtained only if the anode potential falls to zero at the peak of the current wave, an impracticable condition, and a typical practical maximum is $\frac{4}{5}$ of 78·5 per cent., or about 62 per cent.

4.9. Class " C " Amplifiers

The efficiency of Class " B " amplifiers is such that the losses are high and expensive in high-power transmitters.

Reference to Fig. IV.22 shows that much of the loss in the valve occurs when the current is low and the anode–cathode p.d. is high. This part of the loss may be reduced or eliminated by increasing the bias on the valve so that conduction occurs only during part of the positive half-cycle of the signal when the anode–cathode p.d. is low, Fig. IV.23. This is known as Class " C " working, which has, however, two disadvantages relative to Class " B ". The first is that the grid driving potential for a given maximum anode current is greater than that for Class " B " because a greater value of bias has to be overcome. Secondly, the output is not proportional to the input, so that amplitude distortion is occurring, and a Class " C " amplifier cannot be used to amplify a modulated input. If the peak grid signal is not equal to the grid bias less the bias for cut-off, no anode current flows, so that small signals do not drive the valve.

An efficiency at full designed output of 85 per cent. can be obtained with a Class " C " amplifier, and 75 per cent. is commonly reached.

Anode modulation can be employed, the grid then being fed with a constant signal.

In order to draw the rated peak anode current at the trough of the anode potential wave, it is necessary to drive the grid positive with respect to the cathode. Grid current therefore flows, and in large triodes the peak grid current is 5–10 per cent. of the peak anode current. Figs. IV.22 and IV.23 show that the grid current is very peaky, and its effect on the previous stage is reduced by providing a parallel resistor R_1 (Fig. IV.24). We have already seen that waveform distortion is unimportant, as the resonant

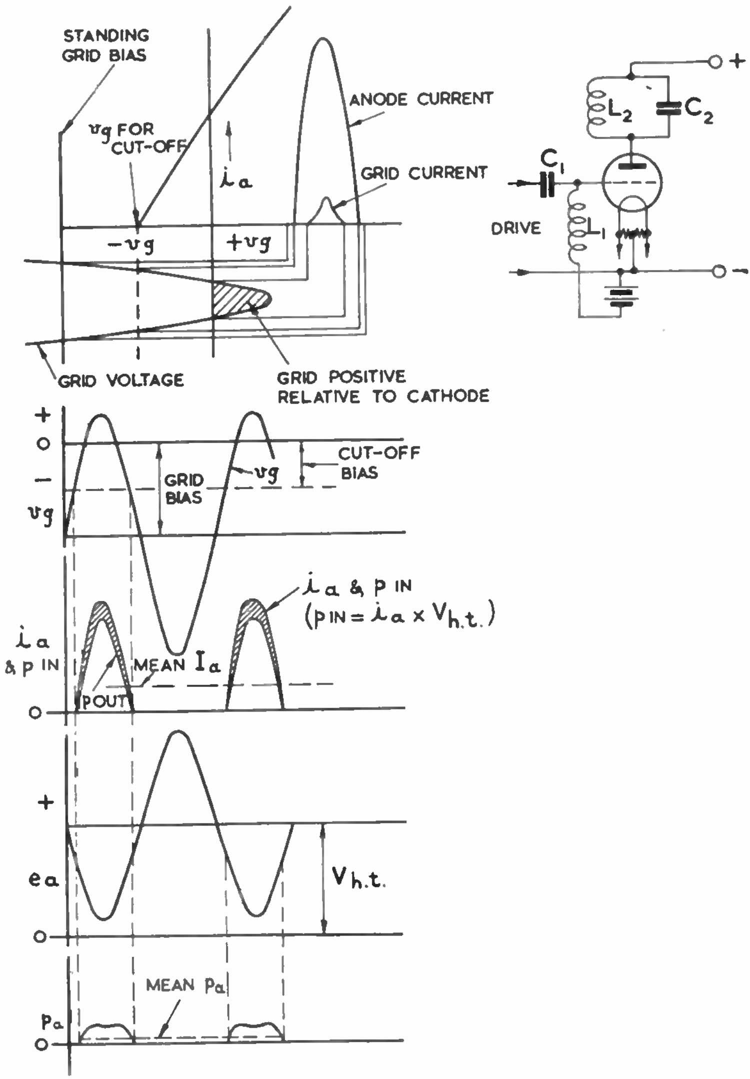


FIG. IV.23.—Class "C" Tuned-anode Amplifier and Waveforms.

circuits restore the sinusoidal waveform and pass only the desired frequency.

4.10. Push–Pull Power Amplifiers

Great advantages accrue from the use of push–pull radio-frequency amplifiers in many respects similar to those obtained in audio-frequency work. An important advantage is the cancellation of second-harmonic distortion. Push–pull amplifiers are commonly used in all but the lowest-powered transmitters because of these advantages.

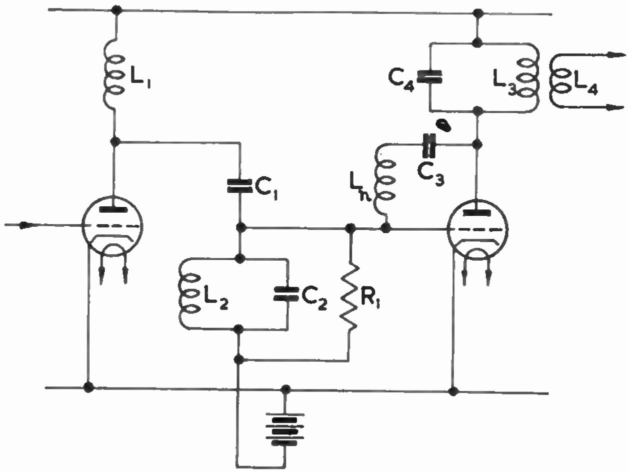


FIG. IV.24.—Showing Parallel Grid Resistor

The circuits are readily neutralised, because of the existence of antiphase points at the anodes of the two valves. The balanced nature of the circuit and good capacitance balance makes the push–pull system easier to operate, especially at the higher frequencies (short wavelengths). The tuned circuits receive power from the valves on each half-cycle, so that there is no production of second or other even harmonics. The second harmonic, being that nearest to the fundamental frequency and the most powerful harmonic, is the most difficult component to eliminate by filtering in a single-sided amplifier.

4.11. Summary of Radio-frequency Power Amplifier Performance

Class "A". Maximum theoretical anode efficiency 50 per cent.; maximum generally obtained, about 40 per cent. Used in low-power stages, *e.g.*, for driving larger valves, also when series modulation is employed (see Volume I, Chapter 9).

Class " B ". Maximum theoretical anode efficiency 78·5 per cent.; maximum generally obtained, about 62 per cent. at full designed output. Output p.d. closely proportional to grid signal p.d., and so is suitable for amplification of modulated carrier. Efficiency with given circuit (anode-load, supply volts, etc.) falls if grid drive and output are reduced so that average efficiency when amplifying a carrier subject to modulation is only one half of 62 per cent.—about 31 per cent.

Class " C ". Maximum practical efficiency, about 85 per cent. Output not proportional to grid input, so cannot be used to amplify a modulated carrier. May be " anode modulated " (see Chapter 7) at high power level. Large value of grid-bias and grid-signal input necessary and high peak grid current usually flows.

In all radio-frequency amplifiers, filament-power consumption decreases the overall efficiency, and economic considerations dictate that valves are usually worked as near to the practical limit as possible. Other losses occur in the tuned circuits, and are made as small as possible. These losses are of the order 5–10 per cent. of the radio-frequency power output, and consist of resistive and eddy current losses in the inductor and a small dielectric loss in the tuning capacitor.

It is instructive to calculate the saving effected by increasing efficiency by a small amount. Consider an amplifier delivering 60 kW (P_0) at an efficiency (η) of 73 per cent. $\left(\dfrac{\text{Ratio of power output}}{\text{H.T. power (final stage)}}\right)$. The H.T. power required (P_i) is obtained from

$$\eta = \frac{P_0}{P_i} \times 100 \text{ per cent.}$$

therefore

$$P_i = \frac{P_0 \times 100}{\eta} = \frac{60 \times 100}{73} = 82{\cdot}2 \text{ kW.}$$

If η is raised to 75 per cent., the output still being 60 kW,

$$P_i = \frac{60 \times 100}{75} = 80 \text{ kW.}$$

In the first case the losses are $82{\cdot}2 - 60 = 22{\cdot}2$ kW, while in the second the losses are $80 - 60 = 20$ kW—a decrease of about 10 per cent. If H.T. power costs 1*d*. per unit and a transmitter operates for 15 hours per day, the saving in running cost is $2{\cdot}2 \times 15 \times 1 = 33d.$ per day, which is equivalent to

$$\frac{33 \times 365}{240} = \text{say £50 per annum.}$$

QUESTIONS

1. A high-power radio-frequency amplifier operates with an H.T. supply at 4,000 volts, and the mean anode current is 500 mA. The radio-frequency circulating current in the tuned circuit is 20 amps., and the effective series resistance is 4 ohms.

Calculate: (*a*) the power dissipated at the anode of the valve; and (*b*) the efficiency of the stage. (C. & G., 1948.)

Answer: (*a*) 400 watts; (*b*) 80 per cent.

2. Calculate the voltage gain between grid and anode of a single-stage intermediate-frequency amplifier operating at 465 kc/s, given that the pentode valve used has a mutual conductance of 5 mA/volt and the anode tuned circuit consists of a 200-μμF capacitor in parallel with an inductor having a Q of 100. (C. & G., 1949.)

Answer: 855.

3. Outline the design and construction of a two-stage tuned radio-frequency amplifier suitable for use at about 5 Mc/s.

If the input power is —40 db relative to 1 watt and the power output is 10 volts across 100 ohms, what is the gain of the amplifier in decibels? (C. & G., 1949.)

Answer: 40 db.

4. What is meant by the Class "C" operation of a radio-frequency amplifier? The final Class "C" amplifier of a transmitter takes a mean anode current of 400 mA at 1,200 volts and delivers a current of 0·75 amps. (r.m.s.) to a load of 600 ohms resistance. What is the efficiency of the stage? (C. & G., 1949.)

Answer: 70 per cent.

5. Give the circuit diagram and typical values for the components of a single-stage 465-kc/s intermediate-frequency amplifier having an approximately uniform response over a frequency band 10 kc/s wide. (C. & C., 1951.)

6. Explain the expression "anode conversion efficiency" or "plate efficiency" used in connexion with a radio-frequency power amplifier. On what factors does this efficiency depend? Include in your answer numerical values of the anode conversion efficiency in some typical cases.

(Brit. I.R.E., Radio Transmission, November 1945.)

7. What is the purpose of: (*a*) neutralising, (*b*) decoupling, as applied to thermionic valve circuits? How are these operations carried out? (C. & G., 1945.)

8. Explain and describe some neutralising methods of producing stability in valve amplifiers. (I.E.E., May 1930.)

9. What are the causes of instability in radio-frequency amplifiers? How are these troubles overcome in practice?

Give a circuit diagram of the radio-frequency circuits of a receiver having two radio-frequency amplifier stages.

(C. & G., 1931.)

10. Why is a high-frequency amplifier consisting of a triode with tuned-grid and tuned anode circuits liable to be unstable? How can such an amplifier be stabilised? (I.E.E., November 1933.)

11. Draw the anode-current/grid potential characteristic of a variable-mu pentode valve and use it to explain one application of this valve in a radio receiver. (C. & G., 1951.)

12. A single-stage thermionic-valve amplifier has a tuned anode circuit having a capacitance of 400 $\mu\mu F$, an inductance of 300 μH and a resistance of 12 ohms. If the plate resistance is 40,000 ohms and the amplification factor is 10, what is the radio-frequency signal p.d. across the tuned output circuit when 0·12 volt is applied to the grid of the valve? (C. & G., 1946.)

Answer : 0·73 volt.

13. The tuned circuit in a tuned-anode amplifier uses a coil having an inductance of 80 μH, and a Q-factor of 85, tuned to a frequency of 1 Mc/s by a parallel loss-free capacitor. Find the stage gain at resonance, and the frequency bandwidth over which the gain exceeds 70·7 per cent. of its resonant value :

(*a*) when using a pentode having a slope resistance of 100,000 ohms and a mutual conductance of 2 mA/volt ;

(*b*) when using a triode having an amplification factor of 10 and a slope resistance of 5,000 ohms.

Answers : (*a*) 60, 16·7 kc/s ; (*b*) 8·95, 112 kc/s.

SPECIMEN ANSWER 1

Q. Outline the design and construction of a two-stage tuned radio-frequency amplifier suitable for use at about 5 Mc/s. (Part question from C. & G., 1949.)

Note : Assume a receiving amplifier is intended.

A. A suitable circuit is shown in Fig. IV.Q.1.

Since a receiving unit is assumed, voltage amplification only is required in the radio-frequency stages, and high-frequency pentodes are used for both valves to avoid instability.

The transformers T_1, T_2 and T_3 are all air-cored and wound on polythene formers. To increase the Q of the coils, heavy gauge conductors are used for the windings, and these conductors may be silver plated, since at 5 Mc/s skin effect begins to be of importance.

All decoupling capacitors should be of the mica dielectric type, since with the paper type dielectric loss and inductance may be appreciable at the operating frequency.

If the tuning of the amplifier is to be variable, the tuning capacitors C_1, C_2 and C_3 will be ganged air dielectric variable capacitors with ceramic trimmers.

If the amplifier is to operate at a fixed frequency, *e.g.*, as the

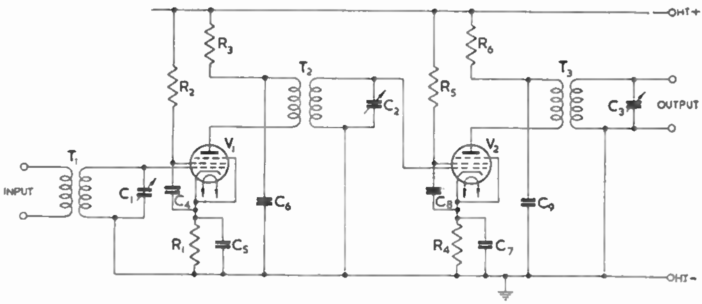


Fig. IV.Q.1

intermediate-frequency amplifier of a television sound receiver, these capacitors will be variable ceramic types.

The screen capacitors C_4 and C_8 are connected directly between screen and cathode rather than between screen and earth in order to improve stability.

In the construction of the amplifier a straight-line principle must be adopted, the components being located in the following order :

Input, T_1, V_1, T_2, V_2, T_3, output.

The two valves are mounted above the chassis, with all other components below. A diagrammatic sketch of the chassis viewed from underneath is shown in Fig. IV.Q.2. in which variable tuning is assumed.

The dotted lines crossing the valve-holders represent screens which extend across the chassis, dividing it into three compartments, the screens being cut away to clear the connecting tags on the valve-holders. Components located in the left-hand compartment are those associated with the input to V_1, viz., T_1, C_1, R_1 and C_5. In the centre compartment are the inter-stage coupling transformer T_2 and its tuning capacitor C_2; the output circuit of V_1, *i.e.*, R_2, R_3, C_4 and C_6; and the input circuit to V_2, R_4 and C_7. The output circuit from V_2 is located in the right-hand compartment, which contains R_5, R_6, C_8 and C_9, as well as T_3 and C_3. The input and output jacks are located as shown, and the power supply connecting strip is to one side of the output compartment. Separate heater leads are run from this strip to the valves V_1 and V_2, and these leads are twisted to reduce the possibility of regenerative feedback via the heater leads. Also care must be taken to avoid couplings caused by earth return currents flowing in the chassis. The earth return leads of each stage should be connected to a common earth point, separate points being used for each stage. At 5 Mc/s, aluminium will be

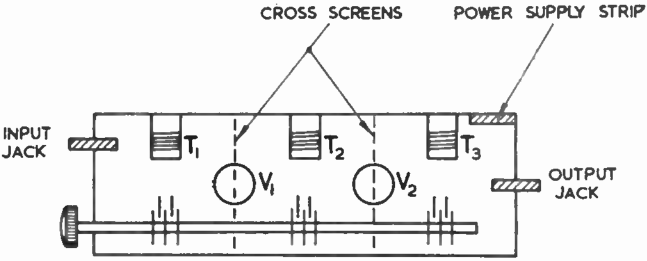


FIG. IV.Q.2

suitable for the chassis and cross-screens, which are earthed and bonded together. (P.O. Eng. Dept.)

SPECIMEN ANSWER 11

Q. The primary and secondary coils of an intermediate-frequency transformer are each tuned to 465 kc/s by 220-pF capacitors, and the coupling is then adjusted to give a bandwidth of 12 kc/s. Assuming that the coefficient of coupling is small, find the mutual inductance of the transformer.

A. Since the same value of capacitance is used to tune both primary and secondary coils, primary and secondary inductances are equal. Thus at resonance, for either primary or secondary coil

$$\omega L = \frac{1}{\omega C}$$

$$L = \frac{1}{\omega^2 C}$$

$$= \frac{10^{12}}{(2\pi \times 465 \times 10^3)^2 \times 220}$$

$$= 5{\cdot}32 \times 10^{-4} \text{ henrys}$$

$$\therefore L = 0{\cdot}532 \text{ mH}$$

When k is small the coupling coefficient is given approximately by

$$k = \frac{\text{Pass band}}{\text{Resonant frequency}}$$

$$\therefore\ k = \frac{12}{465}$$

$$\text{also } k = \frac{M}{\sqrt{L_p L_s}}$$

$$\text{but in this case } L_p = L_s = L$$

$$\therefore\ k = \frac{M}{L}$$

$$\text{and} \quad M = kL$$

$$= \frac{12}{465} \times 0{\cdot}532 \times 10^3 \mu\text{H}$$

$$\therefore\ M = \underline{13{\cdot}7 \mu\text{H}}$$

(P.O. Eng. Dept.)

CHAPTER 5

OSCILLATORS

In this chapter we shall consider in more detail the simple circuits described in Chapter 8 of Volume I, and we shall also describe the crystal oscillator.

5.1. Feedback Oscillators

This class of oscillator consists of a resonant circuit coupled to a valve amplifier so that the valve output supplies the losses occurring in the resonant circuit (Fig. V.1). It is apparent that con-

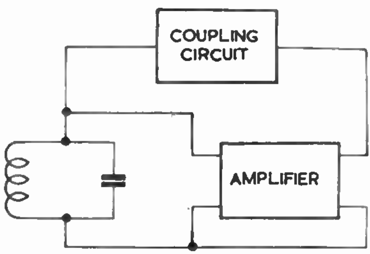


FIG. V.1.—Principle of Oscillator Circuit

tinuous oscillations will occur if the amount of energy fed back to the tuned circuit is at least equal to the energy dissipated in the tuned circuit. Consider the tuned-anode feedback oscillator (Fig. V.2). Here L_1C_1 is the resonant circuit forming the anode load of the valve; L_1 and L_2 are coupled so that an alternating current in L_1 induces an alternating e.m.f. in L_2. If the sense of the induced e.m.f. is correct, then a fall in anode potential causes a rise of potential at the grid end of L_2, so in turn increasing the anode current and causing a further drop in anode potential. Eventually the valve is unable to draw any extra anode current, and C_1 commences to discharge into L_1, so that the cycle is reversed. The circulating current in L_1C_1 causes losses, principally in the resistance of the inductor. The simple circuit of Fig. V.2 can be redrawn including a resistor r to represent these losses and also showing the coupling M between L_1 and L_2 (Fig. V.3). The valve has an effective dynamic mutual conductance g'_m. Let us assume that there is a sinusoidal p.d. V across the tuned circuit and that a sinusoidal current I flows in it. Now at resonance the tuned cir-

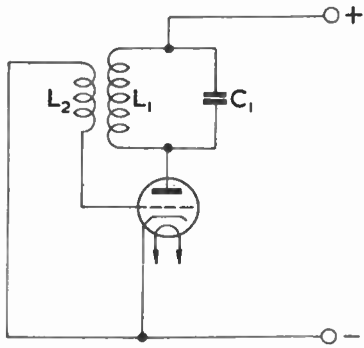


FIG. V.2.—Simple Tuned-anode Oscillator

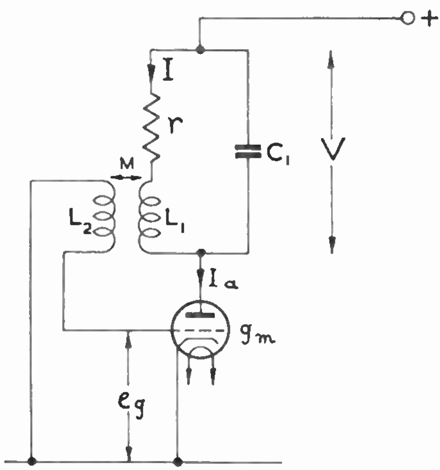


FIG. V.3.—Tuned-anode Oscillator—Analysis

cuit is equivalent to its dynamic resistance $R_D = \frac{L_1}{C_1 r}$, and if I_a is the alternating component of the valve anode current,

$$V = I_a R_D = I_a \frac{L_1}{C_1 r} \text{ when } \frac{\omega}{2\pi} = f = \frac{1}{2\pi\sqrt{L_1 C_1}} \quad . \quad (1)$$

Also $V = I(\sqrt{\omega^2 L_1{}^2 + r^2}) \simeq I\omega L_1$,

since r is small compared with ωL_1 (2)

$$\therefore \quad I\omega L_1 = I_a \frac{L_1}{C_1 r}$$

Re-arranging $$I_a = \omega C_1 r I \qquad (3)$$

Now $$E_g = \omega MI$$

and $$I_a = E_g g'_m$$
$$= \omega MIg'_m \qquad (4)$$

(3) gives the anode current necessary to maintain the p.d. V across the tuned circuit. Therefore the value of I_a from (4) must be at least equal to the value from (3) to maintain oscillation at a frequency determined by

$$f = \frac{1}{2\pi\sqrt{L_1 C_1}}$$

$$\therefore \quad \omega MIg'_m \geqslant \omega C_1 r I$$

$$\therefore \quad M \geqslant \frac{C_1 r}{g'_m}$$

This is known as the maintenance equation for the feedback oscillator.

Another explanation of the operation of the feedback oscillator may be derived from the maintenance equation. In Volume I, Chapter 8, we learnt that an oscillation in a resonant circuit decayed in amplitude because of losses in the components, and we discussed an important type of oscillator making use of the " Negative Resistance " effect obtainable from some tetrode valves. This negative resistance when added to the positive (loss) resistance of the resonant circuit causes continuous oscillations to be set up. The feedback circuit can also be considered as introducing a negative resistance.

The maintenance equation may be re-written

$$r \leqslant \frac{Mg'_m}{C_1}$$

or $$r - \frac{Mg'_m}{C_1} \text{ must be zero or negative}$$

so that the valve and coupling may be considered to introduce a resistance (effectively in series with the inductor L_1) given by $-\frac{Mg'_m}{C_1}$.

If $\frac{Mg'_m}{C_1}$ is numerically less than r, the effective resistance of the resonant circuit is decreased, but not sufficiently for oscillation to occur. This property of the feedback circuit is known as retroaction or reaction, and was at one time widely used to obtain high selectivity by producing circuits having a high effective Q.

5.2. Limitation of Oscillations

The simple circuit of Fig. V.3 does not include any grid bias for the valve, and therefore the grid will become positive with respect

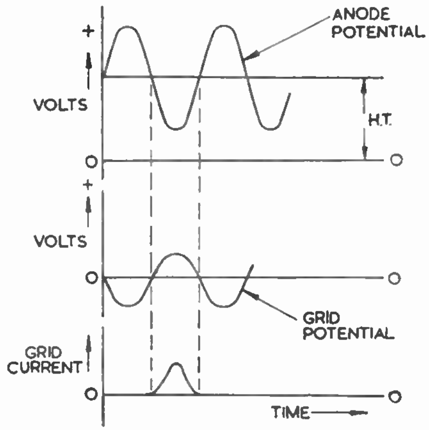


FIG. V.4.—Waveforms in Tuned-anode Oscillator

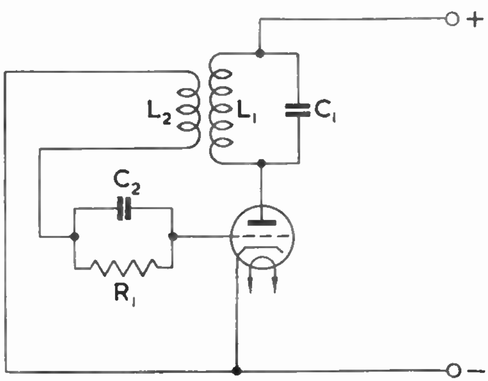


FIG. V.5.—Grid-biassing Circuit

to the cathode during the negative excursion of the anode potential and will draw grid current (Fig. V.4). This grid current is effectively an additional loss in the circuit whose amount increases with the amplitude of oscillation. When this additional loss absorbs the excess energy available as a result of making $M > \frac{C_1 r}{g'_m}$ the amplitude of oscillation settles down to a constant value. The stability of both frequency and amplitude of an oscillator can be much improved by providing grid bias for the valve automatically. Fig. V.5 shows a circuit widely used. C_2 and R_1 apply bias derived from the flow of a small grid current.

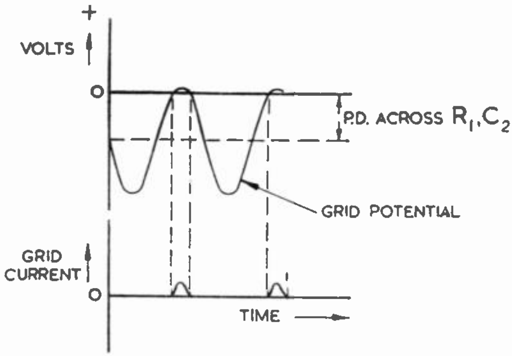


FIG. V.6.—Operation of Grid-bias Circuit

When L_2 drives the grid in a positive direction C_2 becomes charged up by the resulting grid current, and during the rest of the cycle this charge slowly leaks away through R_1. The value of R_1 is chosen to permit a small amount of grid current to flow. Fig. V.6 shows the conditions after the circuit has settled down. This circuit limits the amplitude of the oscillation because the g'_m of the valve is reduced as the grid is made negative with respect to the cathode, and at some value of grid bias the mean value of g'_m is such that the equation $M = \frac{C_1 r}{g'_m}$ is satisfied.

5.3. Tuned-grid Oscillator

Fig. V.7 shows the circuit of this type, the maintenance equation is identical to that for the tuned-anode type, viz., $M \geqslant \frac{C_1 r}{g'_m}$.

5.4. The Hartley Oscillator

Fig. V.8 shows the circuit of a series-fed Hartley oscillator. Here L_1C_1 is the circuit determining the frequency of oscillation.

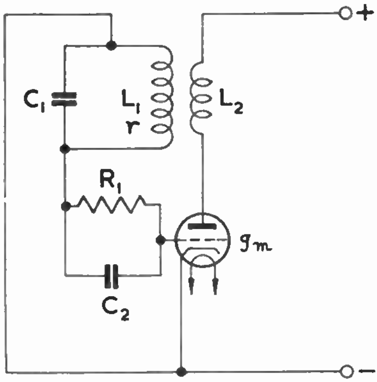


FIG. V.7.—Tuned-grid Circuit

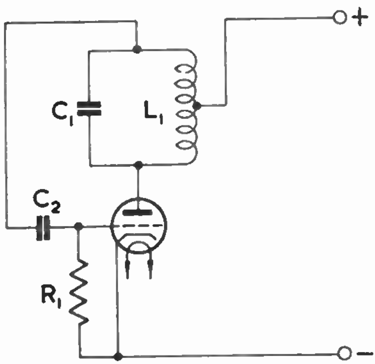


FIG. V.8.—Series-fed Hartley Circuit

C_2 is the grid coupling capacitor, which in conjunction with R_1 provides automatic bias. Oscillation will be maintained, provided that the tap on L_1 is adjusted for the grid p.d. to be at least $\frac{1}{m}$

times the anode p.d., where m is the amplification of the valve under the conditions of operation. The shunt-fed type (Fig. V.9) differs only in that the coil tap is at earth potential. Sometimes

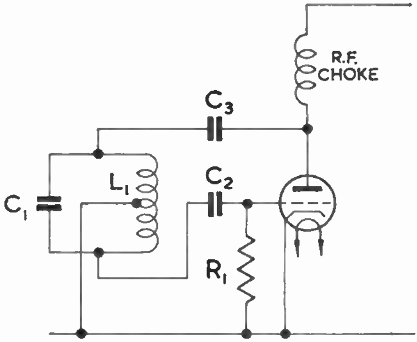


FIG. V.9.—Shunt-fed Hartley Circuit

the radio-frequency choke is replaced by a resistor of suitable value (20,000–100,000 ohms, according to the type of valve and the output required). In both circuits the required phase relation between grid and anode alternating p.d. is automatically obtained. The frequency is given by $f = \frac{1}{2\pi\sqrt{L_1C_1}}$.

5.5. The Colpitts Oscillator

This is similar to the Hartley type, except that a split capacitor is used instead of a tapped coil; Fig. V.10 shows one form. It

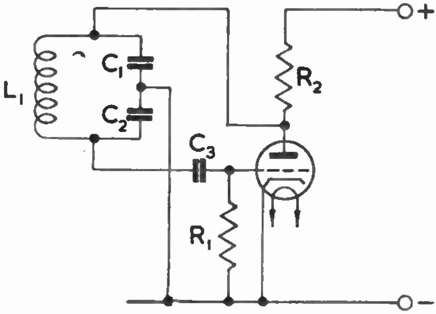


FIG. V.10.—Colpitts Circuit

has the advantage that no tap is required on the coil. The tuning capacitance is the resultant of C_1 and C_2 in series, so that the frequency $f = \dfrac{1}{2\pi\sqrt{L_1 \dfrac{C_1 C_2}{C_1 + C_2}}}$.

5.6. The Tuned-grid Tuned-anode Oscillator

A valve having resonant circuits in both anode and grid will oscillate when conditions are correct because of the coupling via

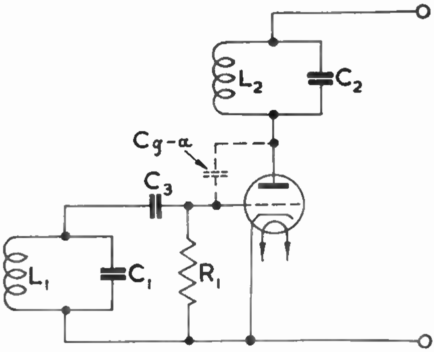


FIG. V.11.—Tuned-grid Tuned-anode Oscillator

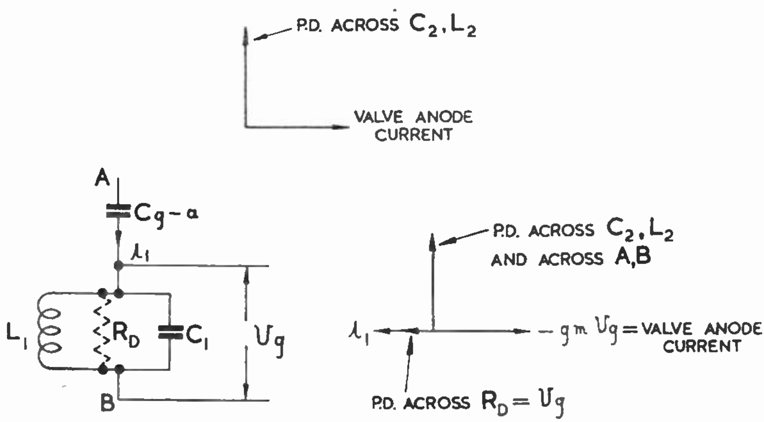


FIG. V.12.—Vector Diagram for Circuit of Fig. V.11

the grid–anode capacitance of the valve. Fig. V.11 shows the circuit. The operating conditions are deduced from the fact that the anode current and grid potential of an amplifier are in anti-phase, and for oscillation to occur the external circuits must reverse the output due to the anode current and apply it to the grid. This condition is satisfied when the resonant frequency of the grid circuit is slightly below the resonant frequency of the tuned-anode circuit as shown in the vector diagram, Fig. V.12. The valve current develops a p.d. leading it by 90° across the anode circuit. This p.d. is then applied to a circuit consisting of the grid–anode capacitance in series with the dynamic resistance of the grid tuned circuit. The reactance of C_{g-a} must be high compared with the dynamic resistance of the tuned-grid circuit for the current here to lead by a further 90°, so that the correct phase relationship is obtained.

5.7. Frequency Stability

The simple theory discussed above does not show that the frequency of oscillation is affected by any circuit values other than those of the components of the resonant circuit. First the valve introduces a capacitance C_{a-k}, which is in parallel with C_1 (Fig.

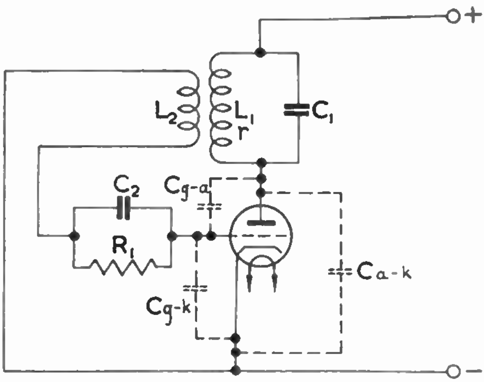


FIG. V.13.—Tuned-anode Oscillator Showing Stray Capacitances

V.13), and so modifies the resonant frequency. Secondly, the resonant frequency is affected by the resistance r representing the losses in the tuned circuit and whose value is also affected by other resistances introduced by the valve. These two effects are minimised by making $\frac{C_1}{L_1}$ as high as convenient so that C_1 is little affected by additional stray capacitances and by making r as small as possible—*i.e.*, by making the tuned circuit of high Q. The resistance introduced by the valve may be made high by using a high-impedance valve and by minimising grid current as shown in Fig. V.5. A resistor R_2 between the valve and the tuned circuit may also be used to reduce the effect of the valve on the frequency of oscillation (Fig. V. 14).

These modifications can ensure a stability of about 10 parts per million due to changes in the valve and supplies. The values of

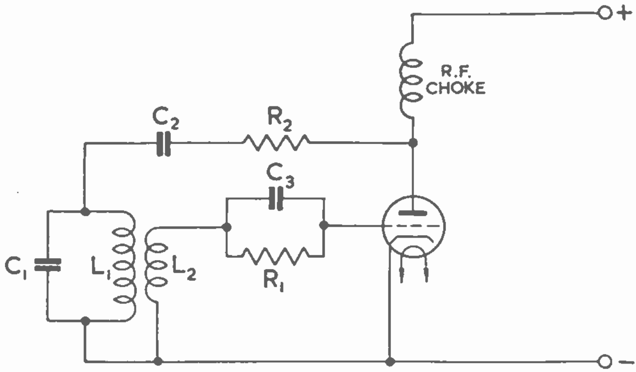


FIG. V.14.—Stabilisation by Resistance

inductance and capacitance in the resonant circuit are likely to vary the frequency by about 25 parts per million per 1° C. change in temperature, and it is therefore necessary to keep the temperature of the components constant to obtain high stability.

In general, simple oscillators may be made sufficiently stable by suitable choice of grid leak to give the right value of bias so that the oscillations are of constant amplitude; an isolating resistor R_2 (Fig. V.14) gives a worthwhile improvement. The effect of R_2 may be calculated as follows:

In Fig. V.14 it is assumed that the impedance of the radio-frequency choke is high compared with the a.c. resistance of the valve, r_a. The gain of the valve, m, is given by $m = \dfrac{\mu(R_2 + R_D)}{r_a + R_2 + R_D}$,

where R_D is the dynamic resistance of the tuned circuit $= \dfrac{L_1}{C_1 r}$.

The p.d. across the tuned circuit, $V = m \times E_g \times \dfrac{R_D}{R_2 + R_D}$

$$\therefore \quad E_g = \frac{V}{m} \times \frac{R_2 + R_D}{R_D} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

Also the p.d. induced in L_2 is given by

$$\frac{V}{\sqrt{\omega^2 L_1^2 + r^2}} \times \omega M \doteqdot \frac{V}{\omega L_1} \times \omega M = \frac{VM}{L_1}$$

(neglecting r in comparison with ωL_1) (2)

For oscillation to occur the induced p.d. given by (2) must be greater than or equal to E_g from (1).

$$\therefore \quad \frac{VM}{L_1} \geqslant \frac{V}{m} \times \frac{R_2 + R_D}{R_D}$$

$$\therefore \quad M \geqslant \frac{L_1}{m} \times \frac{R_2 + R_D}{R_D} \quad . \quad . \quad . \quad . \quad . \quad (3)$$

or

$$M \geqslant \frac{L_1}{m} \times \left(\frac{R_2}{R_D} + 1\right) \quad . \quad . \quad . \quad . \quad . \quad (4)$$

Equation (4) shows that if R_2 is made equal to R_D, M must be double the value required in the absence of R_2 for oscillation to be maintained. As R_D is likely to be greater than r_a, it follows that changes in r_a will be a small part of $r_a + R_2$, which is the impedance "seen" by the resonant circuit, so that its resonant frequency will be little affected by such changes.

From $R_D = \frac{L_1}{C_1 r}$ we have $\frac{L_1}{R_D} = C_1 r$

Also $\frac{m}{R_2 + R_D} = \frac{\mu}{r_a + R_2 + R_D} = g'_m$

Substituting in equation (3) $M \geqslant \frac{L_1}{R_D} \times \frac{R_2 + R_D}{m}$

$\therefore \; M \geqslant C_1 r \times \frac{1}{g'_m}$, which is the well-known maintenance equation.

The effect of the presence of R_2 is to modify g'_m, the dynamic mutual conductance of the valve, so that the maintenance equation is still valid.

The accuracy of setting when a variable capacitor is used to control the frequency in a tunable oscillator is such that further precautions are not warranted; when one or a few fixed frequencies are required temperature control of the oscillator components is warranted.

5.8. The Crystal Oscillator

When certain crystals are subjected to a potential difference between two opposite faces, they distort mechanically. This phenomenon, known as piezo-electricity, is utilised in the crystal oscillator. Any elastic material exhibits mechanical resonance; quartz and Rochelle salt crystals are also piezo-electric. Crystals for oscillators are cut to the form of a thin plate, and sometimes have foils attached to their largest faces to enable a p.d. to be applied. This type is usually mounted by its connecting leads in an evacuated glass envelope like a valve which has pins for connexion to be made. At radio frequencies it is satisfactory to place the crystal between two parallel plates forming a small capacitor,

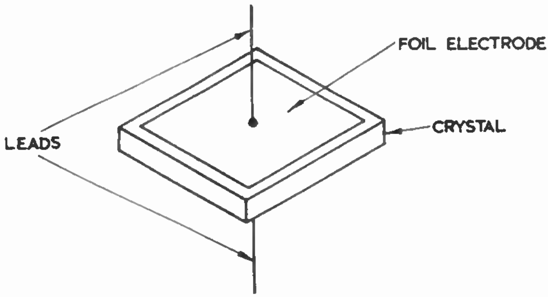


FIG. V.15.—Crystal with Foil Electrodes

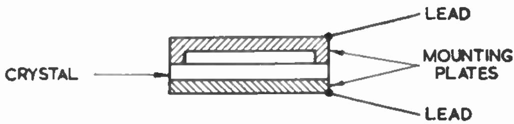


FIG. V.16.—Plate-mounted Crystal

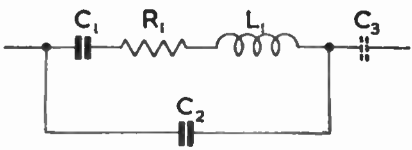


FIG. V.17.—Equivalent Circuit of Crystal

Figs. V.15 and 16. The crystal can be represented electrically by the circuit shown in Fig. V.17. C_1R_1 and L_1 represent the resonant circuit of the crystal, while C_2 is the capacitance of the crystal between the electrodes. C_3 represents the additional series capacitance when the type of mounting in Fig. V.16 is used. The Q value of the crystal, represented by $\frac{\omega L_1}{R_1}$, is very high at resonance, typically 30,000, so that the inherent stability is high. A common circuit known as the Pierce-Miller is shown in Fig. V.18. This functions by feedback through the anode-grid capacitance of the valve. If the tuned circuit is resonant at a frequency slightly higher than the natural frequency of the

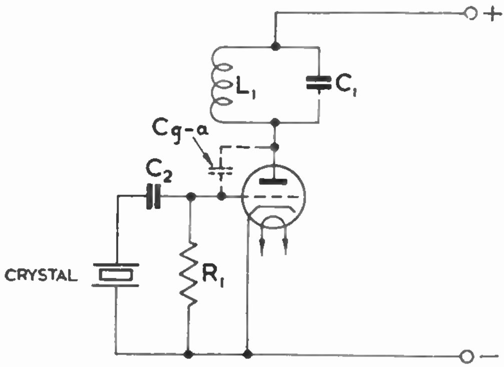


FIG. V.18.—Pierce-Miller Oscillator Circuit

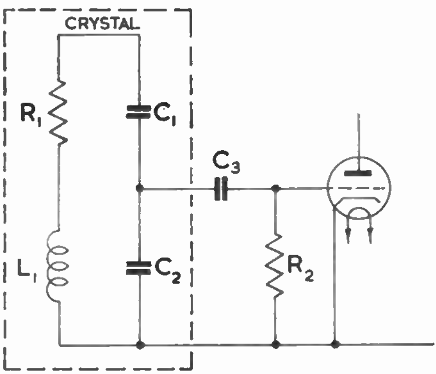


FIG. V.19.—Equivalent Circuit of Crystal in Fig. V.18

crystal, it presents an inductive load to the valve, and the phase relations are correct for oscillation to occur as in the tuned-anode tuned-grid oscillator. The equivalent circuit is shown in Fig. V.19, where the crystal components are designated in the same way as in Fig. V.17. C_2 is much higher in value than C_1, so that the crystal appears as a parallel resonant circuit whose frequency is determined principally by C_1 and L_1, the p.d. across the crystal being greater than the input to the valve, which is effectively tapped down on C_2.

Fig. V.20 shows the circuit of the Pierce crystal oscillator in

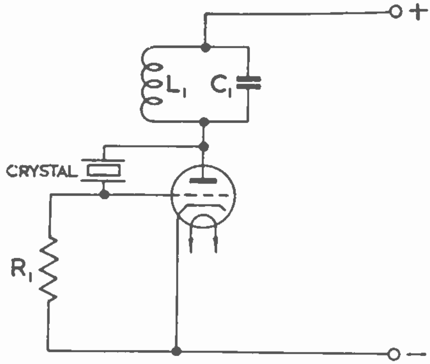


FIG. V.20.—Pierce Crystal Oscillator

which the crystal provides the feedback impedance from anode to grid. Assuming that the crystal is inductive and the tuned circuit is capacitive at the frequency of oscillation, the p.d. across the tuned circuit lags by 90° on the anode current and the p.d. at the grid lags by nearly 90° on the p.d. at the anode. The grid-cathode p.d. and anode current are therefore practically in anti-phase and oscillation is maintained.

5.9. Stability of Crystal Oscillators

The constants, and therefore the resonant frequency, of a crystal depend on its temperature in a manner depending on the way in which the plate is cut from the parent crystal. It is possible to obtain crystals having a very low-temperature coefficient of frequency (*e.g.*, between +20 and —20 parts per million per °C.) by cutting the plate in a certain way, but even so a crystal used to control the carrier frequency of a fixed transmitter is invariably held at a constant temperature. The stability of frequency with temperature is about 20–50 parts per million per °C. in crystals not specially cut for low-temperature coefficient, but because they are superior in respect of purity of output, ordinary cuts held at constant temperature are preferable to special cuts. It is essential not to " over-run " a crystal as breakage may occur if its amplitude of vibration is excessive. The crystal may also rise in temperature because of the energy dissipated in it, and so cause a frequency change which is difficult to control.

5.10. Buffer Stages

In order to prevent the oscillator from being affected by subsequent stages in the equipment, it is usual to take the output from an oscillator via a small capacitor connected to the oscillator anode, or by a weakly coupled secondary or tertiary winding on the inductor and thence to the grid of an amplifying stage operating in Class "A". This valve acts as a "buffer" between the oscillator and the subsequent stages.

Questions

1. Explain the causes of frequency instability in a valve oscillator. Describe how the instability effects can be minimised in the case of: (*a*) an oscillator for use on a single fixed frequency, and (*b*) an oscillator to cover a 3-to-1 frequency range.

Give circuit diagrams of the two oscillators.

(C. & G., Radio II, 1947.)

2. Draw the circuit diagram of a tuned anode triode oscillator, and deduce an equation which must be satisfied in order that oscillations shall be maintained.

3. Why is it necessary to feed back energy from the output of an amplifier to the input in order to maintain continuous oscillations? Explain what factors affect the minimum amount of energy to be fed back, and the amplitude of the resulting oscillations. Draw the circuit diagram of a series-fed, tuned-anode, Hartley oscillator, and explain its action.

4. Describe, with the aid of a circuit diagram, a crystal-controlled master oscillator for a medium-wave broadcast transmitter, stating the precautions necessary to achieve good frequency stability.

(C. & G., 1949.)

5. Give sketches showing one arrangement for the maintenance of electrical oscillations by a triode valve, and discuss mathematically the necessary circuit conditions. (L.U., 1929.)

6. What factors limit the power output and the high-frequency current obtainable from a valve generator? Why is the current in the oscillatory circuit relatively free from higher harmonics?

(I.E.E., May 1931.)

7. Explain how quartz-crystal oscillators can be used in conjunction with a valve to generate high-frequency currents. Discuss the advantages, difficulties and applications of such generators.

(L.U., 1932.)

Specimen Answer

Q. Give the circuit diagram of a variable-frequency oscillator for laboratory use covering the range 1–2 Mc/s and indicate typical values for the components.

State the precautions necessary to obtain good frequency stability, and indicate the order of stability you would expect.

(C. & G., 1952.)

A. Figure V.Q.1 shows the circuit diagram of a Hartley oscillator suitable for laboratory use. The tuning coil has a secondary

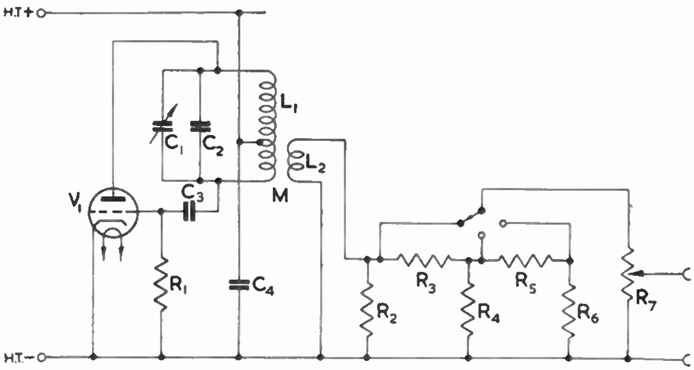


Fig. V.Q.1

$L_1 = 64\ \mu H$	$R_1 = 270\ K\Omega$
$C_1 = 30-330$ pF variable	$R_2 = 68\ \Omega$
$C_2 = 60$ pF mica	$R_3 = 560\ \Omega$
$C_3 = 100$ pF mica	$R_4 = 68\ \Omega$
$C_4 = 1\ \mu F$ paper	$R_5 = 560\ \Omega$
	$R_6 = 68\ \Omega$
	$R_7 = 600\ \Omega$ variable

winding connected to a stepped output control. The values of the resistors used in this control are arranged so that the load on the secondary winding is not affected greatly by the external load connected to the instrument.

The following precautions are necessary for stability of frequency :

(a) *Construction.* The components must be mounted on a rigid structure to prevent relative movement and consequent change of capacitance or inductance. The components of the resonant circuit should be mounted in such positions that they are not subject to the heat from the valve. The whole unit must be mounted in a metal box to prevent interference with or from other equipment.

(b) *Components.* The frequency stability depends chiefly on the stability of the inductor L_1 and the capacitors C_1 and C_2. The inductor L_1 must be wound on a former of rigid construction and of material having a low coefficient of expansion while the capacitors must be of good quality. The wiring between the resonant circuit and the valve must be rigid and as short as possible to reduce the effects of stray capacitance.

(c) *Operating Conditions.* The frequency of a simple oscillator, such as that shown, is affected by changes in the supply p.d.s, and if a high order of stability is required the supplies can be stabilised. The load applied to the secondary winding L_2 affects the frequency. This is reduced by the use of the circuit shown; a further reduction could be obtained if a buffer stage were inserted between L_2 and the output control network R_2, etc.

The order of stability to be expected from such an oscillator at a given setting of the tuning capacitor is 100 parts per million over short periods. The reading accuracy of the dial fitted to the variable capacitor will set a limit to the accuracy of the oscillator, and this is not likely to be better than 0·5 per cent. for a simple mechanism.

CHAPTER 6

POWER SUPPLIES

In this chapter we shall be dealing in more detail with circuits already discussed in Volume I, Chapter 7. In addition, some other H.T. supply circuits are described, and simple stabiliser circuits for improved regulation are considered.

6.1. The Full-wave Bridge Rectifier

The use of metal rectifiers in the form of the familiar " bridge " circuit provides a convenient means of obtaining full-wave recti-

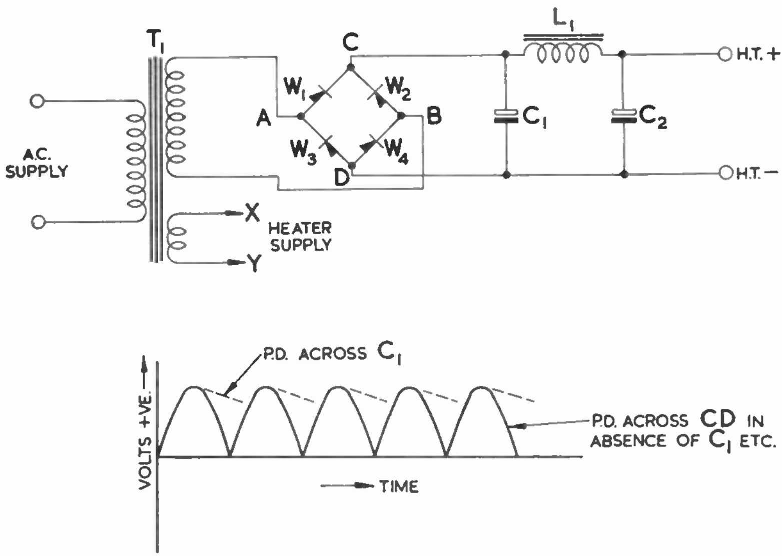


Fig. VI.1—Full-wave Bridge Rectifier and Waveforms

fication. The a.c. is fed to bridge points *AB* via a transformer (Fig. VI.1), and a unidirectional output appears across points *CD*, this being connected to the usual reservoir and filter circuit $C_1C_2L_1$.

Considering the instant when *A* is positive with respect to *B*, then W_1 conducts and current flows through the reservoir capacitor C_1 to W_4, which is also conducting, and so back to point B; rectifiers W_2 and W_3 are then non-conducting. During the next half-cycle the rectifiers W_2 and W_3 are conducting and W_1 and W_4 are non-conducting, the current flowing in the same direction as before into C_1 charges its upper plate positively. The capacitor C_1 discharges through the filter circuit L_1C_2 when a load is connected across the output terminals. The secondary winding XY provides an a.c. supply for the valve heaters.

6.2. The Voltage Doubler Circuit

Where it is necessary to provide a high direct potential, a voltage doubler circuit may be used which will give a direct output

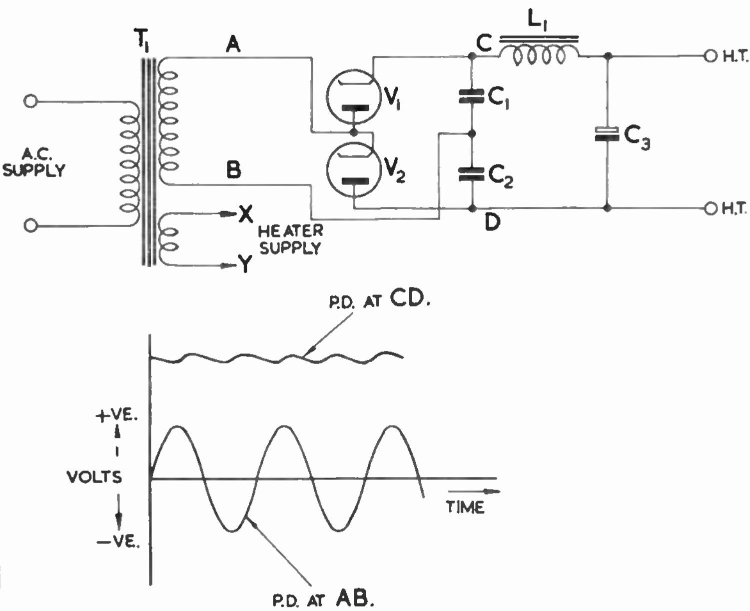


FIG. VI.2.—Voltage Doubler Circuit (Valve Heaters Omitted)

approaching twice the peak value of the alternating input. The circuit is shown in Fig. VI.2, and when point A is positive with respect to point B, diode V_1 conducts and charges capacitor C_1; during the next half-cycle, diode V_2 conducts and charges capacitor

C_2; since the two capacitors are in series, the resultant p.d. is equal to the algebraic sum of the individual p.d.s, and across the two capacitors is a direct potential which is fed to the load via the filter circuit L_1 and C_3.

It will be noticed that the charging current through the secondary of the transformer flows in opposite directions every half-cycle, so that there is no resultant direct magnetisation of the core of the transformer. Double-diode valves are available for voltage doubler working having both diodes contained in a single glass envelope. Also metal rectifiers are available in suitable " stacks " for voltage-doubler working. In the case of the valve heater-cathode potential is high if the heaters are connected in parallel, alternatively, adequate insulation between separate heater windings must be provided in the transformer.

6.3. The Centre-tapped Circuit

Fig. VI.3 shows a bi-phase rectifier circuit employing a centre-tapped transformer. The operation of this circuit is well known, and is described fully in Volume I, Chapter 7.

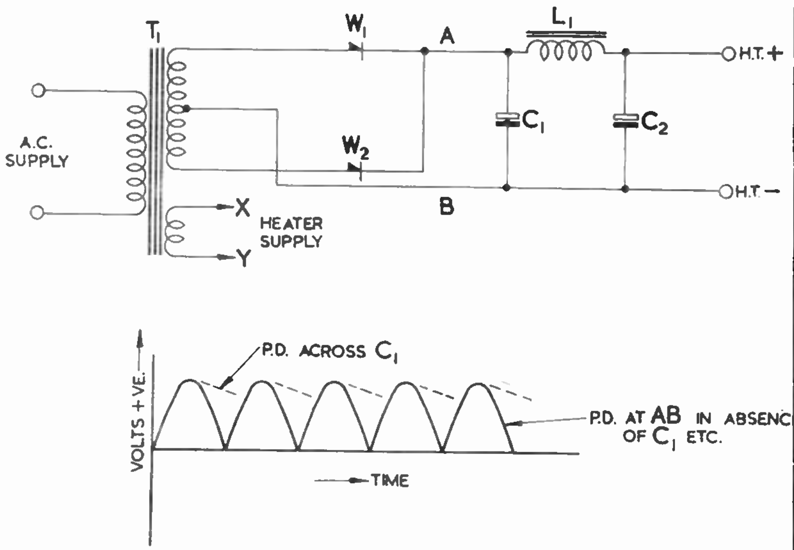


FIG. VI.3—Bi-phase Rectifier Circuit

6.4. Power Supplies for Alternating or Direct Current Working

Equipment for a.c./d.c. working has certain distinguishing characteristics summarised as follows :

1. No transformer (since it is required to work from a d.c. source).
2. Half-wave rectification.
3. Series connexion of valve heaters.

Since no transformer is used, there is direct connexion of the mains to the equipment. It is possible that the mains leads will act as an aerial and be carrying some radio frequency, which may cause interference in the receiver fed by the supply unit. This interference is prevented by a simple filter, comprising a radio-frequency choke in each input lead to the supply unit shunted by a capacitor whose reactance at radio-frequency is very small (Fig. VI.4(*a*)).

Radio-frequency signals may be present in the power supply due not only to the input mains leads, but also to the receiver itself. Lack of suitable screening or decoupling may cause radio-frequency signals to be modulated at supply frequency by the non-linear action of the rectifier and passed back through the receiver to be heard from the loudspeaker as an objectionable hum. This may be minimised by connecting a capacitor (C_4) whose reactance is small at radio-frequency in parallel withthe rectifier, thus preventing modulation taking place. The " raw " output from the rectifier is shown in Fig. VI.4(*b*). When working on d.c. mains the rectifier acts as a low series resistor ; it is, of course, necessary to connect the supply with the correct polarity to obtain H.T.

The order of series connexion of heaters in a receiver is important, and to reduce the amount of hum to a minimum, when the equipment is operating from a.c. mains, the valves handling the smallest signals are connected nearest to H.T. negative so that there is the minimum potential difference between these heaters and H.T. negative. A typical arrangement for a superheterodyne receiver would be as shown in Fig. VI.4(*c*). Other features of Fig. VI.4(*a*) not already mentioned are the resistor R_1, which is a series or " dropping " resistor to determine the correct current for operation of the heater chain ; R_2 is included in series with the rectifier to limit the peak current in the rectifier, and is specified by the valve manufacturer (usually 50–100 ohms) ; C_3 is a small-value capacitor to earth H.T. negative as far as radio frequency is concerned, but to prevent the possibility of earthing the high-potential side of the mains, with consequent undesirable results.

Because the H.T. negative line (usually chassis) may be at some considerable p.d. from earth, certain precautions are necessary when operating such equipment. It is essential for a.c./d.c equipment to be enclosed either in a case of non-conducting material, with all screws which make contact with the chassis covered with wax, or in a properly earthed metal case, the chassis being insulated from it, so that the operator is not able to touch any live parts.

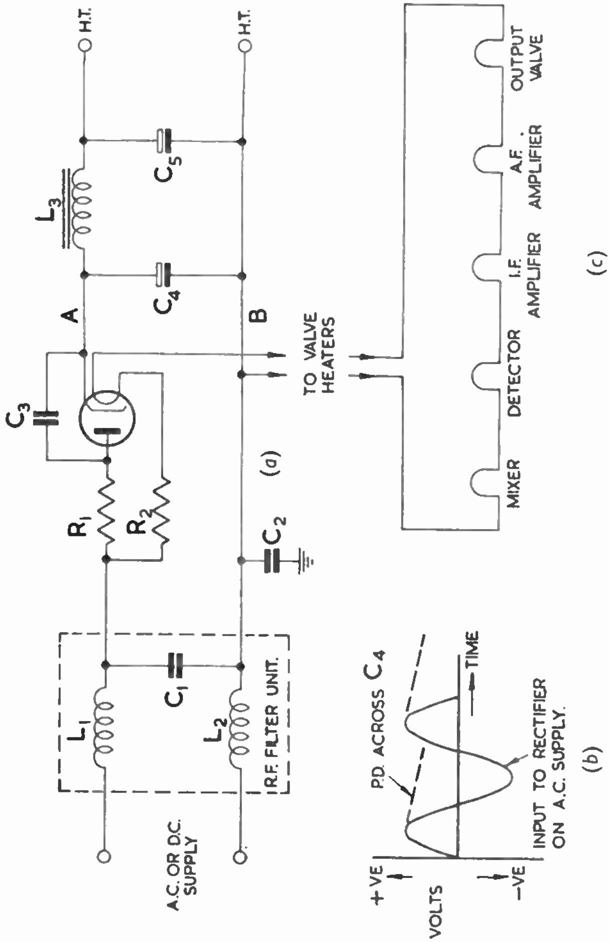


FIG. VI.4.—" Universal " or a.c./d.c. Supply Circuit

6.5. Advantages and Disadvantages of Various Types of Power Supply

Unless it is unavoidable (*e.g.*, a.c./d.c. apparatus), half-wave rectification is seldom used for supplies to broadcast receivers and associated equipment. It suffers from two serious disadvantages :

1. The ripple is at supply frequency which increases the cost of the filter circuit compared with the full-wave or bi-phase case (see below for more details).
2. If a transformer is used the direct load current flows through the secondary winding resulting in direct core magnetisation.

With bi-phase rectification there is no resultant core magnetisation because the load current flows equally and in opposition in the two halves of the secondary winding. Also the ripple frequency is twice the mains frequency, resulting in a less costly filter. The full-wave bridge rectifier offers the same advantages. A centre-tapped transformer as used in the bi-phase rectifier system is called upon to deliver twice the peak value of alternating p.d. as compared with a transformer for half-wave working. However, since there is no resultant direct core magnetisation with centre-tapped working, the transformer core area may be reduced for a given d.c. output. Another important fact is that the single valve in the single-phase half-wave circuit must be larger than each of the two valves in the bi-phase circuit for the same current output, because the mean output is divided between the two valves in the latter case. The advantage of the half-wave rectifier is, of course, its cheapness and simplicity.

Considering now the full-wave bridge rectifier employing four metal rectifier units, shown in Fig. VI.1, this method has the advantages of a simple transformer having only one secondary H.T. winding. There is no resultant core magnetisation, and the ripple frequency is at twice supply frequency. These advantages, however, are obtained at the expense of an increased internal impedance of the source due to two rectifiers in series being in circuit at any instant. Electronic valves are seldom used in this connexion in low-power supplies owing to the complication of extra heater windings. Bridge rectification is widely used in high-power circuits and in meter circuits, battery charging, etc., using metal rectifiers. Valves are used in bridge circuits where high potentials are required, but in receivers diodes or metal rectifiers in the conventional centre-tapped circuit shown in Fig. VI.3 are most common, suitable transformers and valves being readily available.

6.6. Simple Stabilised Power Supplies

Much electronic apparatus in use today, particularly instrument and television transmission equipment, requires a constant value of H.T. potential. This value must be maintained to close limits despite variations in supply p.d. and load current. In order to meet these requirements, some "stabilisation" of the power supply must be effected (Fig. VI.5).

Heater supplies are not normally stabilised, as they are generally working at a fixed load, and furthermore valve heaters of the oxide-coated type are designed to operate over a range of ± 7 per cent. of the nominal rating, thereby rendering stabilisation unnecessary.

It has already been seen in Chapter 2 that neon tubes have the

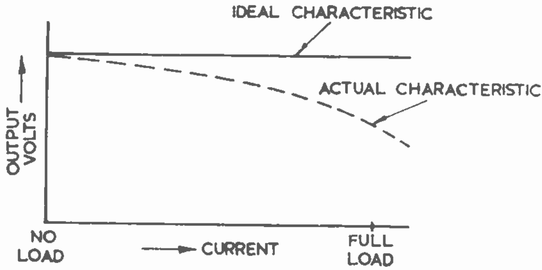


FIG. VI.5.—Load Characteristics of Power Supplies

property of striking at a certain critical p.d. and maintaining a constant p.d. across their terminals over a wide range of current through the tube. This property is used to provide some stabilisation of the H.T. potential in power supplies. Reference to Fig. VI.5 shows that as the load current taken from the power supply increases, the output potential falls. This undesirable variation occurs if the load is variable, and Fig. VI.6 shows a way of reducing the change in H.T. potential with load and approaching the ideal characteristic shown in Fig. VI.5. The neon tube with a resistor R_1 in series with it is placed across the H.T. output terminals of the power supply. The resistor R_1 is chosen so that the current taken by the tube cannot exceed the maximum value stated by the manufacturers and so cause damage to it. As the current through the load varies, so the current through the neon will vary inversely, but providing the neon remains "struck", its action is to maintain a constant potential across its terminals and hence across the output terminals. The current taken by a typical neon tube stabilising a 150-volt 40-mA H.T. supply is 45 mA, and this type of stabilisation can hold the output p.d. within ± 3 per cent. of the nominal from no-load to full-load current.

A varying load current taken from the supply causes changing

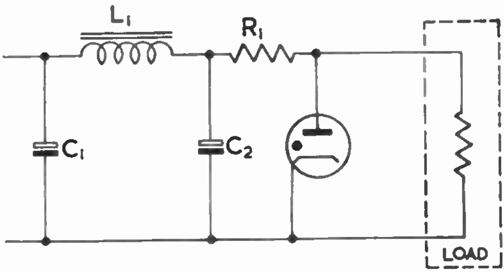


FIG. VI.6.—Simple Stabilised Supply

losses in the components of the power supply so that variations of p.d. at the output terminals occur. Also variations of supply p.d. produce corresponding variations of output p.d. at a given load current; provided that the neon tube is operated within its limits, it gives stabilisation against variations from both causes.

6.7. Regulation of Power Supply

Variations of output potential are compared with the output obtained with no load connected, and the ratio is termed the *regulation* of the power supply (Volume I, Chapter 7), *i.e.*,

$$\text{Regulation }\% = \frac{\text{No-load terminal p.d.} - \text{Full-load terminal p.d.}}{\text{No-load terminal p.d.}} \times 100$$

6.8. The Filter Circuit

There are two main types of filter circuit with which we are concerned, namely filters having either:

1. Shunt capacitor input.
2. Series inductor or choke input (not used on single-phase half-wave rectifiers because the mean p.d. is too low).

6.9. The Shunt-capacitor Input Filter

In this arrangement the rectifier is connected directly to a reservoir capacitor C_1 (Fig. VI.7), which is followed by a smoothing choke L_1 and capacitor C_2. When such an arrangement is first switched on, the charge on C_1 is zero, and during successive half-cycles the two halves of the rectifier pass a current which charges C_1. Current will only flow through the rectifiers when the secondary

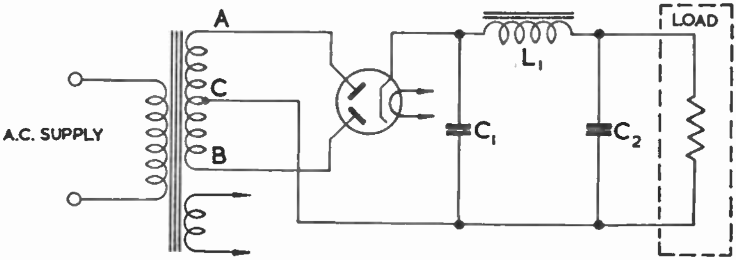


FIG. VI.7.—Capacitor-input Filter Circuit

e.m.f. exceeds the p.d. across C_1; as this p.d rises, so an increasing current will flow through L_1 to the load. During the period of the input cycle, when neither rectifier is conducting (*i.e.*, as the input wave passes through zero), the charge on C_1 falls because the load

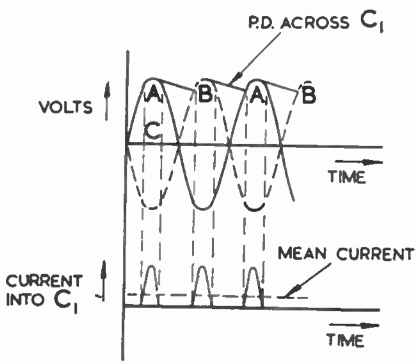


FIG. VI.8.—Waveforms for Fig. VI.7

takes current, but this charge is replaced each half-cycle during the conducting period of the rectifiers. The waveforms of current and p.d. are shown in Fig. VI.8.

During the conducting period the current from the rectifier is determined by the difference between the secondary e.m.f. and the p.d. across C_1 and by the total resistance of the circuit, that is, transformer secondary and valve resistances. If these resistances are low, the p.d. across C_1 will approach the peak value of the transformer e.m.f. and if C_1 is large, this value may be well maintained during discharge. Such conditions require a high peak current in the rectifier and transformer, which may be ten times the mean d.c. output. It is usually necessary to increase the circuit resistance artificially to limit the peak current, and valve makers specify the minimum circuit resistance and maximum value of reservoir capacitance. Normal working conditions are usually chosen to give a mean p.d. at full load of 1–1·2 × the r.m.s. input volts.

When the load current falls, the direct p.d. rises, because the charge on C_1 falls at a lower rate; with no load the p.d. will equal the peak value of the input $= \sqrt{2} \times$ r.m.s. volts $= 1{\cdot}414 \times$ r.m.s. volts.

6.10. Smoothing Circuit

The reservoir capacitor C_1 is followed by the filter L_1C_2, whose function is to filter out the ripple p.d. across C_1.

The effect of inductance is to oppose changes of current and to produce an e.m.f. to oppose such changes. This property is made use of to reduce the ripple at the output of the supply unit. To obtain adequate smoothing the *reactance* (ωL) of the choke to the unwanted ripple frequency should be as high as possible, and the reactance of the smoothing capacitor C_2, $\left(\frac{1}{\omega C_2}\right)$, should be as low as possible so that the ripple-frequency p.d. across it is also low.

The ripple p.d. across C_2 (Fig. VI.7) can be calculated as follows. Let us assume that the ripple at the input to L_1 is E_r volts, and that the ripple current flows through L_1 and C_2 in series, none flowing through the load R.

The impedance of L_1 and C_2 in series is

$$j\omega L_1 - \frac{j}{\omega C_2}, \text{ where } \omega = 2\pi \times \text{ripple frequency}$$

The ripple current is then

$$I_r = \frac{E_r}{j\omega L_1 - \dfrac{j}{\omega C_2}}$$

and the ripple p.d. across C_2 is given by $I_r \times \dfrac{j}{\omega C_2}$.

The ratio of $\frac{\text{ripple p.d. across } C_2}{E_r}$ is then equal to

$$\frac{I_r \times \frac{j}{\omega C_2}}{E_r} = \frac{\frac{E_r}{j\omega L_1 - \frac{j}{\omega C_2}} \times \frac{j}{\omega C_2}}{E_r}$$

$$= \frac{\frac{j}{\omega C_2}}{j\omega L_1 - \frac{j}{\omega C_2}}$$

$$= \frac{1}{\omega^2 L_1 C_2 - 1}$$

(multiplying numerator and denominator by $\frac{\omega C}{j}$).

Therefore the filter has reduced the ripple by the factor $\frac{1}{\omega^2 L_1 C_2 - 1}$ or in general, the ratio

$$\frac{\text{Alternating p.d. across load}}{\text{Alternating p.d. at filter input}} = \frac{1}{\omega^2 LC - 1}$$

ω being that value appropriate to the ripple frequency.

6.11. The Choke Input Circuit

Fig. VI.9 shows a filter having a series inductor input followed by a smoothing capacitor C_1 feeding into a resistive load.

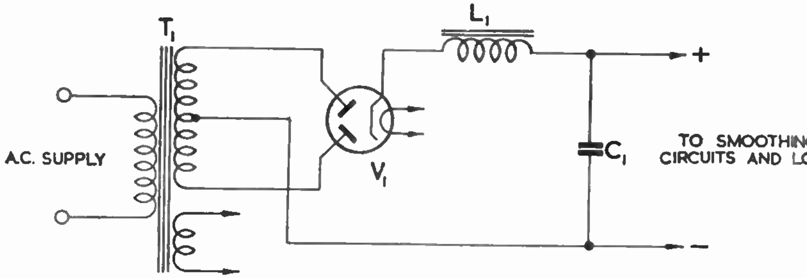


FIG. VI.9.—Choke Input Circuit

The object of using choke input is to obtain continuous conduction in the rectifier and so cause the output p.d. to approximate to the mean value of the rectified pulses. The iron-cored choke L_1 in Fig. VI.9 is chosen large enough in value to maintain a continuous current from the rectifier. Fig. VI.10 shows the form of output obtained with various values of inductance, L_1. It can be seen that too low a value of L_1 produces a discontinuous current through the filter, and by increasing the value of L_1, the variation of current or ripple is reduced.

The mean value of the current through the inductor is, of course, the d.c. output, and the resultant current is the d.c. component added to the ripple current in the inductor. This ripple current (assuming that there is negligible ripple p.d. across the smoothing capacitor C_1) is equal to the fundamental ripple frequency p.d. at the rectifier divided by the reactance of the inductor at ripple

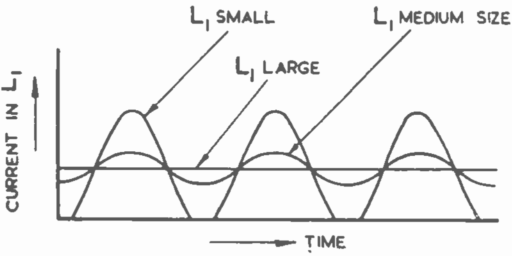


FIG. VI.10.—Effect of Value of Input Inductance

frequency. To ensure that the current in the rectifier is continuous (*i.e.*, does not fall to zero) the peak value of the ripple current must be less than the mean d.c. component.

In a centre-tapped or bridge rectifier the peak value of the ripple-frequency p.d. at the rectifier output is 0·667 of the mean output p.d. The peak value of ripple current in the inductor is therefore $\frac{0{\cdot}667 \times \text{mean output p.d.}}{\omega L_1}$, where ω is $2\pi \times$ fundamental ripple frequency. The mean value of the current is given by $\frac{\text{mean output p.d.}}{\text{total circuit resistance}}$. The limiting case of continuous conduction occurs when the peak ripple current is equal to the mean current, that is, when

$$\frac{0{\cdot}667 \times \text{mean output p.d.}}{\omega L_1} = \frac{\text{mean output p.d.}}{\text{total circuit resistance } (R_T)}$$

$$\therefore \quad L_1 = \frac{0{\cdot}667 R_T}{\omega} \text{ henrys}$$

On a 50-c/s supply with full-wave or bi-phase rectification $\omega = 2\pi \times 100$

$$\therefore \quad L_1 = \frac{0{\cdot}667 R_T}{628} = \frac{R_T}{940} \text{ henrys}$$

Any greater value of L_1 may be used, and the variation of current in the rectifier is then reduced.

Example

It is desired to design a rectifier filter having a series inductor input to operate from a 50-c/s power supply and to deliver 350 volts d.c. at 120 mA, with not more than 5 per cent. ripple. The d.c. resistance of the choke may be neglected.

$$R_T = \text{load resistance} = \frac{350 \times 10^3}{120} = 2{,}920 \text{ ohms}$$

$$\therefore \quad L = \frac{2{,}920}{940} = 3{\cdot}1 \text{ say } 3{\cdot}5 \text{ henrys (since } 3{\cdot}1 \text{ is a minimum value).}$$

For 5 per cent. ripple $\dfrac{1}{\omega^2 LC - 1} = 0{\cdot}05$,

$$\therefore \quad \frac{1}{(\omega^2 \times 3{\cdot}5 \times C) - 1} = 0{\cdot}05$$

or

$$0{\cdot}05\,(\omega^2 \times 3{\cdot}5 \times C - 1) = 1$$

$$\omega^2 \times 3{\cdot}5 \times C - 1 = \frac{1}{0{\cdot}05} = 20$$

$$\therefore \quad \omega^2 \times 3{\cdot}5 \times C = 21$$

$$\therefore \quad C = \frac{21 \times 10^6}{\omega^2 \times 3{\cdot}5}\,\mu F = \frac{21 \times 10^6}{40 \times 0{\cdot}05 \times 3{\cdot}5} \times 10^4 = \frac{10^2}{6}$$

$$\therefore \quad C = 16{\cdot}7\,\mu F$$

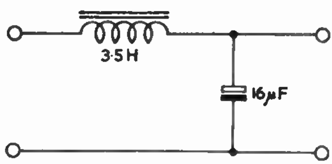


FIG. VI.11.—Numerical Example

(In practise a 16-μF capacitor would be used as the next standard value, 32 μF, would give an unnecessarily low ripple.)

The filter is therefore as shown in Fig. VI.11.

6.12. Effect of Variation of Load on Series Inductor Input Filter

With practical inductors variation of d.c. through them causes variation of inductance (Fig. VI.12). If therefore the load current varies with a rectifier employing a series inductor input filter, $\frac{\omega L}{R_T}$ may become less than 0·667 at low loads (high R_T). This will cause the current to become discontinuous, and the direct output p.d. will rise and approach the peak value

This effect may be reduced by connecting a " bleed " resistor across the rectifier output so that the current drawn by the resistor is an appreciable amount of the total load current and ensures that conduction is continuous.

This remedy, however, requires a larger power supply and choke

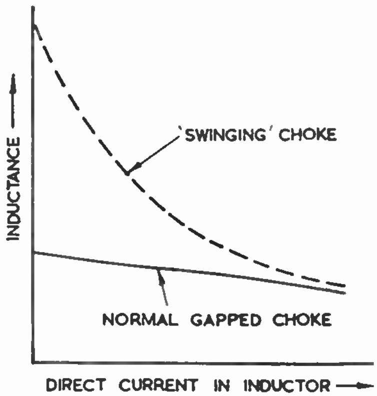


FIG. VI.12.—Characteristics of Typical Chokes

to supply the larger total current (load plus bleeder) and to maintain the same value of ripple volts. Larger components mean more cost, so that the use of a bleeder resistor is not economical. The variation of inductance with direct current may be controlled by means of the air gap (Chapter 4, Volume I) in the core of the inductor. With a small air gap, large values of current through the inductor produce small variations of inductance (Fig. VI.12), but at low values of current the rate of change of inductance with current is large. Therefore the inductor may be used over a wide range of current to obtain continuous conduction in the rectifier by keeping $\frac{\omega L}{R_T} \geqslant 0{\cdot}667$. Such a choke is called a " swinging choke ".

6.13. Comparison between Capacitor Input and Inductor Input Filters

The capacitor input circuit enables the direct output p.d. to be equal to or greater than the r.m.s. value of the alternating input.

As the p.d. across the reservoir capacitor has a smaller ripple component than the input in Fig. VI.9, the choke may be smaller for the same output ripple than when choke input is employed. The r.m.s. current in the transformer winding is higher, and the regulation is higher than in a well-designed choke–input system. For a steady load, such as that of a radio receiver, however, the capacitor input is satisfactory, and it is essential when using single-phase half-wave rectification, as in a.c./d.c. equipment, in order to obtain an output p.d. of the same order as the input.

With the choke–input circuit the direct p.d. is approximately equal to the average value ($=0{\cdot}9 \times$ r.m.s.) of the applied p.d. The r.m.s. value of the transformer secondary current is less than with capacitor input, and although a higher p.d. is required, the transformer is usually smaller for a given output. If continuous conduction is maintained, the regulation is small and is mainly due to resistive drops in components.

6.14. Mercury-vapour Rectifiers

The low impedance and constant p.d. across a conducting mercury-vapour tube (see Chapter 2) make this type of valve ideal where good regulation is required and high currents are involved. It is necessary to allow time for the valves to warm up before H.T. is applied to the anodes; a delay of 30 seconds to 5 minutes is necessary, increasing with valve rating. A small power pack uses two valves in the centre-tap circuit, and it is instructive to calculate the regulation. Fig. VI.13 shows the circuit.

Example

Calculate the output p.d. at 50 and 250 mA for the circuit of Fig. VI.13, assuming that the transformer delivers 500–0–500 volts, the secondary resistance is 150 ohms, and each choke has a resistance of 30 ohms. Assume a fall in potential of 18 volts in the valve during conduction. The smoothing is adequate to maintain continuous conduction in the rectifier. Express the results as a percentage regulation.

The half secondary e.m.f. is 500 volts r.m.s. and the corresponding mean value is $\dfrac{500}{1{\cdot}11} = 451$ volts. Subtracting the constant valve p.d. of 18 volts, the mean input to the choke is $451 - 18 = 433$ volts.

At 50 mA the drop in the circuit resistance, consisting of 75 ohms for a half secondary and 60 ohms for the chokes, is 0·05 (75 + 60) = 6·75 volts, so that the output p.d. at 50 mA is 433 — 6·75 = 426·25, say 426 volts.

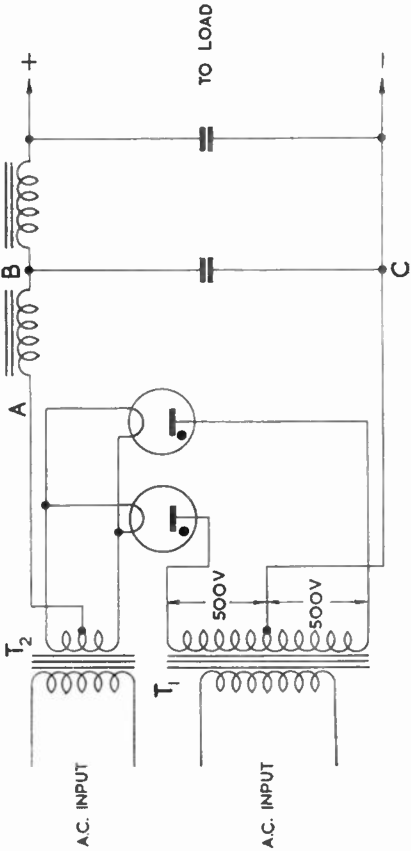


FIG. VI.13.—Bi-phase Rectifier Circuit

Similarly at 250 mA the output p.d. is 433 — 0·25 (75 + 60) = 433 — 33·75 = 399·25 volts, say 399 volts.

The regulation between 50 and 250 mA is therefore

$$\frac{426 - 399}{426} \times 100 \text{ per cent.} = 6{\cdot}35 \text{ per cent.}$$

6.15. Vibrator Power Supplies

These were briefly discussed in Volume I, Chapter 7. They are used to obtain H.T. supplies from d.c. sources, and in principle

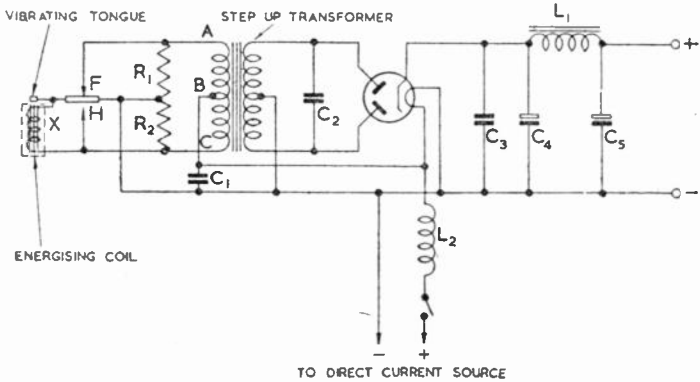


FIG. VI.14.—Vibrator H.T. Supply with Valve Rectifier

comprise a means of operating a transformer from a direct supply. Fig. VI.14 shows the circuit of a vibrator H.T. supply using a valve rectifier. The vibrator causes successive pulses of current

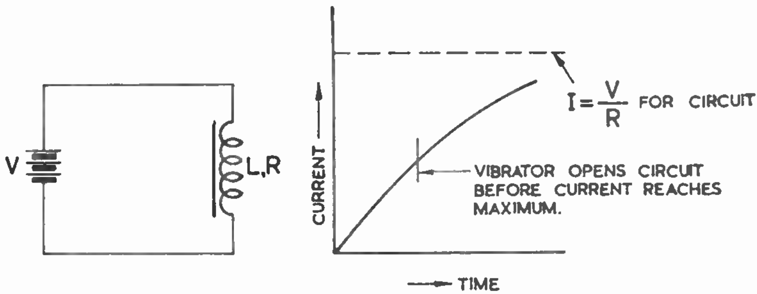


FIG. VI.15.—Circuit and Current—Primary Winding of Vibrator-fed Transformer

to flow in opposite directions through the two halves of the centre-tapped transformer primary winding.

Because there is a constant p.d. across either half of the primary while it is connected to the source, the current will rise during

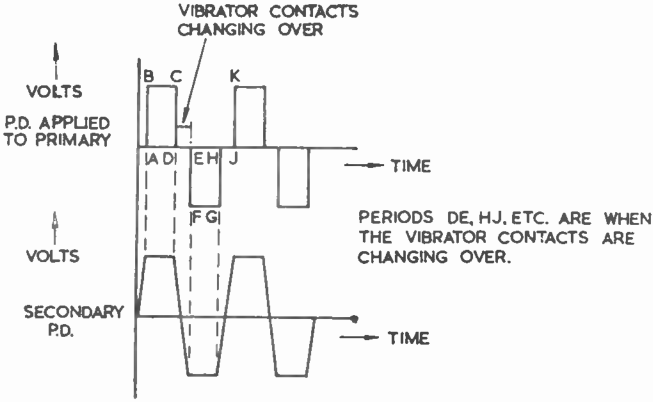


Fig. VI.16.—Waveforms in Vibrator Circuits

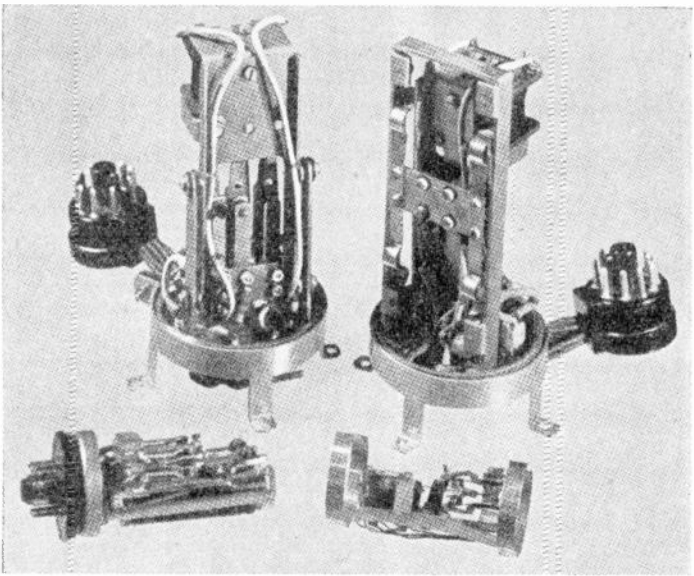


Fig. VI.17.—Typical Vibrator Units—Covers Removed connexion, the circuit being that of an inductor connected to a direct supply. Fig. VI.15 shows the conditions. The e.m.f. in the secondary will be of similar form to the primary p.d., viz., a series of rectangular pulses—Fig. VI.16. The rectifier is a normal centre-tapped circuit. Fig. VI.17. shows the construction of a

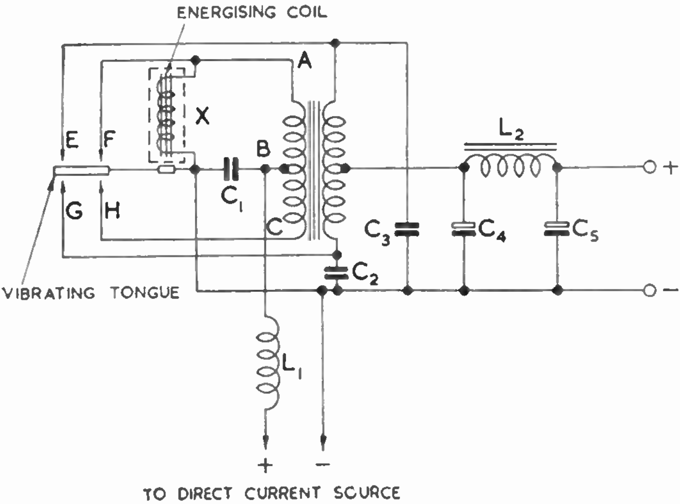


FIG. VI.18.—Self-rectifying Vibrator Unit

typical vibrator. The frequency of vibration is typically 100–120 c/s, and the transformer must be designed for the special conditions of operation.

The "synchronous" vibrator contains additional contacts which are connected in place of a valve to rectify the transformer output (Fig. VI.18). These are economical in space, but the conditions of operation for the high-voltage contacts are rigorous, and adjustment of the contacts is critical.

6.16. Choice of Power-supply Circuit

The arrangement for a particular purpose depends to some extent on the availability of rectifier valves and components. Metal rectifiers can be readily built for any required output, and they or single diodes are commonly used in a.c./d.c. circuits up to 250 mA output. Double-diode rectifiers for use in a centre-tap circuit are available for d.c. outputs up to 250 mA at 500 volts, and this circuit is therefore suitable for all types of receiver operated from a.c. mains.

High-power equipment using hard diodes or mercury-vapour diodes in centre-tap or bridge circuits may be built to supply any amount of power required by electronic equipment. A pair of the smallest mercury-vapour valves used in a centre-tap circuit may deliver 0·5 amp at 1,000–1,500 volts, and four such valves in full-wave bridge will deliver 0·5 amp. at 2,000–2,500 volts. The mercury-vapour valve is most useful at the higher current ratings because of the high efficiency obtainable.

Vibrator supplies are essential when only d.c. sources are obtainable, *e.g.*, mobile equipment for operation from vehicle or aircraft batteries of 6, 12 or 24 volts, and special vibrator equipment has been used on d.c. supplies up to 250 volts for providing an alternating output. In such cases, the vibrator frequency is 50 c/s, so that standard equipment designed for a.c. supplies may be used.

Questions

1. Sketch and explain a skeleton circuit diagram of the H.T. and heater power supplies used in a typical a.c./d.c. superhet receiver for domestic use. Assume that the valves are triode-heptode, variable-μ pentode, double-diode-triode, output tetrode, power rectifier, and the receiver is to operate from 200-, 230- and 250-volt mains supply. Include a surge-limiting resistor in the heater supply and explain its action.

Give reasons for the relative positions of the heater connexions of the various valves. Mention particular points to which special attention must be paid in the design of such a receiver in the interests of the user's safety.

(Brit. I.R.E., Radio Reception, November 1951.)

2. Describe a vibrator type of power supply using

(*a*) an additional vacuum rectifier ;

(*b*) no additional rectifier.

Suggest an arrangement of filter components designed to keep the audio-frequency ripple and the radio-frequency interference at a suitably low level.

(Brit. I.R.E., Radio Reception, November 1950.)

3. Give the circuit diagram of a smoothing circuit for use with the rectifier circuit of a domestic receiver working from a 50-c/s supply. Mark on the diagram suitable component values.

(Brit. I.R.E., Principles of Radio Engineering, May 1951.)

4. Describe, with the aid of a circuit diagram, the power-supply arrangements for an all-mains broadcast receiver (a.c. mains).

Show how a stabilised H.T. potential could be provided to one of the valves.

If the a.c. r.m.s. potential applied to the rectifier is 300 volts, what is the maximum value of the d.c. potential available as the H.T. supply? (C. & G., 1947.)

Answer : 424·4 volts.

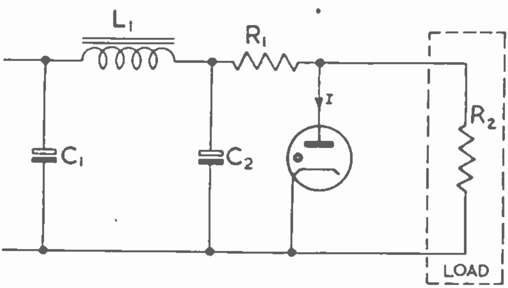
5. Describe a common type of metal rectifier, and indicate how it may be used to derive a d.c. output from an a.c. supply. Explain by means of free-hand graphs how the potential across the reservoir capacitor of a simple rectifier circuit varies with time. What is meant by ripple?

6. How would you provide the H.T. and L.T. supplies for a four-valve receiver from a 12-volt accumulator, the total H.T. consumption being approximately 50 mA at 250 volts, and the valve heaters each consuming 0·3 amp. at 6·0 volts? (C. & G., 1948.)

7. A mains-operated receiver working in the long-wave band experiences radio-frequency interference, which is conveyed to the receiver via the supply mains. Describe how the interference could be minimised at the receiver. (C. & G., 1949.)

8. Describe a circuit for supplying H.T. and heater power to an a.c./d.c. receiver, indicating typical values for the components, potentials and currents. (C. & G., 1949.)

9. A 150-volt neon stabiliser operates from a 250-volt d.c. supply using the circuit shown below :



If the load resistance R_2 is 10,000 ohms and the current I through the neon stabiliser is 15 mA, determine the resistance of, and the power dissipated in, the resistor R_1.

If the load resistance drops to 7,500 ohms, determine the modified value of the current through the neon stabiliser, assuming that it operates ideally. (C. & G., 1951).

Answer : 3·3 kΩ, 3·0 watts, 10 mA.

10. Indicate the methods adopted in practice for the provision of d.c. power from a single-phase a.c. supply. Explain how the output-potential ripple may be minimised and indicate how good regulation may be obtained.

(Brit. I.R.E., Radio Engineering, November 1945.)

11. Explain, with the aid of a circuit diagram, the principles of operation of a H.T. supply unit employing a full-wave bridge network of metal rectifiers. Illustrate your answer with sketches of the potential and current waveforms at suitable points in the circuit. (C. & G., 1952.)

CHAPTER 7

TRANSMITTERS

THE term " transmitter " is properly applied to the equipment used for generating and supplying the aerial with modulated radio-frequency power.

It comprises a radio-frequency oscillator, amplifier, modulator and a power amplifier. A block diagram of a typical transmitter

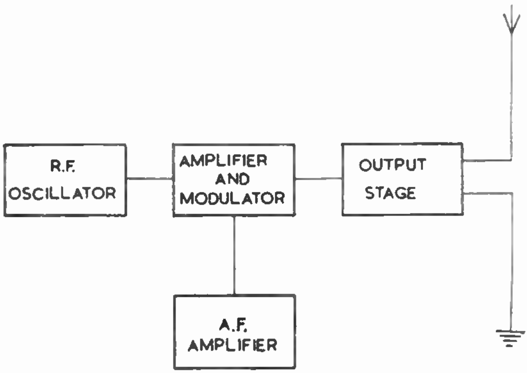


FIG. VII.1.—Block Diagram of Transmitter

is shown in Fig. VII.1. Various power supplies are also necessary for the operation of these stages, but these will not be considered in this volume.

The output power of a transmitting station in this country is usually given in terms of the power fed to the aerial, and transmitters delivering above 10 kW to the aerial are usually considered to be " high-power " transmitters. If the aerial input is 10 kW or less, the station is " low-power ".

7.1. Radio-frequency Oscillators

The generation of radio-frequencies has been discussed in Chapter 5, and as constancy of frequency is of prime importance in a transmitter, a crystal oscillator is most commonly used. Broadcasting stations on frequencies below 1·6 Mc/s are required by International Radiocommunications Regulations to operate within ± 20 c/s of their nominal frequency, corresponding to an accuracy of ±12·5 parts in a million at 1·6 Mc/s. Above 1·6 Mc/s an accuracy of ±30 parts per million is required (100 parts per million if the power is less than 200 watts). Low-power fixed stations for communications, *e.g.*, ship to shore, are permitted a maximum tolerance of ±500 parts per million below 1·6 Mc/s and ±30 parts per million up to 30 Mc/s. Mobile stations, *e.g.*, ships, aircraft, are permitted a tolerance of ±1 part per thousand up to 1·6 Mc/s and ±100 parts per million up to 30 Mc/s.

Between 30 and 100 Mc/s a general tolerance of ±200 parts per million applies. In high-power transmitters the carrier frequency must be held so closely to the allotted value that the crystal and its associated stage is kept at a constant temperature under steady operating conditions.

Low-power and portable transmitters do not need to be so closely controlled, nor are such elaborate schemes practicable in the latter case.

The highest frequency for which a quartz crystal of reasonable dimensions can be made is about 20 Mc/s, and if a higher frequency is required, frequency-multiplying stages must be used.

7.2. Frequency-multiplication

Although for the purposes of calculation valve characteristics are usually considered to be straight lines, there is inevitably some curvature present.

The requirement of an amplifying stage is to increase either the amplitude or the power of the signal without changing its characteristics. The circuit is therefore designed so that the dynamic characteristics are as straight as possible.

Consider now an amplifier in which the valve is biased to cut-off in the absence of a signal, so that when a sinusoidal p.d. is applied to the grid, anode current flows only during the positive half-cycle of the signal (in other words, Class " B " operation). Assuming that the valve characteristics are straight lines, the anode current wave will have the same shape as the signal p.d. during the latter's positive half-cycle, as shown in Fig. VII.2. Fig. VII.3 shows graphically the operation of frequency multiplication by such a stage.

The first curve shows the variation of the grid potential when the signal is applied, and the second curve shows the corresponding anode current. If the mean anode current, that is the current which would be indicated by a moving-coil ammeter in the anode circuit, is taken as datum, the instantaneous anode current varies above and below this line at signal frequency. It therefore contains a component at signal frequency, although its waveform is different from that of the signal.

The third part of the diagram shows a sinusoidal wave at signal frequency and another sinusoidal wave at twice signal frequency, whose relative amplitudes have been specially chosen. The amplitude of the signal or fundamental frequency component is one-half of the maximum value of the anode current, while

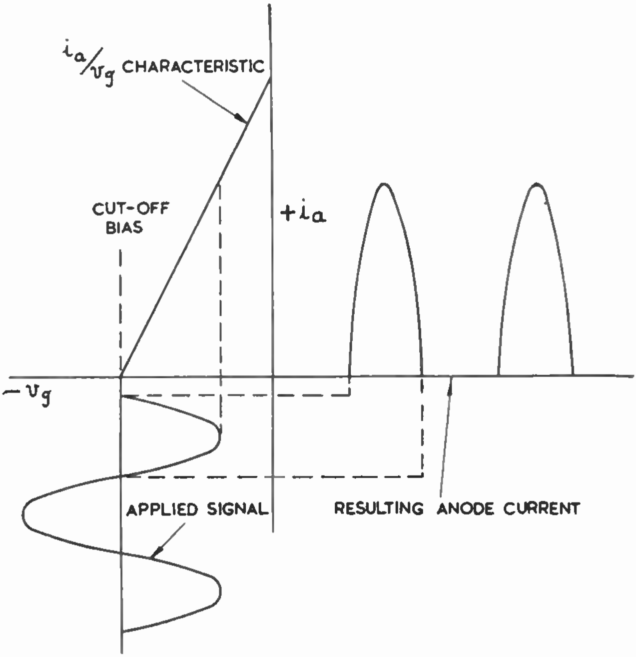


FIG. VII.2.—Waveforms in Class "B" Amplifier

that of the double-frequency or second harmonic component is $I_{\text{max.}}\left(\frac{1}{2} - \frac{1}{\pi}\right)$. When the two component waves are at their positive maxima we have

$$I_{\text{resultant}} = \frac{I_{\text{max.}}}{2} + I_{\text{max.}}\left(\frac{1}{2} - \frac{1}{\pi}\right)$$

$$= I_{\text{max.}} - \frac{I_{\text{max.}}}{\pi}$$

$$= I_{\text{max.}}\left(1 - \frac{1}{\pi}\right)$$

and when the signal frequency component is at its negative peak, we have

$$I_{\text{resultant}} = \frac{I_{\text{max.}}}{2} + I_{\text{max.}}\left(\tfrac{1}{2} - \frac{1}{\pi}\right)$$

$$= -\frac{1}{\pi} I_{\text{max.}}$$

The excursion of current is thus from $I_{\text{max.}}\left(1 - \frac{1}{\pi}\right)$ to $-\frac{1}{\pi} I_{\text{max.}}$

that is, a total excursion of $I_{\text{max.}}$

Fig. VII.3 shows how these values fit the curves.

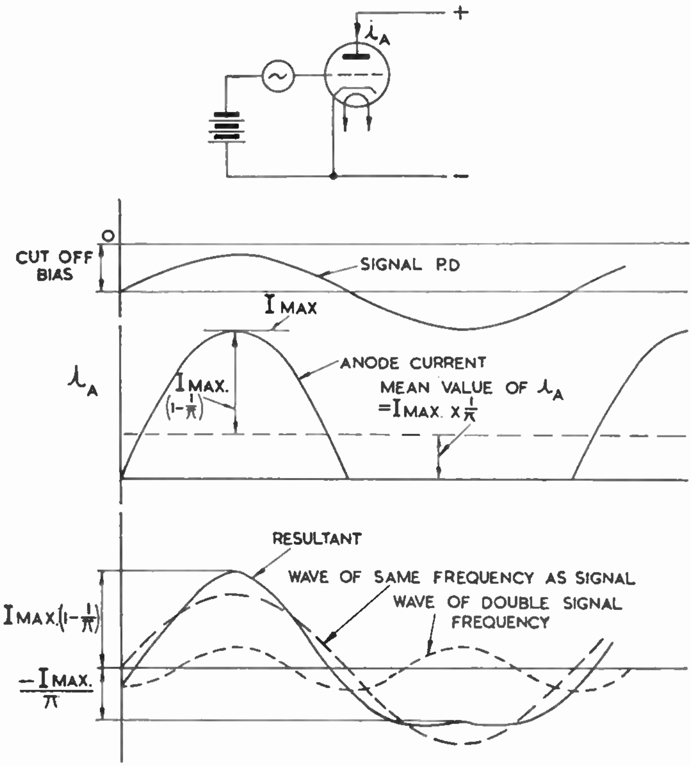


FIG. VII.3.—Anode Current in Class "B" Amplifier

Thus, by the addition of two sinusoidal quantities, one of the same frequency as the signal p.d. and the other of twice that frequency a waveform is synthesised which has a somewhat similar shape to that produced in the anode circuit of Fig. VII.2 when a sinusoidal p.d. is applied to the grid.

If the anode load of such a stage is a parallel resonant circuit adjusted to resonate at twice the signal frequency, an appreciable p.d. at this double frequency will appear across the tuned circuit. Fig. VII.4 shows the circuit of such a frequency doubling stage.

The waveform synthesised in Fig. VII 3 is not identical with that produced in the anode circuit of Fig. VII.2, but by adding more components at higher multiples of the signal frequency the synthesis can be made to approach identity. It is therefore possible to obtain an output at three, four or more times the frequency applied to the grid by tuning the anode load to the appropriate frequency, but the amplitude falls as the multiple rises. Treble-frequency output is often utilised in practice.

The curvature of valve characteristics which exists in practice increases the amplitude of the higher harmonics above the theoretical values.

By following one multiplying stage by others, considerable multiplication is possible while maintaining the same degree of stability of frequency. The highest frequencies at which normal techniques are used (say 100 Mc/s) can thus be derived from a master crystal oscillator operating at 20 Mc/s or less. For example, if a carrier frequency of 56 Mc/s is required, a 14-Mc/s crystal followed by two doublers could be used—$14 \times 2 \times 2 = 56$ Mc/s.

The single-valve multiplier is satisfactory for use in stages requiring only voltage amplification and where no appreciable power has to be developed, because the peak value of the second-harmonic component is only about $\frac{1}{5}$ the peak value of the valve current. Since the peak current in the valve is a limiting factor, it is clear that a doubling stage makes use of only a fraction of the power which the valve is capable of delivering.

7.3. The Push–Push Doubler

By feeding the grids of two valves biased to cut-off and whose anodes are connected in parallel, with signals which are 180° out-of-phase with each other, two pulses of anode current are produced in every cycle of the drive.

Fig. VII.5 shows the arrangement which is known as a " push–push " doubler. The two grids are driven from the ends of a

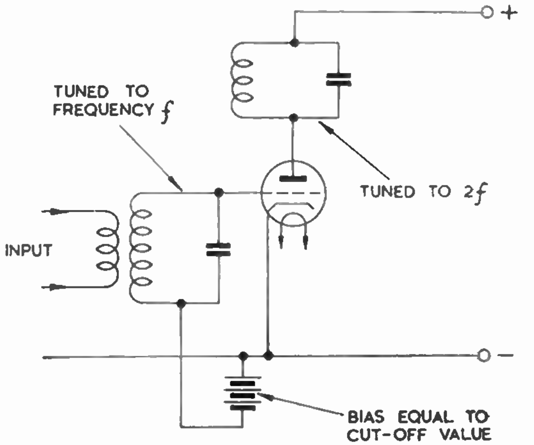


Fig. VII.4.—Class "B" Amplifier Operating as Frequency-doubler

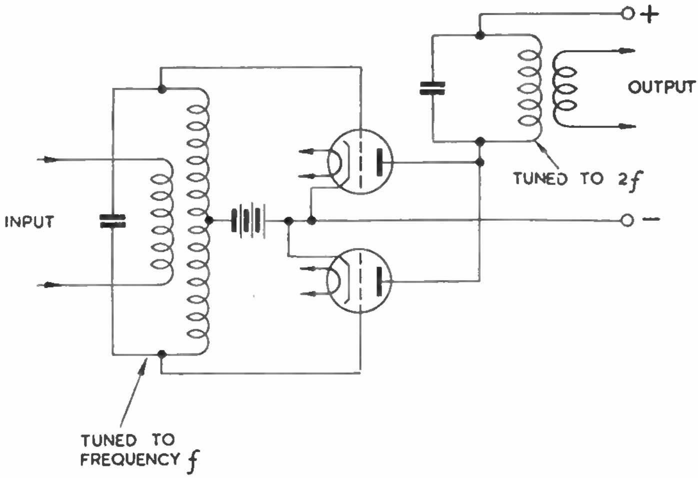


Fig. VII.5.—Push–Push Frequency Doubler

centre-tapped coil whose centre point is taken to the grid-bias supply. The common anode load of the two valves is a tuned circuit whose resonant frequency is twice that of the grid signal. Fig. VII.6 shows the potentials and currents at various parts of the circuit, and the major component in the anode circuit is now a current at second-harmonic frequency, there being no fundamental component present.

The name of the circuit is derived from the fact that the valves are " pushed " into operation on a common load on alternate half-

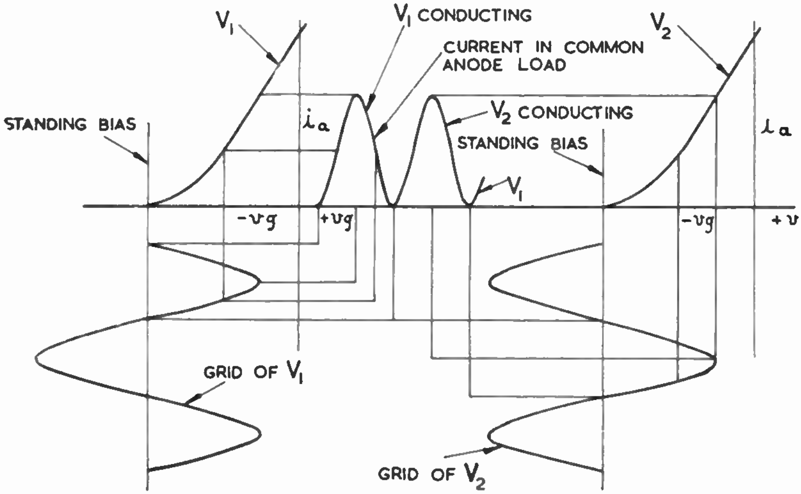


FIG. VII.6.—Waveforms for Push-Push Frequency Doubler

cycles. The output frequency having the largest component is that at twice the input frequency.

7.4. Modulation

Fig. VII.7 shows a typical amplitude-modulated radio-frequency wave. Reference to Chapter 9, Volume 1 will show that when a carrier wave (v_r) of the form $V_{max.} \sin \omega t$ is amplitude modulated by an audio frequency (v_a) of the form $m V_{max.} \sin pt$, the resultant modulated carrier (v) is $V_{max.} \sin \omega t\,(1 + m \sin pt)$, where m is the modulation index $= \left(\frac{\%\ \text{modulation}}{100}\right)$, and $p = 2\pi \times$ audio frequency.

Expanding, we have $v = V_{max.} \sin \omega t + mV_{max.} \sin \omega t \sin pt.$

$$\therefore v = V_{max.} \sin \omega t + \frac{mV_{max.}}{2} \cos(\omega - p)t - \frac{mV_{max.}}{2} \cos(\omega + p)t.$$

This result shows the existence of two side frequencies, one below and one above the carrier frequency. The difference between the two side frequencies is twice the modulating frequency $(p/2\pi)$, and the amplitude of each of the side frequencies is $\frac{m}{2}$ times that of the carrier wave amplitude.

The total energy in an amplitude-modulated wave is obtained

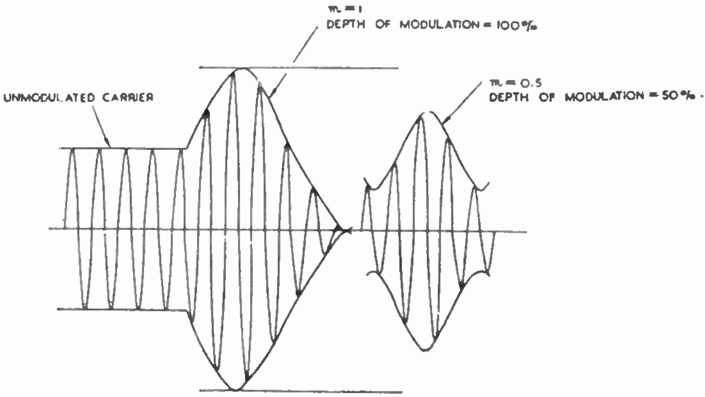


FIG. VII.7.—Amplitude-modulated Radio-frequency Wave

by adding the energy contents of the carrier and the two side bands. (Each modulating frequency produces two side frequencies, one upper and one lower, so that when more than one modulating frequency is present the term " side bands " is used to indicate the upper and lower bands of frequencies produced as a result of modulation.)

The power in the carrier is proportional to $V_{max.}^2$. Power in the lower side frequency is proportional to $\left(\frac{mV_{max.}}{2}\right)^2 = \frac{m^2V_{max.}^2}{4}$. Power in the upper side frequency is proportional to $\left(\frac{mV_{max.}}{2}\right)^2 = \frac{m^2V_{max.}^2}{4}$ so that the increase of power when modulated 100 per cent. by one frequency is proportional to $\frac{V_{max.}^2}{4} + \frac{V_{max.}^2}{4} = \frac{V_{max.}^2}{2}$

$\therefore$ the total power radiated is proportional to

$$V_{\text{max.}}^2 + \frac{V_{\text{max.}}^2}{2} = \text{carrier power } (1 + \tfrac{1}{2})$$

In general, the effective amplitude of a sinusoidally modulated wave is proportional to $\sqrt{1 + \frac{m^2}{2}}$ so that the total power radiated when the depth of modulation is denoted by m is proportional to $V_{\text{max.}}^2 + \frac{m V_{\text{max.}}^2}{2} = \text{carrier power} \times \left(1 + \frac{m^2}{2}\right)$

e.g., a 20-kW transmitter when modulated to 40 per cent. will radiate a total power of :

$$20\left(1 + \frac{0{\cdot}4^2}{2}\right) = 21{\cdot}6 \text{ kW}$$

The same transmitter, however, when modulated to a depth of 80 per cent. will radiate a total power of :

$$20\left(1 + \frac{0{\cdot}8^2}{2}\right) = 26{\cdot}4 \text{ kW}$$

and at 100 per cent. modulation will radiate a total power of :

$$20\left(1 + \frac{1^2}{2}\right) = 30 \text{ kW}$$

The percentage modulation will vary about the mean value because of the varying intensities of speech or music, and to give a good signal/noise ratio it should be kept as high as possible.

From the point of view of the radio-frequency amplifier, it is desirable to introduce the modulation as late in the chain as possible. The side frequencies caused by modulation must be passed through the tuned circuits of the radio-frequency amplifiers in true proportion. Since each tuned circuit causes some attenuation of these side frequencies, particularly the higher ones, it is well to reduce the number of such circuits. Series modulation is restricted to low-power stages because the modulator must work in Class " A " conditions, and the H.T. power required becomes excessive when the method is applied to high-power stages.

Amplifier stages which have to handle the modulated signal must be designed for 100 per cent. modulation conditions, when the carrier alternates at modulation frequency between zero and twice the unmodulated carrier amplitude. For a considerable proportion of time, therefore, they will not be working to full capacity, and the valves and power supplies are of uneconomic size. Further, the efficiency of a given simple Class " B " amplifier is proportional to the drive, and so if the efficiency at full-power (100 per cent. modulation) is 60 per cent., the efficiency at half-power (normal carrier only) will be only about 30 per cent.

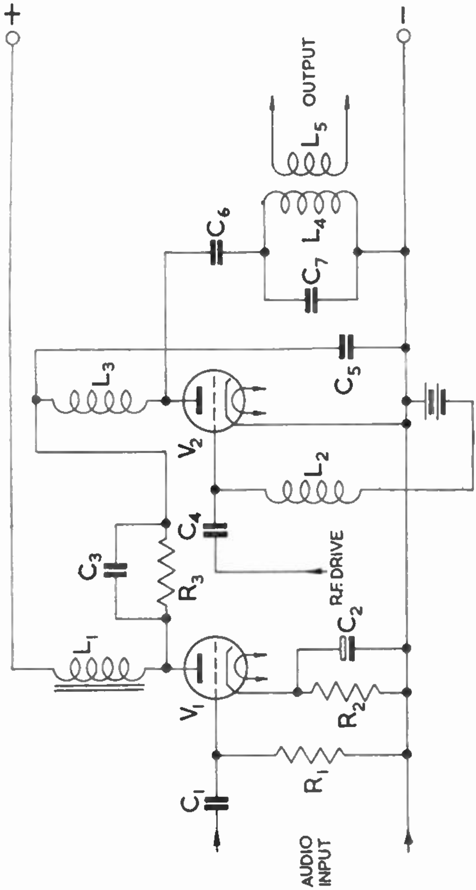


FIG. VII.8.—Anode Modulation

Special circuits have been developed which have greater efficiency than a straight-forward Class " B " amplifier, but the perfection of high-power audio-frequency amplifiers has enabled anode modulation to be applied to the final stage of transmitters of the highest power. This arrangement has a high overall efficiency.

Figs. VII.8 and 9 show typical circuits for anode and grid modulation; these arrangements can be used for any stage in the chain. In Fig. VII.8 the components L_3 and C_5 form a radio-frequency

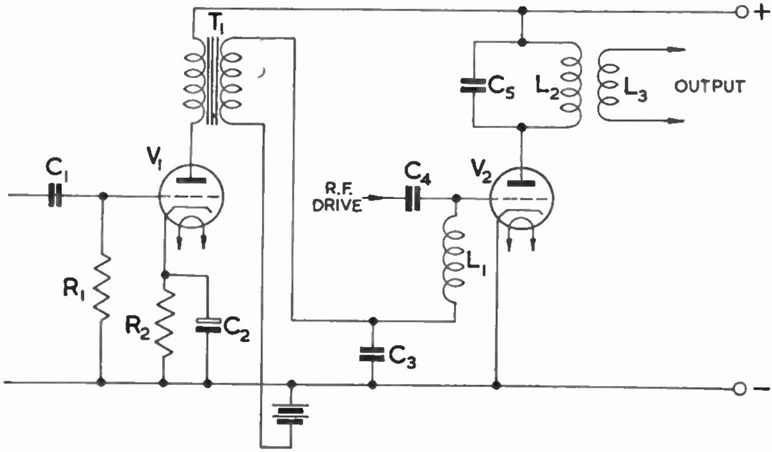


FIG. VII.9.—Grid Modulation

filter to prevent radio-frequency from reaching V_1. R_3 reduces the H.T. applied to V_2, while C_3, which has a low impedance at audio frequencies, passes the full swing from the anode of V_1 and permits modulation to approach 100 per cent. In Fig. VII.9 L_1 and C_3 form a radio-frequency filter. The circuits shown are for single-sided radio-frequency amplifiers; Fig. VII.10 shows an anode-modulated push–pull radio-frequency amplifier.

7.5. Telegraphy Transmitters

In the case of C.W. telegraphy, the carrier has to be switched on and off according to the code of transmission. It is therefore a special case of modulation by relatively simple means.

A C.W. telegraphy transmitter is designed for one running condition only, namely full carrier, and to attain a high efficiency has a Class " C " power amplifier.

Fig. VII.11 shows the principal units in a telegraphy transmitter. The master oscillator is followed by a " buffer " stage, whose function is mainly to prevent subsequent stages from loading, and so altering the frequency of the oscillator. The buffer stage will

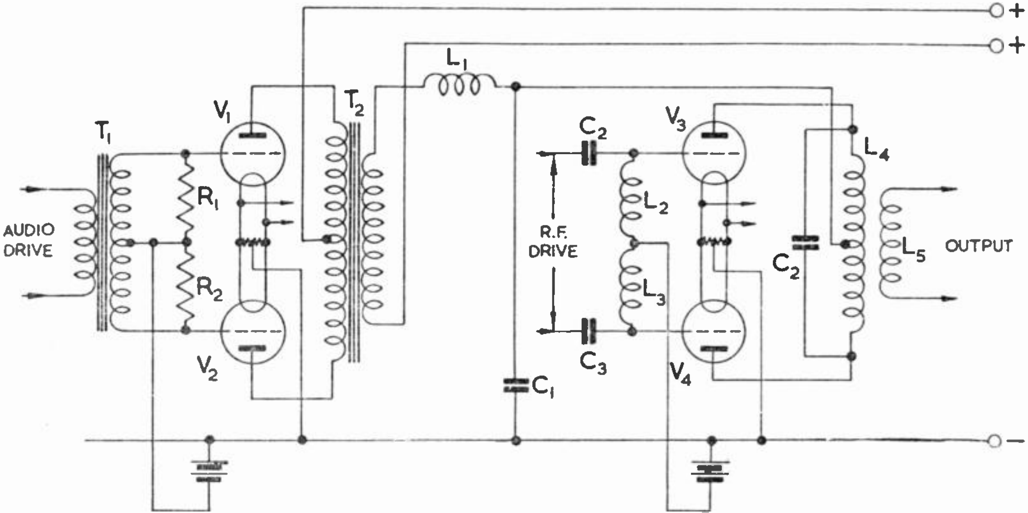


FIG. VII.10.—Anode-modulated Push-Pull Radio-frequency Amplifier probably give some gain; it may be a frequency doubling stage if required.

The switching, or "keying" as it is called, may be done by various methods.

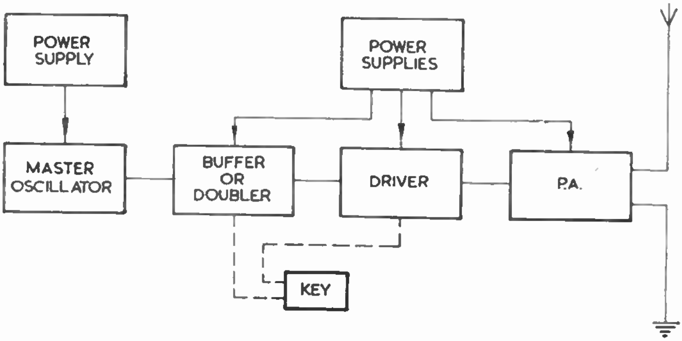


FIG. VII.11.—Block Diagram of Telegraphy Transmitter

A very common method in small transmitters is to remove the anode supply from one or more stages, particularly the output stage. This may be done either by switching the supply direct to the amplifier or by switching off the power-supply unit on the primary side. Fig. VII.12 shows the principle of both of these

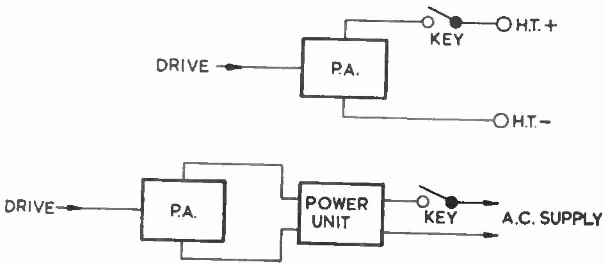


FIG. VII.12.—Methods of Keying

methods. When the power of the output stages is greater, this method can still be applied to the earlier stages.

Another method which can be applied to the buffer stage or to a later stage is grid-bias keying. In this case, the valve being keyed is given an excessive grid bias so that it is cut off and gives no output. The key is arranged to switch over to the normal value of bias during the " on " or " mark " periods only, and the valve then performs its normal function (Fig. VII.13).

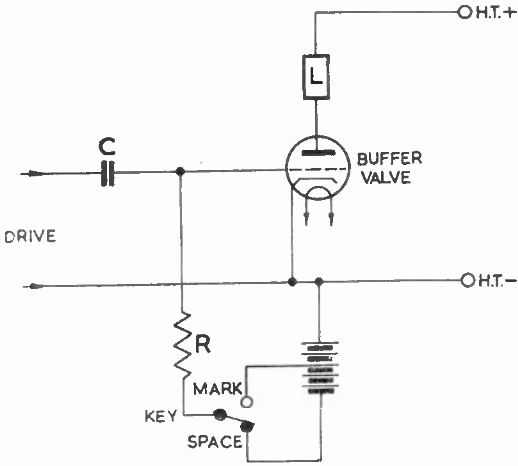


FIG. VII.13.—Grid-bias Keying

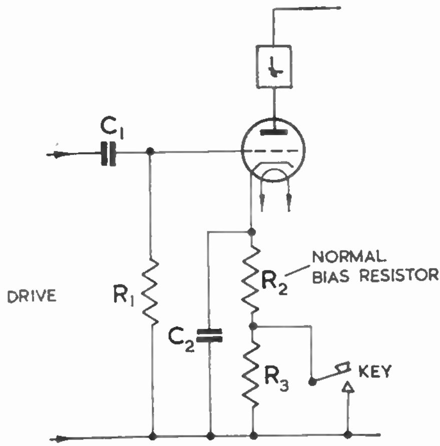


FIG. VII.14.—Cathode Keying

Another method is to use cathode keying, in which the bias is increased by inserting a high resistance in the cathode circuit of the keyed valve during the " off " or " space " period (Fig. VII.14).

7.6. Key-clicks

If a transmitter is switched on or off very suddenly the effect is comparable to modulating with rectangular waves, which have very high component frequencies, and cause the radiation of high side-frequencies. This again makes the transmitter occupy a wide band of frequencies in the spectrum which naturally is

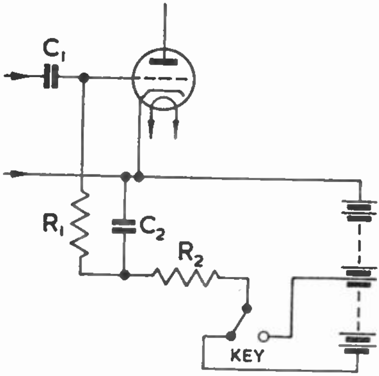


FIG. VII.15.—Filter for Grid-bias Keying

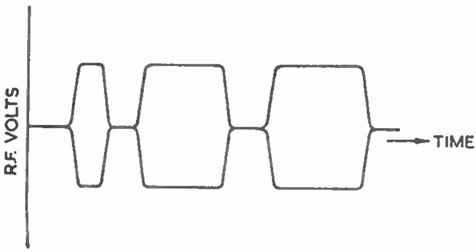


FIG. VII.16.—Envelope of Telegraphy Transmission

undesirable, and will cause interference with transmitters on nearby channels. One object of using telegraphy—apart from the use of codes to preserve secrecy—is to reduce the bandwidth occupied by a transmitter, and so permit a large number of services to be operated.

Sudden switch-on and switch-off is prevented by means of resistance-capacitance filters in the keying circuits.

In the case of primary keying, the necessary smoothing of the switching occurs in the rectifier filter circuits, and the only precaution necessary is some spark-suppression device on the main contacts performing the switching.

When grid-bias switching is employed, a simple R-C filter to prevent sudden changes of grid bias on the controlled valve suffices (Fig. VII.15). Fig. VII.16 shows the envelope of a typical telegraphy transmission.

7.7. Tank Circuits

The tuned circuit in the anode of the final stage is known as the "tank circuit", because its function is to act as a reservoir of energy at the frequency of transmission. The inductor of the tank circuit usually acts also as the primary winding of a trans-

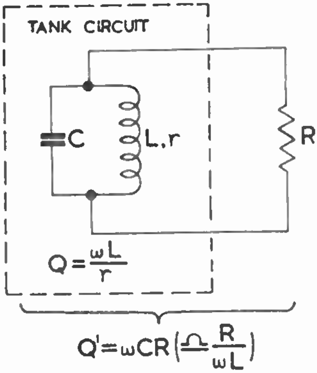


FIG. VII.17.—Effect of Aerial Load

former; the load is fed from the secondary winding, and by adjustment of the turns ratio may be properly matched to the valve. The losses which occur in the tank circuit reduce the useful output, and it is therefore essential to keep these losses as small as possible (see Chapter 1). In order to prevent the radiation of harmonics, the Q of the tank circuit, when loaded, is usually made about 10. Fig. VII.17 shows the equivalent circuit of the resonant tank, and the resistive load due to the aerial. This shows the effective loaded Q of the circuit $Q' = \omega CR$. The Q of the tank circuit considered alone, $Q = \frac{\omega L}{r}$, which should be high—at least 100—for high circuit efficiency. The efficiency of the circuit is equal to

$$\frac{\text{power dissipated in } R}{\text{power supplied to the whole circuit}} = \frac{Q - Q'}{Q} \times 100 \text{ per cent.}$$

so that if $Q = 100$ and $Q' = 10$ the circuit efficiency η is given by $\frac{100 - 10}{100} \times 100$ per cent. $= 90$ per cent.

7.8. Monitoring of Transmitters

Large transmitters are fitted with voltmeters and ammeters to enable the conditions of each stage to be checked. The criterion is, of course, that power is being delivered to the aerial, and this is indicated by an ammeter. Small transmitters have an instrument of the thermo-couple type connected between the output coil and earth, while larger ones use a current transformer in one of the aerial feeders.

It is most important not to exceed the maximum ratings of valves and components, and protective devices are used to switch off H.T. supplies if excessive current flows. Protection against failure of cooling water or air is also provided.

7.9. Initial Alignment

Let us consider a small transmitter comprising a master (crystal) oscillator, a Class " A " buffer stage, a Class " B " driver stage and a Class " C " output stage (p.a.). The oscillator and buffer stages are energised and allowed to reach their working temperatures, and the cathodes of the driver and p.a. are energised. Grid bias is applied to the driver and p.a. valves. H.T. is then applied to the driver, slowly raising the supply p.d. until a measurable fraction of normal anode current flows in the driver stage. The anode load of the driver is then tuned until the anode current is a minimum, and the neutralising capacitor is adjusted to reduce the anode current still further. The alternating p.d. on the driver anode load is then a maximum, indicating correct resonant frequency and neutralisation. Some increase in the H.T. on the driver may be necessary to obtain a useful indication, followed by alternate adjustment of resonant frequency and neutralising capacitor for greater accuracy of setting.

H.T. is then slowly increased on the p.a. stage and similar adjustments made. When the resonant circuits and neutralisation have been adjusted, H.T. on the p.a. and driver are increased. The anode ammeters are watched to ensure that the maximum current is never exceeded, and tuning and neutralisation are checked periodically. The p.a. load is then applied by increasing the coupling to the aerial, again watching the p.a. ammeter and

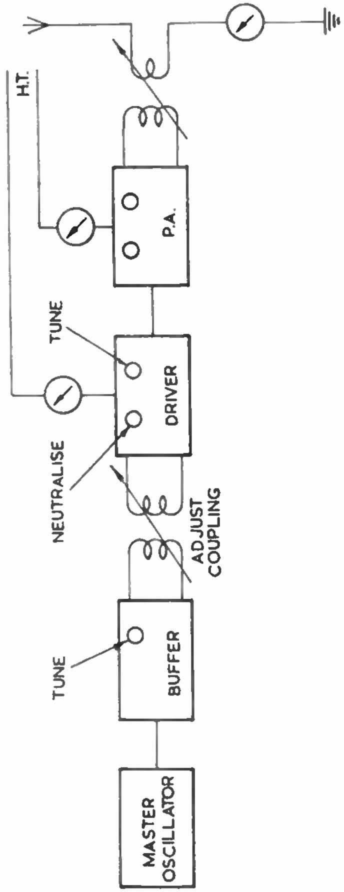


FIG. VII.18.—Transmitter Controls

the aerial ammeter. Final adjustment of drive, by tuning the buffer or increasing the coupling between buffer and driver, may then be necessary (Fig. VII.18).

QUESTIONS

1. Describe briefly two methods of speech amplitude modulation in a radio frequency transmitter. (C. & G., 1946.)

2. Describe, with the aid of a block schematic diagram, a crystal-controlled telegraph transmitter suitable for operation on a frequency of 15 Mc/s and giving an output of 1 kW. Indicate a suitable method for on–off keying the transmitted carrier. (C. & G., 1951.)

3. Give an outline description, with block diagrams, of a medium-frequency amplitude-modulated radio-telephony transmitter. (C. & G., 1948.)

4. Explain, with the aid of diagrams showing the grid potential and anode current waveforms, what is meant by Class " B " and Class " C " operation of transmitting valves. State the maximum efficiencies to be expected in each case. (C. & G., 1947.)

5. Explain, with the aid of a block schematic diagram, the principles of operation of a simple amplitude-modulated telephony transmitter. (C. & G., 1952.)

SPECIMEN ANSWER

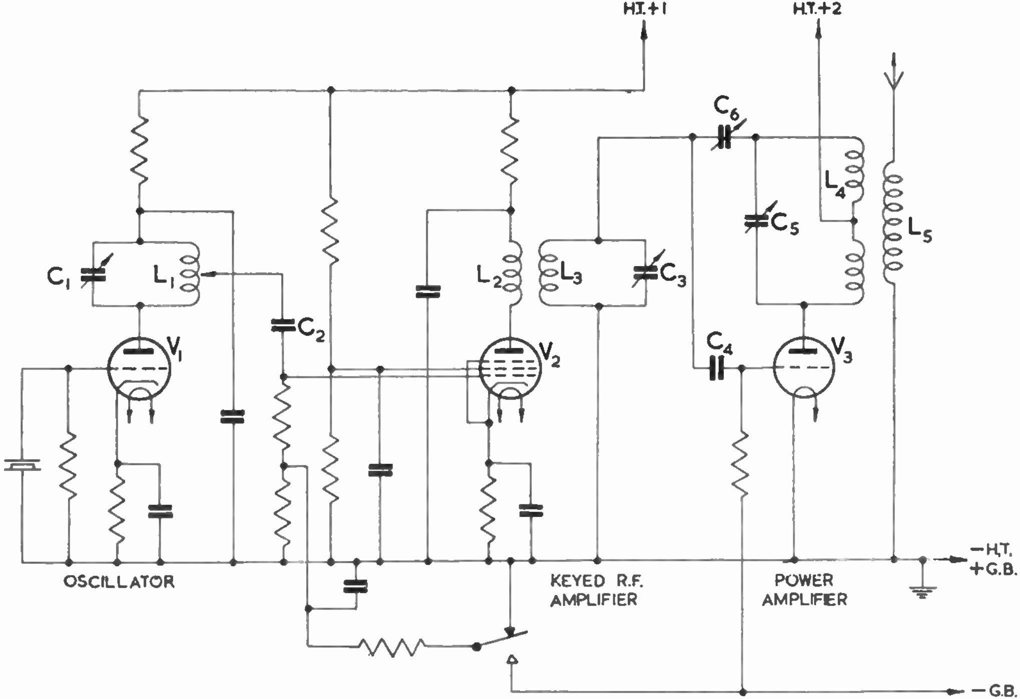
Q. Describe the essential features of a short-wave telegraph transmitter for continuous-wave operation. Show how the transmitter may be keyed, and indicate the waveform of the radiated signals. (C. & G., 1947.)

A. The circuit arrangement shown in Fig. VII.Q.1. represents a three-stage radio-telegraph sender for continuous-wave operation.

The stages used are a crystal-controlled master oscillator, a radio-frequency low-power amplifier and a power-output stage coupled to the aerial.

The oscillator is a tuned-anode tuned-grid circuit operating at the natural frequency of the quartz crystal in the grid circuit The anode load of V_1, consists of the circuit L_1C_1 tuned to a frequency higher than the operating frequency, so as to present an inductive load. If the operating frequency is higher than about 10 Mc/s the crystal will operate at about 5 Mc/s, and a harmonic of the crystal frequency will be used to provide the carrier frequency.

The oscillator is coupled by C_2 to the lower-power amplifier V_2 with a tuned anode load provided by L_2, L_3 and C_3. Keying is effected at this stage, the normal bias provided by R_2 being replaced by a bias potential sufficient to cut off the valve in the spacing condition. Keying may be by a direct key in the grid circuit or by a remote key controlling a telegraph relay as shown.



H.T.+1
H.T.+2
C_6
L_4
C_5
L_5
C_1
L_1
L_2
L_3
C_3
C_2
V_1
V_2
C_4
V_3
OSCILLATOR
KEYED R.F. AMPLIFIER
POWER AMPLIFIER
−H.T.
+G.B.
−G.B.

Fig. VII.Q.1

The output stage V_3 is operated as a Class " C " amplifier to obtain high efficiency, thus the standing bias is sufficient to cut-off the valve, but the radio-frequency drive from the V_2 stage is sufficient to overcome this bias. The aerial is mutually excited from this stage via L_4, L_5 tuned by C_5. Neutralisation of the output stage is effected by the neutralising capacitor C_6.

The ideal waveform of the radiated signal is as shown in Fig.

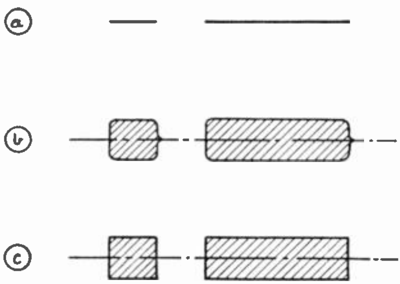


FIG. VII.Q.2

VII.Q.2(*b*), corresponding to key operations indicated in Fig. VII.Q.2(*a*). If the modulation envelope shown in Fig. VII.Q.2(*c*) were radiated it would require frequency components considerably removed from the carrier frequency. This would mean that the transmitted bandwidth would be excessive and cause interference with channels working on adjacent frequency allocations. To avoid this, the rounded waveform of Fig. VII.Q.2(*b*) is produced by introducing a time-delay circuit R_3, C_7 in the keying circuit

(P.O. Eng. Dept.)

CHAPTER 8

RECEIVERS

RECEIVERS fall into three main classifications: (1) communications receivers, (2) broadcast receivers, (3) vision receivers, whose functions differ considerably. Communications receivers fall into other classifications according to their duty: (1*a*) receivers for certain allotted frequency bands, (1*b*) receivers for general coverage, (1*c*) mobile receivers for special services (*e.g.*, Police).

The two main classifications are derived in the following way: (1) Receivers for communication purposes should be capable of producing an intelligible signal of code or speech from the desired transmission under almost any circumstances. (2) Broadcast receivers are intended to provide entertainment from speech and music of a high quality from a few relatively near stations, and are usually also designed to provide reasonable entertainment from a considerable number of more distant stations. Vision entertainment is also transmitted for relatively near reception from special stations.

Although the basic principles of all classes are identical, circuit details vary according to the use of the receiver.

8.1. Sensitivity

A communications receiver must have the highest possible sensitivity, and to obtain it a careful division of the total gain between the various sections of the receiver must be made. In order to obtain the required total gain, and to have a high enough selectivity (which should be controllable), such a receiver is invariably of the supersonic heterodyne type (see Volume I, Chapter 10). A block diagram is shown in Fig. VIII.1. A mixer stage is prone to introduce noticeable " noise " if its input is small (less than about 20 μV), and to ensure that a reasonable signal level at the mixer input is obtained one or two stages of signal-frequency amplification may be provided. The total gain of a superheterodyne receiver may be much higher than that of a straight receiver because the amplification is obtained at two or more different frequencies, and instability caused by feedback is less likely to be troublesome.

Modern broadcast receivers are also invariably of superheterodyne type, and as the required sensitivity is less than that of a communications receiver the signal-frequency amplifier is often omitted (see Fig. VIII.2). Important reasons for the use of superheterodynes as broadcast receivers are : (1) ease of operation, (2)

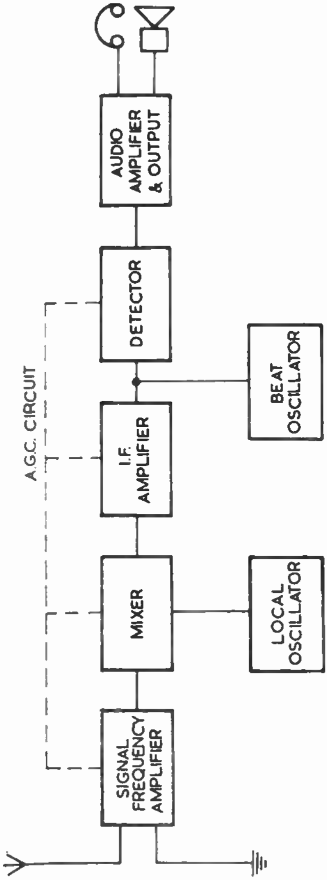


FIG. VIII.1.—Communications Receiver

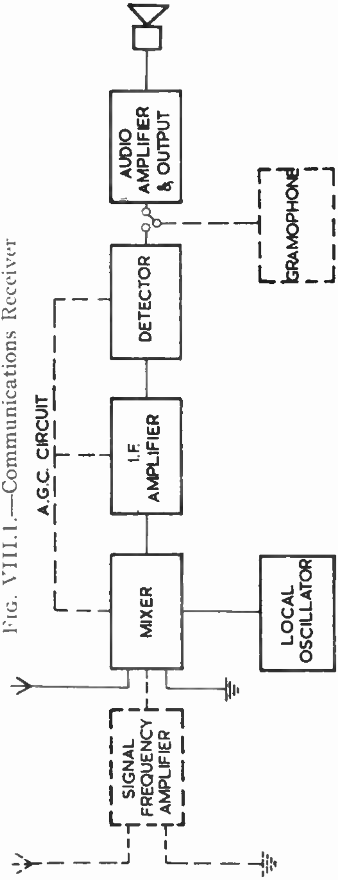


FIG. VIII.2.—Broadcast Receiver

good selectivity, (3) ease of production in a form to give (1) and (2).

The sensitivity of the receiver must be high enough to ensure reception of the weakest signal to be received, and it will therefore be too high for stronger signals. To avoid unnecessary manual adjustment of gain when the carrier strength varies (as it probably will when tuning from one carrier to another), an automatic control is normally arranged to give a fairly constant audio output for all signals above a minimum " threshold " value. This arrangement is known as " Automatic Gain Control " (A.G.C.), and controls the gain of the early stages according to the carrier level.

8.2. Selectivity

An advantage of the superheterodyne is the constant selectivity imparted by the intermediate (fixed) frequency amplifier. With an adjustable tuned circuit of constant Q, as is closely the case in practice, the bandwidth for a given attenuation is a constant fraction of the resonant frequency (see Chapter 4). The bandwidth therefore varies over the normal tuning range on one coil by about $2\frac{1}{2}$ to 1 in frequency.

The intermediate frequency (i.f.) is usually between 450 and 470 kc/s, as there are few transmitting stations working about these values, and direct reception at intermediate frequency is unlikely. A very common intermediate frequency is 465 kc/s, and suitable components are readily available. A reason for the use of a higher value of intermediate frequency in some cases is given below (see " Interference ").

The intermediate-frequency amplifier has band-pass tuned circuits (see Chapter 4), and is very selective against signals near (in frequency) to the desired one. The fixed frequency enables simple circuits to be used to give adjustable bandwidth if required.

8.3. Mixing

In the mixer stage the incoming signal modulates a local oscillator, and one of the products of modulation, usually the frequency equal to the difference between the two frequencies, is selected for further amplification.

In the pentagrid mixers (see Chapter 2) the anode current depends on the potential of the first grid and the signal grid. The electron current passing the first grid is proportional to e_1g_{m1}, where e_1 is the potential of the first (oscillator) grid and g_{m1} its mutual conductance. This current is then modulated by e_4g_{m4}, where e_4 is the potential of the fourth (signal) grid and g_{m4} its mutual conductance. " Surplus " current flows to g_3, the current to g_2 (the oscillator " anode ") being relatively small. The resultant final anode current is thus proportional to $e_1g_{m1} \times e_4g_{m4}$. A similar argument applies to the triode-hexode except that the order of modulation by g_1 and g_3 (signal and local oscillator grids respectively) is different.

Assuming constant values of g_{m1} and g_{m4} over the working range and sinusoidal p.d.s $E_o \sin \omega_o t$ (local oscillator) and $E_s \sin \omega_s t$ (incoming signal), the anode current is proportional to $g_{m1}\ g_{m4}$ $E_o \sin \omega_o t \,.\, E_s \sin \omega_s t$, that is,

$$\begin{aligned} &\propto E_o E_s \sin \omega_o t \sin \omega_s t \text{ (g_{m1} and g_{m4} being assumed constant)} \\ &\propto E_o E_s \{\tfrac{1}{2} \cos (\omega_o t - \omega_s t) - \tfrac{1}{2} \cos (\omega_o t + \omega_s t)\} \\ &\propto \tfrac{1}{2} E_o E_s \{\cos (\omega_o - \omega_s)t - \cos (\omega_o + \omega_s)t\} \end{aligned}$$

The output therefore contains a component $\{(\omega_o - \omega_s)t\}$ at a frequency equal to the difference between the local oscillation and the incoming signal frequencies and another component $\{(\omega_o + \omega_s)t\}$ at their sum. This is known as "multiplicative" mixing, because the two inputs are in effect multiplied together.

It is also possible to mix two frequencies in a valve whose i_a/e_g characteristic is some form of square-law. For example, if

$$i_a \propto K_1 e_g + K_2 e_g^2 + \ldots$$

and both signal and local oscillator p.d.s are applied to the same grid, that is

$$e_g = E_o \sin \omega_o t + E_s \sin \omega_s t$$

we have

$$\begin{aligned} i_a &\propto K_1(E_o \sin \omega_o t + E_s \sin \omega_s t) + K_2(E_o \sin \omega t + E_s \sin \omega_s t)^2 \\ &\propto K_1(E_o \sin \omega_o t + E_s \sin \omega_s t) \\ &\qquad + K_2(E_o^2 \sin^2 \omega_o t + E_s^2 \sin^2 \omega_s t + 2E_o E_s \sin \omega_o t \sin \omega_s t) \end{aligned}$$

The last term in the K_2 bracket is similar to that obtained when considering multiplicative mixing, so that outputs at sum and difference frequencies are obtained.

The first two terms in the K_2 bracket may be evaluated thus:

$$\sin^2 A = \tfrac{1}{2}(1 - \cos 2A)$$

$$\begin{aligned} \therefore \quad K_2 E_o^2 \sin^2 \omega_o t &= \frac{K_2}{2} E_o^2 (1 - \cos 2\omega_o t) \\ &= \frac{K_2}{2} E_o^2 - \frac{K_2}{2} E_o^2 \cos 2\omega_o t \end{aligned}$$

which is a constant quantity depending on the value of E_o together with a term at a frequency corresponding to $2\omega_o$. Similarly, there is a term at frequency corresponding to $2\omega_s$. The output from a square-law mixer therefore contains components at both the applied frequencies, together with their second harmonics and terms at the sum and difference of the applied frequencies.

The lower part of the i_a/e_g characteristic (including the bottom bend) of any valve is a law of this type, and the effect is increased if the valve is cut off by the negative half-cycles of the applied signals.

Multiplicative mixing in special valves (described in Chapter 2) is most commonly used. Typical circuits are shown in Fig. VIII.3 for a " Pentagrid Converter " or " Heptode " mixer and in

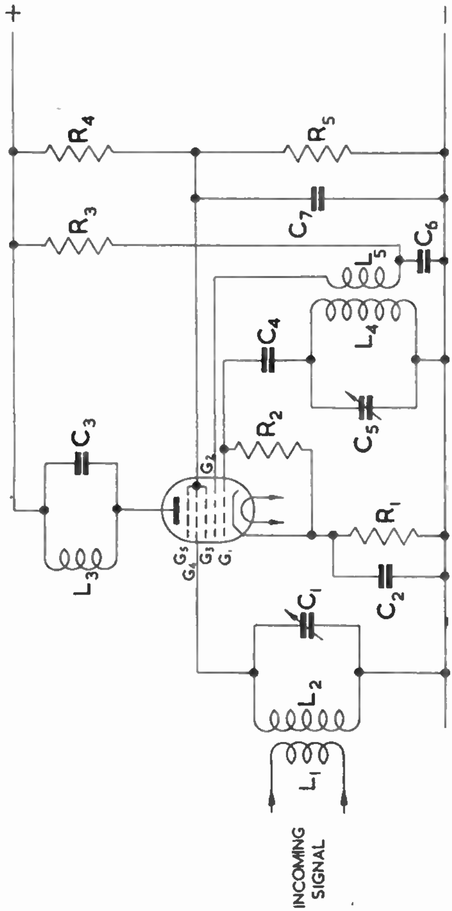


Fig. VIII.3.—Pentagrid Converter

Fig. VIII.4 for a " Triode-Hexode " valve. The Pentagrid is satisfactory for frequencies up to about 5 Mc/s, and with care may be used up to 20 Mc/s. At these higher frequencies the triode-hexode is easier to operate because of the greater mutual conductance of the oscillator section, which ensures ready self-oscillation at high radio-frequencies. The important parameter of a mixer valve is its "conversion conductance", which is the ratio of $\dfrac{\text{anode current change at difference frequency}}{\text{volts input at signal frequency}}$ in mA/volt. This ratio varies with the amplitude of the local oscillation, and the best working conditions are stated by the valve manufacturer. The conversion conductance of a triode-hexode frequency changer rarely exceeds 0·75 mA/volt compared with 2–3 mA/volt mutual conductance for the average radio-frequency pentode so that less gain is obtained. The signal grid of the mixer has a variable-mu characteristic like a radio-frequency pentode.

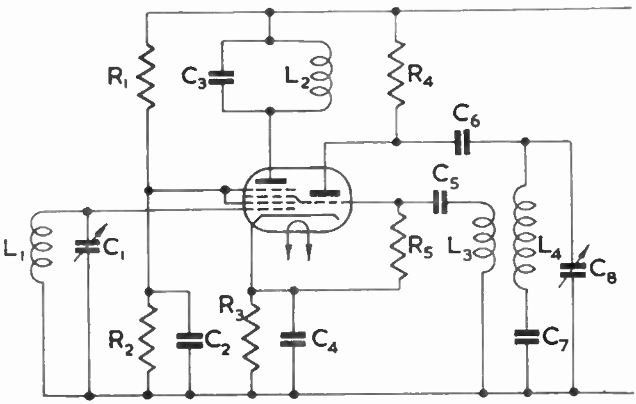


FIG. VIII.4.—Triode-hexode Mixer

The coupling between the local oscillator electrodes and the signal grid is greater in a pentagrid than in a triode-hexode because the electron stream is modulated first by the local oscillation. This is another reason for its greater difficulty of operation at high frequency. The signal grid is the first electrode to control the electron flow in the hexode section of a triode-hexode, and the coupling between oscillator and signal grids is therefore the least possible.

The local oscillator frequency is equal to signal frequency plus intermediate frequency. This is essential when the signal frequency is less than the intermediate frequency, and is more convenient than the reverse when ganged tuning (that is, tuning of oscillator and of signal-frequency circuits by mechanically coupled capacitors) is used, as the extra components to keep constant frequency difference are associated with the circuit resonant at the higher frequency. For example, if the intermediate frequency is 465 kc/s and the medium-wave tuning range is 1500–600 kc/s

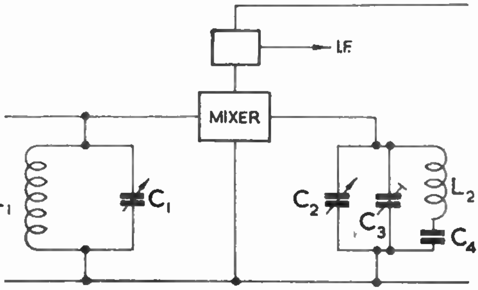


FIG. VIII.5.—Showing Trimming Capacitor C_3 and Padding Capacitor C_4 in Oscillator Resonant Circuit

(200–500 metres) the oscillator frequency range is 1965–1065 kc/s. By using parallel trimming and series padding capacitors with an inductor of smaller value than that for the signal circuits, the oscillator frequency is higher than that of the signal circuits by a

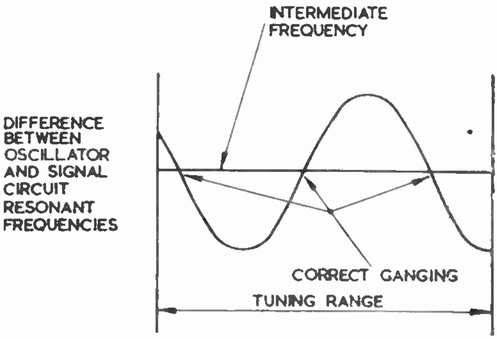


FIG. VIII.6.—Gauging Error Curve

sufficiently constant value if the tuning capacitors are identical (Figs. VIII.5 and 6). The adjustment of the trimming and padding capacitors is known as " alignment ", and is a simple procedure if the component values are properly calculated. The design of these circuits has been investigated by a number of writers, who have published charts and equations giving their results.

The " noise " level of a mixer stage is higher than that of a normal amplifier stage, and it is necessary to apply a signal large in comparison with this noise. The input to a mixer should be at least 20 μV.

8.4. The Signal-frequency Amplifier

This has already been discussed in detail in Chapter 4. It is common to omit this stage from domestic receiver circuits, as a

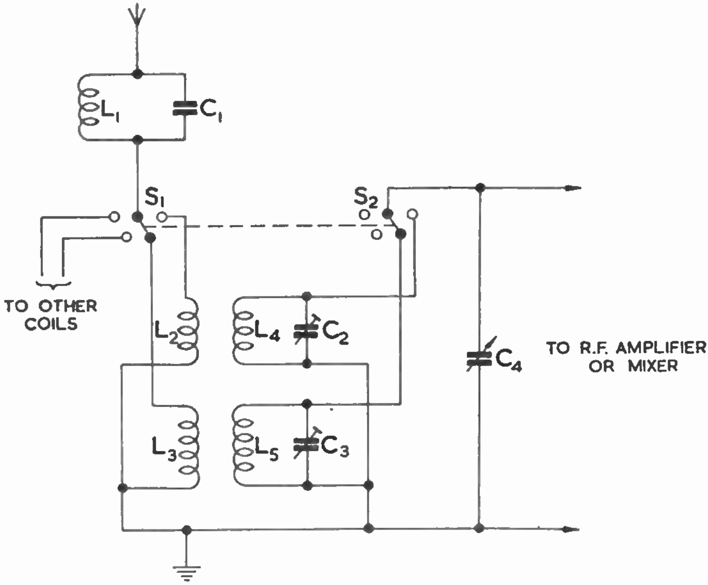


FIG. VIII.7.—Transformer Input Circuits

reasonable signal level is obtainable at the mixer stage from many stations without amplification.

Receivers intended for reception of distant stations, and particularly communications receivers, invariably have one and sometimes two signal-frequency amplifiers. The valves are of the variable-mu type to permit gain control (see Chapter 4).

On medium- and long-wave ranges transformer coupling between the aerial and the first tuned circuit is often used, because a good transfer of signal is possible without damping the tuned circuit, and changes of aerial impedance have a small effect. Fig. VIII.7 shows a typical circuit. The resonant circuit L_1C_1 is tuned to the intermediate frequency, and prevents any signals at that frequency from reaching the receiver. This is known as an "intermediate-frequency rejector". Range switching is carried out by connecting into circuit the appropriate pairs of coils: at the same time the local oscillator coil will be switched.

Short-wave input circuits on domestic receivers for ordinary aerials often use direct capacitance coupling to the first tuned

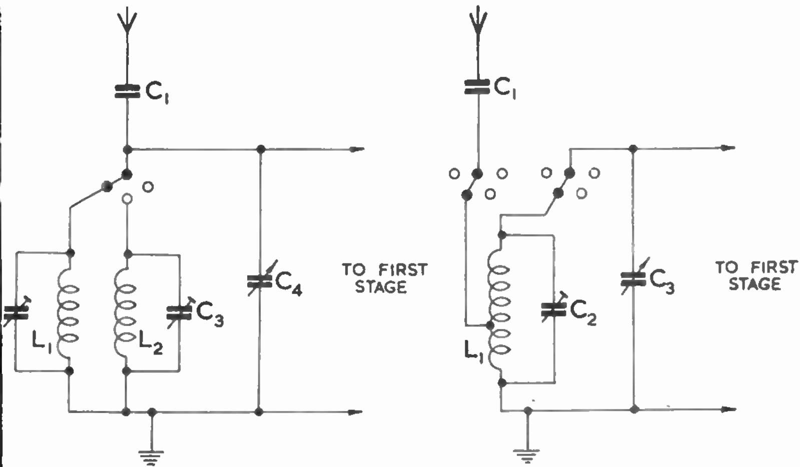


FIG. VIII.8.—Capacitance Input Coupling

circuit or to a tapping on the inductor (Fig. VIII.8), and this method is sometimes used on medium and long waves. In this case the intermediate-frequency rejector circuit is usually a series-tuned circuit (Fig. VIII.9).

Communications receivers with dipole aerial systems (see Chapter 3, Volume I), as used for fixed-frequency stations, use transformer coupling, the transformer ratio being made correct to match the aerial impedance to the dynamic resistance of the first tuned circuit (Fig. VIII.10), and where several known frequencies are to be received, switches or links are provided to connect the appropriate circuits and aerials.

The first tuned circuit is adjusted by a variable capacitor which forms one section of a multi-gang unit, or which in specialised communications receivers may be an individual capacitor to permit more accurate adjustment. A small trimming capacitor is associated with each inductor for alignment purposes, and on some receivers the trimmer C_2 associated with the short-wave ranges is

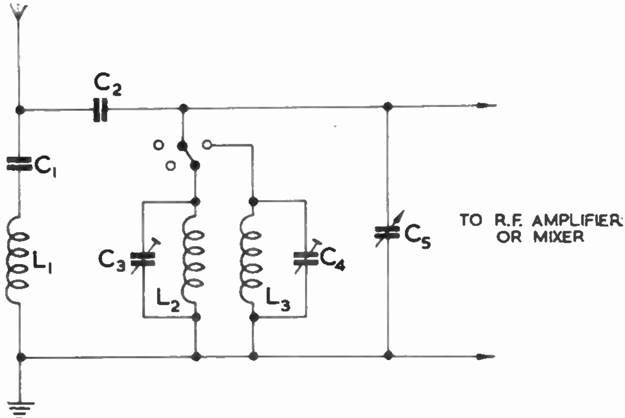


FIG. VIII.9.—Intermediate-frequency Rejector Added to Circuit of Fig. VIII.8

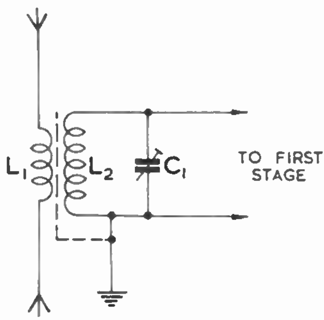


FIG. VIII.10.—Input Coupling for Dipole Aerial

brought out as a manual control, again to obtain the best performance (Fig. VIII.11).

Greater selectivity is obtainable in the input circuits if a band-pass coupling is employed, examples of this circuit being shown in Fig. VIII.12. This type of circuit has been discussed in Chapter **4**; it is sufficient to note here that it is more critical in respect of

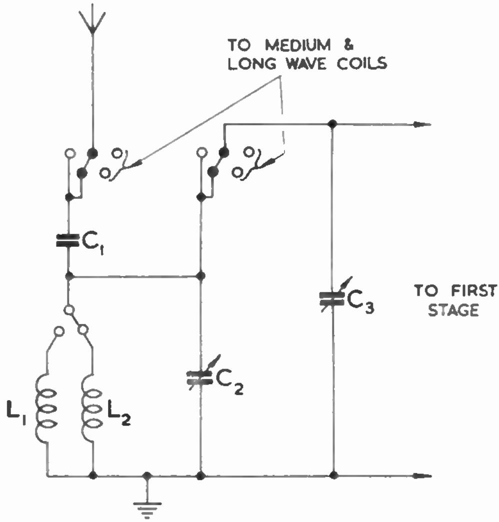


Fig. VIII.11.—Showing Hand Trimmer C_2 for Short-wave Ranges

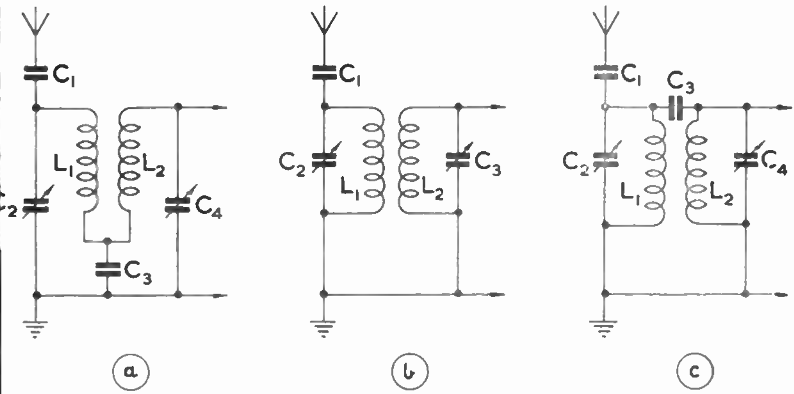


Fig. VIII.12.—Tuneable Band-pass Input Circuits component tolerances and circuit alignment and requires an extra section of the tuning capacitor. It is used only on medium- and long-wave bands, and is now seldom found in superhets, as adjacent-channel selectivity is mainly obtained in the intermediate-frequency amplifier.

8.5. Intermediate-frequency Amplifiers, Detectors and A.G.C. Systems

Intermediate-frequency amplifiers have been discussed in Chapter 4. This part of the receiver again employs variable-mu valves

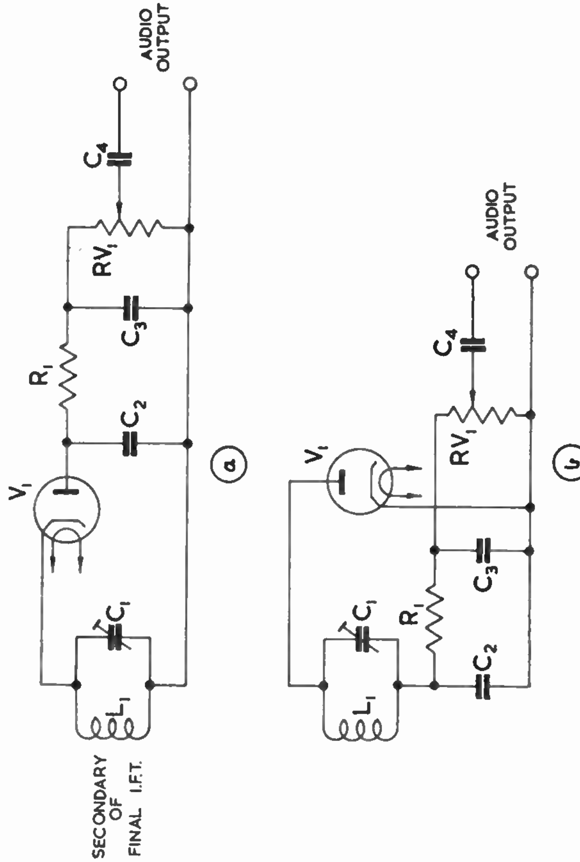


FIG. VIII.13.—Diode Detector Circuits

to permit gain control. The maximum gain of intermediate-frequency amplifier stages varies between 50 and 200, depending on the bandwidth, and the output at intermediate frequency is usually several volts.

The detector is most commonly a diode rectifier, and typical circuits are shown in Fig. VIII.13. The diode is chosen because of its freedom from overloading, and its fidelity when fed with high-level signals as available from the intermediate-frequency amplifier.

A diode detector gives a mean value of direct output proportional to the amplitude of the carrier at the input to the detector. This direct p.d. is modulated with the audio frequency, and its

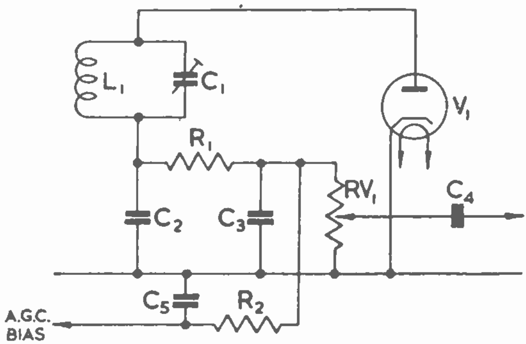


FIG. VIII.14.—Simple A.G.C. Circuit

polarity depends on the diode connexion. If it is made negative with respect to earth, it can be used to regulate the gain of the early stages by feeding it back to the control grids of the variable-mu valves used there. It is necessary to insert a filter to prevent audio signals from affecting the gain, and Fig. VIII.14 shows a simple circuit. Three methods of feeding this control bias to radio-frequency amplifiers are shown in Fig. VIII.15.

A disadvantage of such a simple type of automatic gain control (A.G.C.) is that any signal producing an output from the diode reduces the gain of the receiver by some amount. It is desirable that weak signals, below a pre-determined level, do not affect the gain, and circuits providing this feature are called " delayed A.G.C." circuits, because their action is " delayed " until the input signal exceeds a certain level.

To provide delayed A.G.C., a second diode, which is commonly fed from the anode of the last intermediate-frequency amplifier, is necessary to avoid interfering with the detector circuit. A typical circuit is shown in Fig. VIII.16.

In this circuit V_1 is the final intermediate-frequency amplifier and V_2 is the A.G.C. rectifier. V_1, like all other controlled valves, has its own cathode bias resistor, R_1, to give the correct bias when

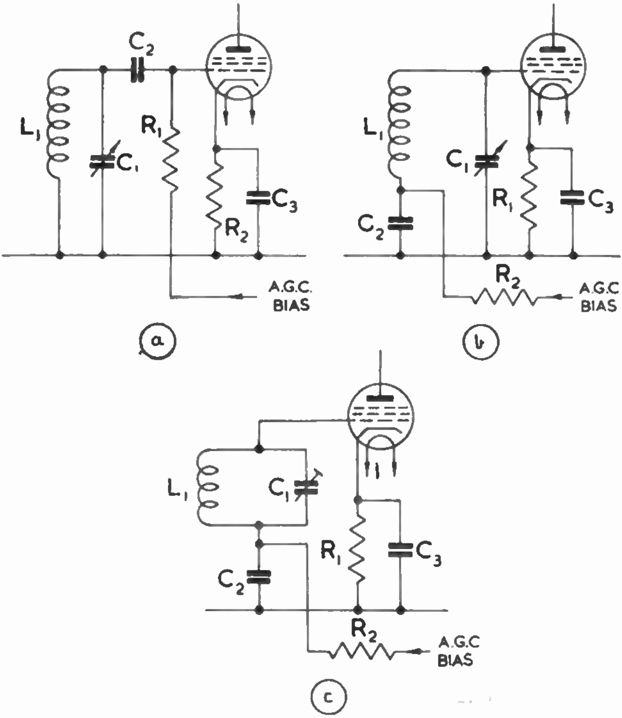


FIG. VIII.15.—Feeding A.G.C. Bias to Variable-mu Amplifiers

A.G.C. is not working. C_4 feeds V_2 from the anode of V_1, and the anode of V_2 is taken to earth via R_4 and R_5. The cathode of V_2 is held positive to earth by R_6 and R_7 across the H.T. supply; this positive potential is sometimes derived from the cathode of a subsequent stage. In the absence of a signal, the anode of V_2 is therefore at a lower potential than its cathode, and the grids of the control valves are held at earth potential by R_2, R_3, R_4 and R_5. This condition continues until a signal whose peak value exceeds the p.d. across R_7 is applied from V_1. V_2 then conducts on the peaks of the signal, so that C_4 charges until the mean anode potential of V_2 is lower than that of its cathode by approximately the peak value of the signal (Fig. VIII.17). The mean potential of V_2 anode relative to earth is then roughly equal to the p.d.

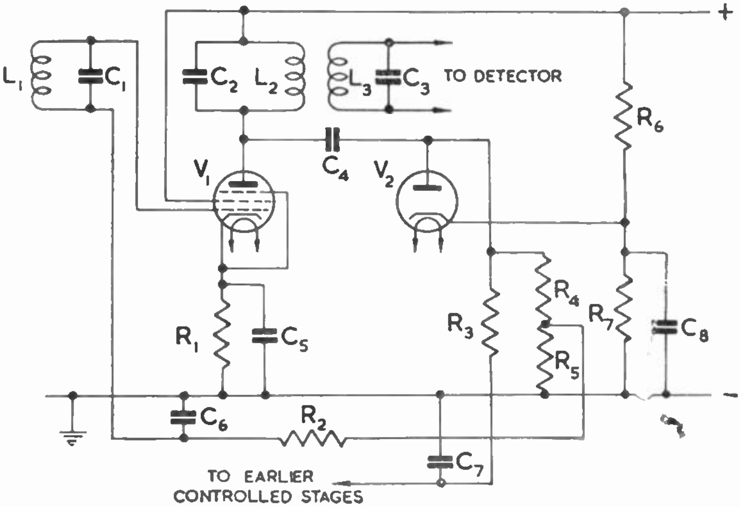


FIG. VIII.16. - A Circuit for Delayed A.G.C.

across R_7 minus the peak value of the signal from V_1 which is a negative quantity, thus biasing the controlled stages and reducing their gain. As the input to the receiver is increased, extra A.G.C. bias is produced by a necessarily increasing signal from V_1, but the increase at that point is very many times less than that of the input. Typical A.G.C. characteristics are shown in Fig. VIII.18.

8.6. Audio-frequency Amplifier

The detector circuit in a superhet receiver delivers a few volts of audio-frequency, which is sufficient to drive a high-gain pentode output valve. Valves of this type are commonly used in small domestic receivers, but they introduce more distortion than power valves of lower gain which are used in receivers designed to give good quality of reproduction. In such cases a stage of amplification must be used between detector and output stages.

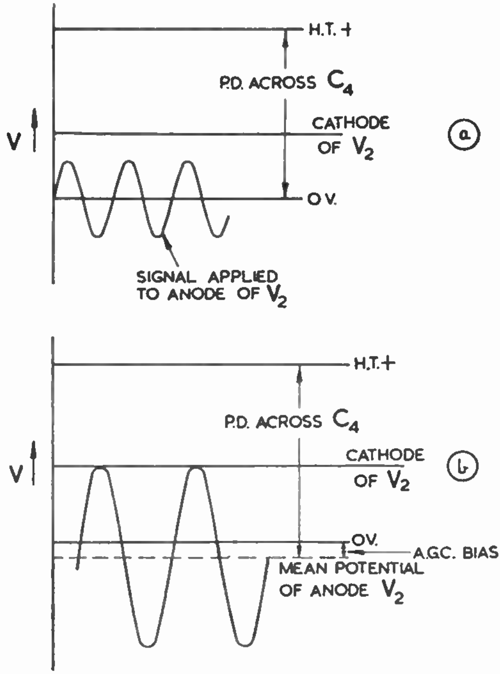


FIG. VIII.17.—Waveforms in Circuit of Fig. VIII.16

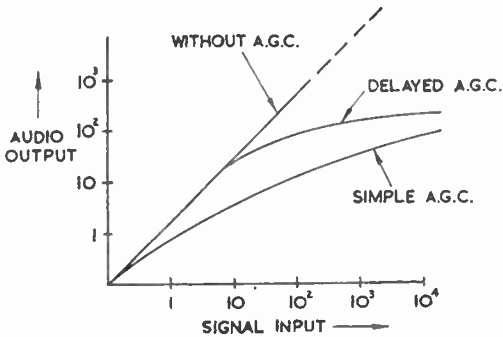


FIG. VIII.18.—Typical A.G.C. Characteristics

Domestic receivers incorporating gramophone equipment require at least one stage of amplification between the pick-up and the output stage, while instruments of high quality generally have a push–pull output stage, and may have a more complicated audio-frequency section with provision for " tone control ".

Communication receivers are usually fitted with a simple low-power output stage for operating headphones and a small loudspeaker. A stage of amplification between detector and output stage is usual to provide a useful output on small signals.

8.7. Alignment

This is the procedure of adjusting the resonant circuits of the receiver to the designed values, and is carried out by setting the various trimming and padding capacitors. Signal generators for this purpose produce a modulated or unmodulated output at will. If a modulated signal is used, an output meter measuring audio-frequency volts across the loudspeaker is a convenient indicator. An unmodulated signal may be used by measuring the direct current in the detector load. The intermediate-frequency amplifier is first adjusted by feeding a signal from a calibrated signal generator to the grid of the mixer valve and preventing the local oscillator and A.G.C. system from working.

The tuning capacitors on the intermediate-frequency transformers are then adjusted to give the greatest possible output, the feed from the signal generator being kept as low as convenient.

To align the signal-frequency amplifier, the local oscillator is put into operation and the signal generator is connected to the aerial terminal of the receiver. Where separate trimmers are provided on each range of tuning and oscillator coils, the ranges can be dealt with in any order, but where some trimmers are common the independent ranges must be adjusted first. The signal generator is set to a frequency near the high-frequency end of the range being aligned, corresponding to a displacement of about 15° from the low-capacitance position of the receiver-tuning capacitor, which is likewise adjusted according to the tuning dial. The parallel trimming capacitors on the oscillator circuit and then on the signal circuits are adjusted for maximum output. The signal generator is then set to a frequency corresponding to a setting of the receiver-tuning capacitor about 15° from the full-mesh position. The signal is then tuned in, and the series padding capacitor in the oscillator circuit is adjusted for maximum response while the receiver-tuning capacitor is " rocked ". This is necessary because no further adjustment of the signal-frequency circuits is possible, and their calibration has already been set by trimming them where the trimmer is a large proportion of the total capacitance (the high-frequency end of the range).

The signal generator and receiver are then re-adjusted to the position at which the first adjustment was made, and if necessary, the parallel trimmer on the oscillator is re-adjusted.

These operations are then repeated on the other ranges of the receiver.

Periodical alignment is necessary during the life of a receiver, because of changes of value of some of the components with the repeated temperature changes they suffer, and sometimes because the adjustments shift with vibrations caused by the loudspeaker.

8.8. Interference

Superhet receivers are prone to interference, which can be prevented only by careful design. One form is known as " image " or " second-channel " interference. The intermediate-frequency cir-

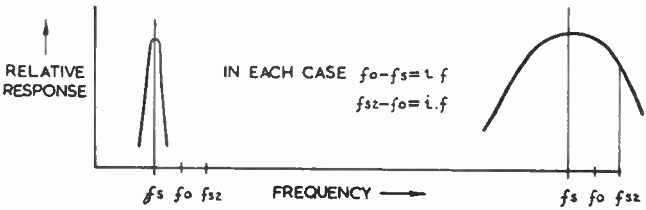


FIG. VIII.19.—Response Curves for Circuits of Similar Q Value at Low and High Radio-frequency

cuits will amplify any signal at intermediate frequency which reaches them, and so any products of the mixing operation carried out by the frequency changer which are of this frequency—say 465 kc/s—will produce an audible output. The desired product of mixing is, of course, that at local oscillator minus signal frequency, but if a signal at local oscillator plus intermediate frequency is present on the signal grid of the mixer this will also appear at the frequency-changer anode at intermediate frequency. Such a signal is known as the image of the desired signal, or the second channel, which when modulated by the local oscillator gives an output at intermediate frequency. It can be rejected by ensuring that the signal-frequency tuned circuits have adequate selectivity, a simple requirement on long and medium waves when the intermediate frequency is 465 kc/s (see Fig. VIII.19). Because bandwidth is proportional to frequency with a given Q value in the tuned circuit, rejection of image signals becomes increasingly difficult at a low value of intermediate frequency, such as 465 kc/s, as the signal frequency rises, as shown in Fig. VIII.19.

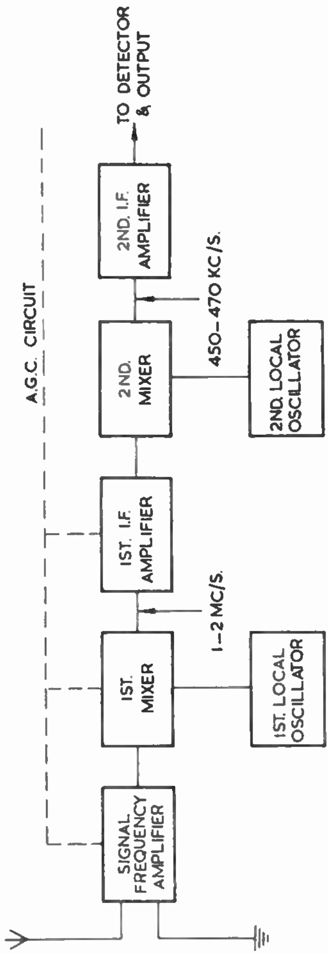


FIG. VIII.20.—Block Diagram of Double Superhet

A method commonly used in communication receivers is to use an intermediate frequency in the range 1–2 Mc/s, 1·6 Mc/s being a common frequency for this purpose. This gives a considerable improvement, and is further assisted by the use of more than one tuned circuit in the signal-frequency stages. The use of a high intermediate frequency brings with it the problem of adjacent-channel interference, and it is usual to change the frequency a second time to 465 kc/s to enable the superior selectivity of circuits at this lower frequency to be obtained. Receivers of this type are known as " double superhets ", and a block diagram is shown in Fig. VIII.20. They have the additional advantage of permitting a very high total gain, as the amplification is obtained at three different frequencies and instability problems are reduced.

Interference caused by direct reception at intermediate frequency is prevented by the use of intermediate-frequency rejector circuits already mentioned and shown in Figs. VIII.7 and 9.

Interference heard as a whistle is also likely, and is caused when signals within the pass band of the intermediate-frequency amplifier, but differing in frequency by a few kc/s, are produced by the mixer. One such signal may be desired, but the other is usually a spurious product of mixing, such as between two transmitters whose frequencies differ by approximately the intermediate frequency of the receiver.

Questions

1. Outline the design and construction of a three-stage intermediate-frequency amplifier for a superheterodyne receiver, the mid-band frequency being 465 kc/s and the bandwidth suitable for: (*a*) speech, and (*b*) continuous-wave telegraph signals. Indicate the precautions taken to avoid instability likely to lead to self-oscillation. (C. & G., 1947.)

2. Discuss the relative merits of straight and superheterodyne receivers suitable for the reception of radio telephony in the band 5–25 Mc/s.

State the factors determining the selectivity desirable in (*a*) the radio-frequency circuits and (*b*) the intermediate-frequency circuits of a superheterodyne receiver. (C. & G., 1947.)

3. What is the difference between second-channel interference and adjacent-channel interference in a superheterodyne receiver? State clearly what factors determine each of these types of interference.

4. Explain the superheterodyne principle. Point out the advantages of using an intermediate frequency and the factors which influence the choice of a suitable value.

Why is it necessary for the oscillator and input circuits to have resonant frequencies differing by a constant amount, and what methods are used in practice to obtain it?

Upon what factors do the selectivity and sensitivity of a superheterodyne receiver depend? (C. & G., 1946.)

5. What is the purpose of delayed automatic gain control in a receiver? Explain how it is obtained. (C. & G., 1946.)

6. Give the circuit and describe the action of a triode-hexode frequency changer for a superheterodyne receiver.
(C. & G., 1948.)

7. Describe one form of automatic gain control for use in a short-wave receiver, indicating by a sketch the relation between the radio-frequency input and the audio-frequency output.
(C. & G., 1949.)

8. A superheterodyne receiver is required to cover the waveband 200–500 metres, and the intermediate frequency is to be 450 kc/s. Calculate:

(*a*) the maximum and minimum frequencies of the beating oscillator;

(*b*) the maximum capacitance of the oscillator tuning capacitor assuming that the minimum value is 100 $\mu\mu$F.
(C. & G., 1949.)

Answer: (*a*) 1,950 kc/s, 1,050 kc/s; (*b*) 344·9 $\mu\mu$F.

9. Explain, with the aid of a circuit diagram, how a valve having a square law anode-current/grid-potential characteristic could be used as a frequency-changer. (C. & G., 1949.)

10. Why is it difficult to obtain both good selectivity and high fidelity in a broadcast receiver? Are these qualities more readily obtained in superheterodyne receivers than in straight receivers, and if so, why? (C. & G., 1951.)

11. A tuned circuit comprises a 1,000-μH inductor of series resistance 30 ohms in parallel with a 100-$\mu\mu$F capacitor. An e.m.f. of 0·001 volt (r.m.s.) at the resonant frequency of the tuned circuit is induced in series with the inductor. What is the voltage developed across the tuned circuit? What additional resistance would be required to make the magnification factor of the circuit equal to 100? (C. & G., 1951.)

Answer: 0·105 volt, 1·62 ohms.

12. Give a brief explanation of the following: (*a*) Image sideband; (*b*) Miller effect. (C. & G., 1944.)

13. Explain the need and method of application of automatic gain control in a broadcast receiver. Give some approximate idea of the output level of such a receiver for varying input levels.
(C. & G., 1940.)

14. Give a brief explanation of the meaning and cause of three of the following effects: (*a*) side-band interference; (*b*) motor boating; (*c*) Miller effect; (*d*) second-channel interference. (C. & G., 1940.)

15. Describe a method whereby the audio-frequency output of a broadcast receiver is automatically controlled to compensate for variations in the level of the received signal due to fading. (I.E.E., November 1933.)

16. What are the advantages of A.G.C. when applied to a radio-telephone receiver? How does the time of operation of the control affect the performance of the receiver? Give a diagram of a control system suitable for a high-class broadcast receiver. What can be done to avoid excessive noise when tuning between stations? (C. & G., 1934.)

17. Under what circumstances is A.G.C. desirable in a radio receiver? Describe, with the aid of a circuit diagram, a method of obtaining delayed A.G.C. in a superheterodyne receiver. (L.U., 1940.)

18. State the precautions to be observed in ganging the radio-frequency amplifier and oscillator sections of a superheterodyne receiver.

What do you understand by: (*a*) padding; (*b*) trimming; (*c*) tracking? (C. & G., 1952.)

19. The anode current of a hexode valve can be represented approximately by:

$$i_a = kv_1v_2$$

where v_1 and v_2 are the potentials at the two grids and k is a constant. If v_1 and v_2 are sinusoidal in form but are of different frequencies f_1 and f_2, explain how such a valve can be used as a frequency-changer, and deduce an expression for the component of the anode current of frequency $f_1 - f_2$. (C. & G., 1953.)

20. Describe, with a block schematic diagram, the essential features of a "communications" type of receiver suitable for amplitude-modulation telegraphy and telephony in the frequency range 150 kc/s to 30 Mc/s. (C. & G., 1953.)

21. Describe the method of use of a cathode-ray oscilloscope and associated equipment for the alignment of the tuned circuits of a radio receiver. (C. & G., 1953.)

22. Describe a simple circuit for the automatic gain control of a radio receiver and illustrate your answer by a curve of the control characteristic. (C. & G., 1953.)

23. Describe one method whereby an intermediate-frequency amplifier stage can be made to have a substantially uniform response over a band of frequencies 10 kc/s wide.

State the factors in the design of the circuit that determine: (*a*) the width of the pass-band, and (*b*) the gain of the stage.

(C. & G., 1953.)

Specimen Answer

Q. Explain with a diagram the operation of the triode-hexode frequency-changing stage in a superheterodyne receiver. What are the advantages and disadvantages of the triode-hexode valve for this purpose?

A. The triode-hexode valve used (see Fig. VIII.Q.1) consists of a triode and a hexode, with a common cathode and heater, enclosed

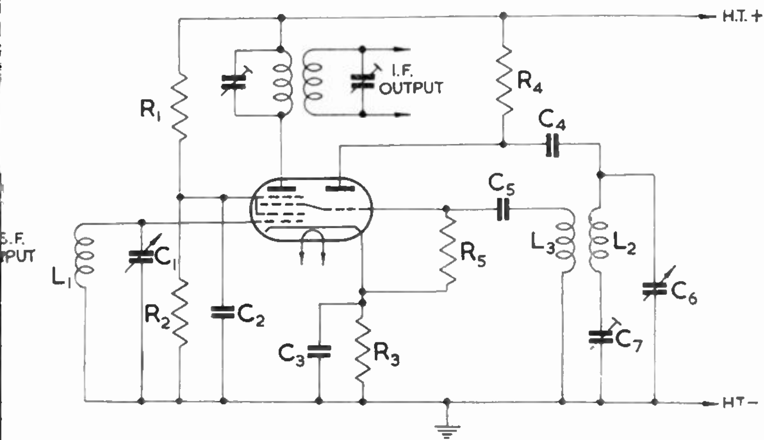


Fig. VIII.Q.1

within the same evacuated envelope. A direct internal connexion is made between the control grid of the triode and the third grid of the hexode. The triode section operates as a tuned anode oscillator, its tuned circuit being L_2C_6 coupled to the grid circuit via L_3 and C_5. The triode anode is shunt fed via the anode resistor R_4, C_4 being the coupling and blocking capacitor.

The radio-frequency input signals are applied to the control grid of the hexode via the tuned circuit L_1C_1. The hexode grids 2 and 4 are maintained at a positive potential by the voltage divider R_1R_2, decoupled by C_2, the valve acting as a screened-grid amplifier. The screens also isolate the oscillator from the signal and intermediate-frequency circuits. The anode load of the hexode consists of the band-pass-tuned transformer feeding the first intermediate-frequency stage. This intermediate-frequency transformer selects the required band of frequencies from the modulation products resulting from the injection of the local oscillator frequency into the hexode section.

Automatic gain control is normally applied to the hexode control grid, but this is not shown in Fig. VIII.Q.1.

The principal advantages of the triode-hexode are :

(*a*) Inclusion of mixing and oscillating sections within the same envelope. This enables an electrode arrangement with internal screening to be employed, which gives efficient operation at high frequencies, and thus overcomes the disadvantage of a separate oscillator required by other frequency-changing valves when used for short-wave reception.

(*b*) The hexode portion has variable-mu characteristics between the signal grid and anode so that gain control may be easily applied. The mutual conductance between the hexode grid 3 (to which the local oscillator signal is applied) and the anode is not variable, however, so that a small heterodyne signal is sufficient for full modulation of the anode current.

(*c*) The high mutual conductance of the triode section ensures satisfactory operation at high frequencies.

(*d*) Interaction between signal and oscillator circuits is negligible because of the electrode arrangement and internal screens.

(*e*) The signal-to-noise ratio is greatly improved by the amplification of the signal in the hexode section before modulation with the heterodyne signal.

The principal disadvantage is that the input capacitance of the hexode section varies with the bias on the signal grid. This, however, is a defect inherent in all radio-frequency amplifiers, and may be minimised by suitable circuit arrangements.

(P.O. Eng. Dept.)

CHAPTER 9

MEASUREMENTS

In this chapter the measurement of resistance, capacitance and inductance is discussed. Methods of measurement of the Q factor of circuits are mentioned, and a brief description is given of the cathode-ray oscilloscope and its use for various measurements. The measurement of aerial-impedance characteristics is discussed.

9.1. Measurement of R, C and L at Low Frequencies

In radio work it is necessary to measure resistance, capacitance and inductance to within 5 per cent. or better. This requires the use of a bridge network employing a known standard component. An example is the well-known " Wheatstone Bridge " for measurement of resistance (Fig. IX.1).

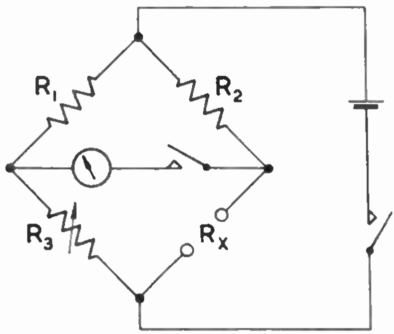


FIG. IX.1.—Wheatstone Bridge

The general balance conditions for all types of bridge, whether they be a.c. or d.c., is that the product of the impedances of the opposite arms of a bridge are equal in magnitude and phase.

In Fig. IX.1 $R_1R_x = R_2R_3$ (all arms being non-reactive)

or
$$R_x = \frac{R_2R_3}{R_1}$$

Measurement of inductance and capacitance is conveniently carried out with a bridge supplied by some form of a.c. generator, *e.g.*, an oscillator. The measurements are generally divided into two main classes :

(*a*) audio- or low-frequency measurements (20 c/s–20 kc/s) ;
(*b*) radio- or high-frequency measurements (20 kc/s upwards).

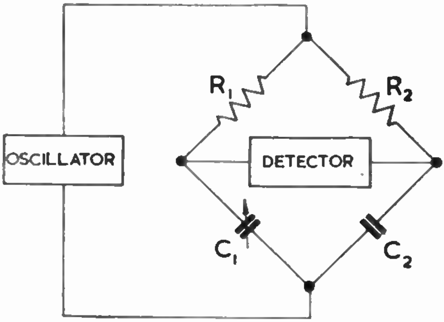


FIG. IX.2.—Simple a.c. Capacitance Bridge

Fig. IX.2 shows a bridge for the measurement of capacitance at audio-frequency. Capacitor C_1 may be a calibrated standard capacitor which is varied to obtain balance against C_2, the unknown to be measured.

At balance $$\frac{R_1}{j\omega C_2} = \frac{R_2}{j\omega C_1}$$

$$\therefore \quad \frac{R_1}{C_2} = \frac{R_2}{C_1}$$

and if C_2 is the unknown, $$C_2 = \frac{R_1 C_1}{R_2}$$

If C_1 is variable, it is usual to make $R_1 = R_2$, so that $C_1 = C_2$ at balance. The range of the bridge may be extended by adding additional known values of capacitance in parallel with C_1. Many simple bridges of this type have variable ratio arms (Fig. IX.3), and several known values of capacitance can be connected at C_1 to cover a wide range—say 100 μF to 100 pF. Although it is possible to apply the above technique for the measurement of inductance, this is not usually done because of the difficulty of producing suitable variable inductors. It is usual to balance the positive reactance of the unknown inductor with the negative reactance of a calibrated standard capacitor in the opposite arm of the bridge. The best-known bridge of this type is due to Maxwell, and is known as the "Maxwell Inductance Bridge" (Fig. IX.4). Resistors R_2

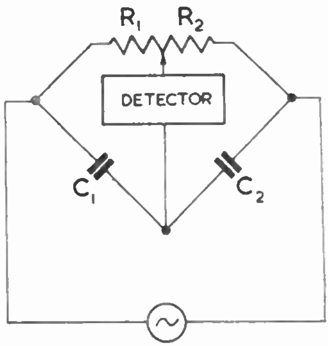


FIG. IX.3.—Simple a.c. Bridge

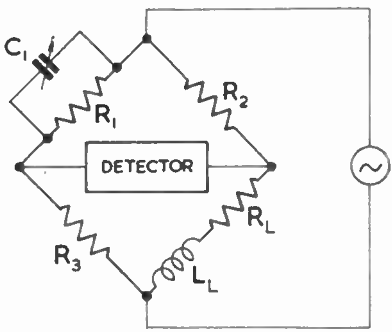


FIG. IX.4.—Maxwell Inductance Bridge

and R_3 are equal, and the bridge is balanced by adjusting C_1 and R_1, the latter being necessary because of the loss in the effective resistance of the inductor. The resultant of R_1 and C_1 in parallel is given by

$$Z_1 = \frac{R_1 \times \dfrac{1}{j\omega C_1}}{R_1 + \dfrac{1}{j\omega C_1}} = \frac{R_1}{j\omega C_1 R_1 + 1}$$

Therefore at balance $\dfrac{R_1}{j\omega C_1R_1 + 1} \times (R_L + j\omega L_L) = R_2R_3$

$$\therefore \quad R_1R_L + R_1j\omega L_L = R_2R_3(j\omega C_1R_1 + 1)$$
$$= j\omega C_1R_1R_2R_3 + R_2R_3$$

Equating the real terms we have

$$R_1R_L = R_2R_3$$
$$\therefore \quad R_L = \frac{R_2R_3}{R_1}$$

and equating the imaginary terms we have

$$R_1j\omega L_L = j\omega C_1R_1R_2R_3$$
$$\therefore \quad L_L = C_1R_2R_3$$

Unless very accurate results are required and/or the measurements are made with a low input level from the oscillator, no special precautions are required for audio-frequency bridges. The input to, and the output from, the bridge is usually via transformers having unity turns ratio; the oscillator may be any fixed or variable audio-frequency oscillator, depending on the conditions required. The detector may be head-phones (if the frequency is within say 100–3,000 c/s), an amplifier whose output is rectified and fed to a voltmeter or an oscilloscope, balance being obtained in each case by adjusting for minimum output from the bridge. Each of the bridges described above is independent of frequency.

9.2. Radio-frequency Measurements

Rough measurements of capacitance and inductance at radio-frequency are often made by the "Tuned-circuit substitution

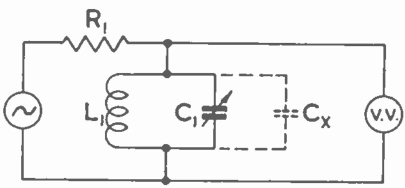


FIG. IX.5.—Circuit for Substitution Method

Method". Fig. IX.5 shows a parallel resonant circuit fed from a high-impedance source (an oscillator with added resistor R_1). The variable capacitor C_1 is a calibrated component, while the inductor L_1 is chosen to resonate with C_1 at a frequency obtainable from the oscillator. A valve-voltmeter is connected across the tuned circuit.

If an unknown capacitor is to be measured, C_1 is first adjusted near its maximum capacitance and the oscillator frequency is adjusted until resonance is obtained, as indicated by a maximum reading on the valve-voltmeter. The unknown capacitor C_x is then connected in parallel with C_1, which is reduced in value until resonance is obtained at the same frequency as before. The unknown capacitance is then given directly by the difference between

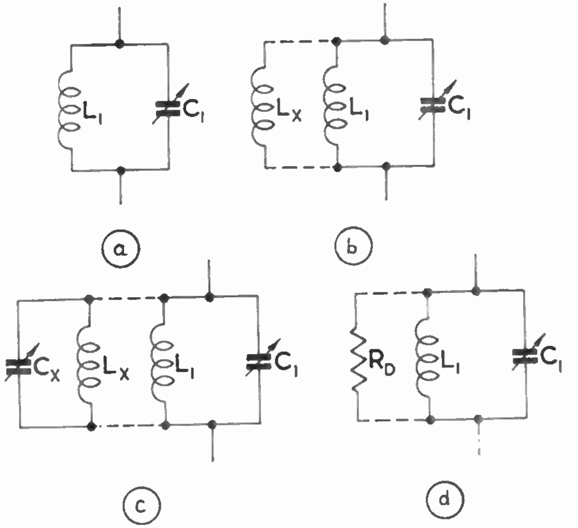


FIG. IX.6.—Measurement of Inductance by Circuit of Fig. IX.5

the two readings of C_1. The frequency need not be known for this measurement, but it must be constant, while, of course, C_1 must have a greater range of adjustment than any value to be measured.

Consider now a resonant circuit L_1C_1 as shown in Fig. IX.6(*a*). Suppose that an inductor L_x is connected in parallel with L_1 as shown in Fig. IX.6(*b*). The resultant of L_1 and L_x in parallel is less than either L_1 or L_x, so that the resonant frequency of the combination is increased. Now let a capacitance C_x be placed in parallel with L_x so that the resonant frequency of L_x and C_x is the same as that of L_1 and C_1 as shown in Fig. IX.6(*c*). This may also be represented by Fig. IX.6(*d*), where R_D represents the dynamic resistance of L_xC_x at their resonant frequency. The resonant frequency of the combination is now restored to that of L_1 and C_1. Returning to Fig. IX.6(c), the capacitance C_x may be provided by increasing the value of C_1. If the frequency is the same for all the diagrams in Fig. IX.6, then from $\omega^2 = \frac{1}{LC}$ we have

$$L_x = \frac{1}{\omega^2 C_x}$$

where $\omega = 2\pi \times$ frequency in c/s; L_x = inductance in henrys; C_x = capacitance in farads.

In this case the measurement is made by adjusting C_1 to a low value and tuning the oscillator for resonance (Fig. IX.5). The unknown inductor is connected in parallel with the resonant circuit and C_1 increased until resonance occurs. If C_x = change of capacitance necessary after connecting L_x, then $L_x = \frac{1}{\omega^2 C_x}$. The frequency of the applied p.d. must be known in this case.

If the unknown inductor L_x is small compared with L_1 it may be impossible to adjust C_1 to resonate with the parallel combination of L_1 and L_x. In that case L_x should be connected in series with L_1. The inductive reactance of the circuit is then increased, and for resonance of the combination to occur at the same frequency as that of L_1C_1, the capacitive reactance must also be increased, or the value of capacitance reduced. The same equation as for the parallel case applies, but in this instance C_1 should be initially at a high value. In all these measurements it is essential that there is negligible mutual coupling between the two coils L_1 and L_x.

The measurement of Q may also be made with the arrangement shown in Fig. IX.5. In Chapter 4 it was stated that the band-width $2\delta f$ of a tuned circuit is defined as the frequency range between the points where the response is $\frac{1}{\sqrt{2}}$ of the maximum and that $2\delta f = \frac{f_r}{Q}$, where f_r is the resonant frequency and $Q = \frac{\omega L}{r}$ = ratio of reactance to effective series resistance. The oscillator frequency is adjusted so that resonance is obtained, and readings of the valve-voltmeter and oscillator frequency are noted. The oscillator frequency is then adjusted to the two values, one above and one below resonance, at which the valve voltmeter reading is $\frac{1}{\sqrt{2}}$ times the reading at resonance, and the frequency at each of these points is noted (see Fig. IX.7). Q is then given by $Q = \frac{f_r}{2\delta f}$, where f_r is the resonant frequency and $2\delta f$ is the difference between the two frequencies at the $\frac{1}{\sqrt{2}}$ points. For this measurement the series resistor R_1 and the input impedance of the valve-voltmeter should be high compared with the dynamic resistance of the tuned

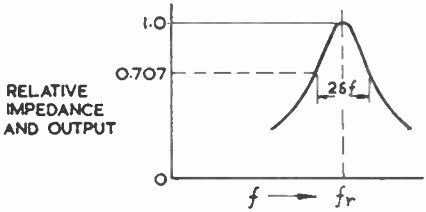


FIG. IX.7.—Pertaining to "Q"

circuit and the oscillator output p.d. should remain constant. The accuracy of this measurement is low, because $2\delta f$ is a small difference between two large quantities and is not determinable with precision from the oscillator frequency scale.

9.3. The Q-Meter

The Q value of an inductor is most satisfactorily measured on a "Q-meter" designed for the purpose. One type consists basically of a low-impedance signal source, a valve-voltmeter and a low-loss

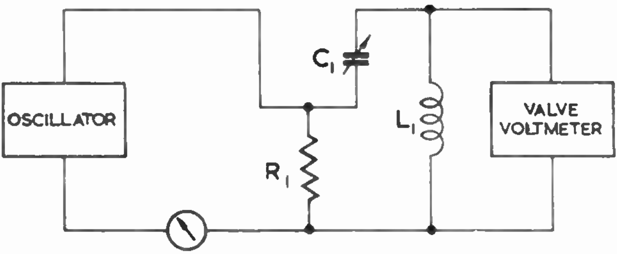


FIG. IX.8.—Circuit of Q-meter

variable capacitor to form a resonant circuit with the inductor under test. The circuit is shown in Fig. IX.8. The radio-frequency oscillator supplies a current which develops a p.d. across a low resistor R_1, and is indicated by a thermal ammeter. Typical values are 0·5 amp. and 0·04 ohm, so that the p.d. across R_1 is 0·02 volt. Capacitor C_1 is then adjusted to resonate with inductor L_1 at the applied frequency. At resonance the p.d. across the inductor is a maximum and is indicated by the reading of the valve-voltmeter. Now capacitor C_1 and inductor L_1 are effectively a series resonant circuit fed from a known p.d. V_1. The impedance, therefore, of L_1 and C_1 in series is the effective resistance r_1 of inductor L_1 (assuming C_1 is perfect). The current in L_1 and C_1 is therefore $\frac{V_1}{r_1}$, and the p.d. across L_1 is $\frac{V_1}{r_1} \times \omega L_1 = V_1 \times Q$. If V_2 is the reading of the valve voltmeter $\frac{V_2}{V_1} = Q$. The ammeter is usually marked with a standard value of current which develops a known p.d. across R_1, and the valve voltmeter can then be calibrated directly in terms of Q. The oscillator covers a wide range of frequencies to enable measurements to be made on radio-frequency coils at their working frequency, and the method of measurement is satisfactory above about 50 kc/s. Below this frequency it is usual to use the Maxwell bridge to measure Inductance and Effective Resistance, and hence to calculate Q.

9.4. Radio-frequency Bridge Measurements

Bridge measurements may also be made at radio-frequency, but the form of the bridge is often different from low-frequency types.

In radio-frequency bridges the following precautions are usually observed :

1. The connexions from source to bridge and from bridge to detector are made by using either a balanced and screened pair or a coaxial cable, the screen in both cases being connected to earth potential.
2. To provide screening from external fields the bridge is housed in a metal case.
3. The leads from oscillator to bridge and bridge to detector are as short as possible.
4. The connexions to the unknown are as short as possible, as they might otherwise offer appreciable reactance.
5. The bridge input and output are connected via transformers whose primary and secondary windings are screened from each other.
6. The transformers are designed to operate satisfactorily over a band of radio frequencies.

Fig. IX.9 shows a typical radio-frequency impedance bridge which has one side earthed. This is known as an unbalanced bridge.

The variable arm R_3 and the differential capacitor $C_a + C_b$ are calibrated to give the value of the unknown component connected across R_4 in terms of parallel resistive and capacitive components.

When measuring inductive impedances on such a bridge the frequency must be known because the impedance is given in terms of

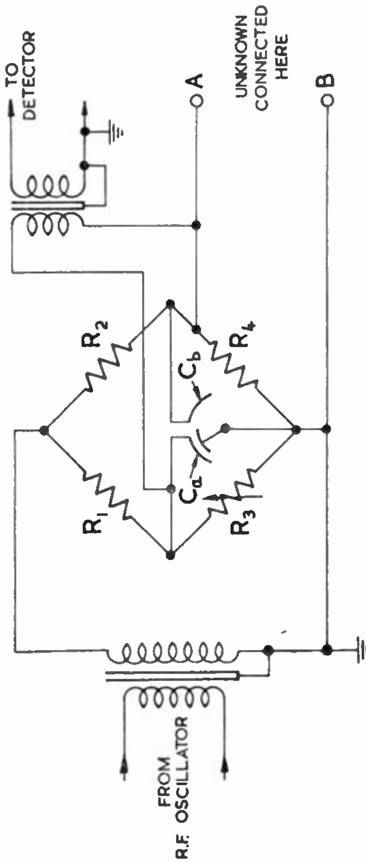


FIG. IX.9.—Radio-frequency Impedance Bridge

capacitance (C_x). Conversion to inductance is by the equation $L_x = \dfrac{1}{\omega^2 C_x}$.

A bridge of this type is suitable for the measurement of aerial characteristics, which must be accurately known in order to design efficient coupling circuits between transmitter power amplifier and aerial. Practical aerial characteristics differ from calculated

G

values because of the influence of other masts, aerials, guy wires, etc., in the immediate vicinity, so that actual measurements must be made " on site ". The impedance that is important is the one into which the transmitter works, and this is termed the " driving-point impedance " of the aerial. In Volume I it was shown that

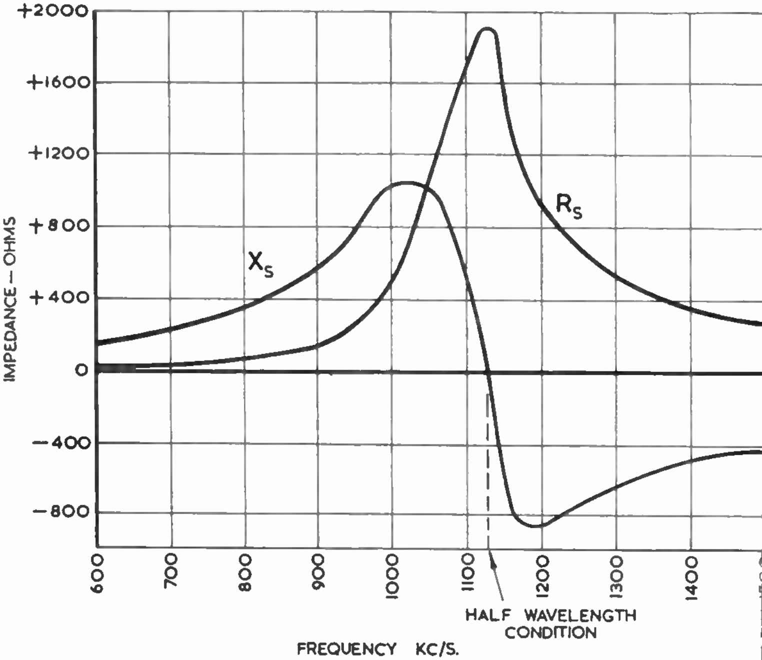


FIG. IX.10.—Characteristics of Dipole Aerial

the aerial impedance was a pure resistance when the effective height of the aerial was a quarter or half wavelength. Further, it was shown how the frequency of operation and the effective height of the aerial were related so that alteration of one affected the other. With a fixed height of aerial it is possible to plot the impedance-frequency characteristic of the aerial by varying the frequency and taking measurements of the aerial driving-point impedance. These measurements are made by the use of a radio-frequency bridge.

Figs. IX.10 and 11 show the results of plotting actual bridge measurements to a base of frequency which extends, in this case, over the medium-wave band. R_s and X_s are the equivalent series resistance and reactance respectively.

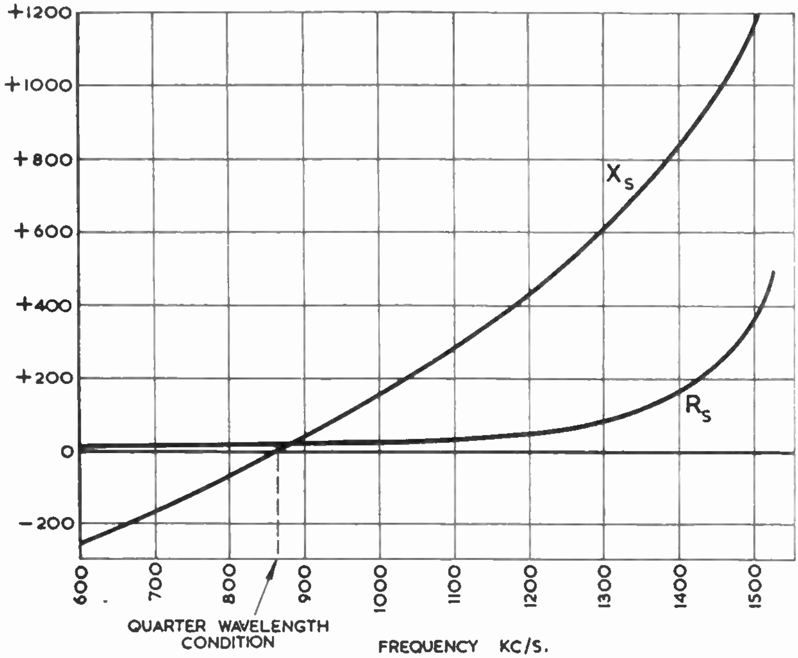


FIG. IX.11.—Characteristics of Quarter-wavelength Aerial

9.5. The Cathode-ray Oscilloscope

The Cathode-ray Oscilloscope (C.R.O.) consists of a cathode-ray tube and associated apparatus in a single unit. In addition to the tube a C.R.O. contains an amplifier for increasing the amplitude of applied signals, a means of producing a deflexion of the spot proportional to time and power supplies (Fig. IX.12).

The scanning or " time-base " generator is used to cause a repeated linear traverse of the fluorescent spot from left to right at a controllable rate, and comprises a circuit developing a periodic p.d. of " sawtooth " form (Fig. IX.13). This p.d., amplified if necessary, is applied to the deflector plates causing horizontal or " X " deflexion of the spot.

Application of another periodic signal to the vertical or "Y" deflector plates causes it to be delineated as an ordinate with time

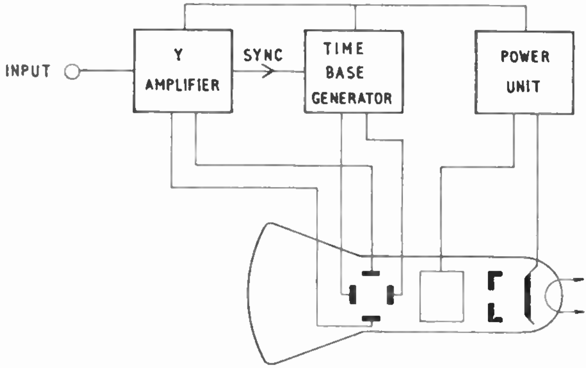


FIG. IX.12.—Block Diagram of Cathode-Ray Oscilloscope.

as the abscissa on the face of the tube, and a stationary picture is obtained if the periodicity of the X and Y p.d.s are integrally related. In order to examine unknown signals of small amplitude an amplifier is provided whose output is fed to the Y plates.

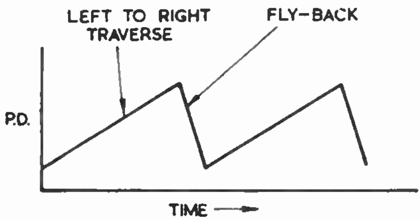


FIG. IX.13.—Form of Scanning p.d.

9.6. The C.R.O. as a Voltmeter

The deflexion of the electron beam of a cathode-ray tube is proportional to the applied potential, whether it be alternating or direct. It is therefore possible, by means of a mask covering the face of the tube and divided into small squares (usually millimetres), to measure the value of an applied potential, if the sensitivity of the tube is known. The sensitivity of a tube is the ratio between p.d. and corresponding deflexion for a p.d. applied directly to the plates, and is stated by the manufacturer. A typical value is 30 volts/cm. This means that a deflexion of 1 cm. is observed for a change in applied potential of 30 volts (Fig. IX.14).

In order that this sensitivity may be checked periodically, it is

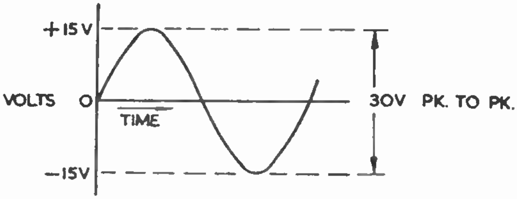


FIG. IX.14.—Illustrating Volts Peak to Peak

usual to provide an output of known value from the mains transformer within the C.R.O.

For conversion to r.m.s. values the peak-to-peak value of a sine wave is halved and then divided by $\sqrt{2}$ (or alternatively multiplied by 0·707). Therefore a signal of 30 volts peak–peak is equivalent to $\frac{30}{2} \times 0{\cdot}707 = 10{\cdot}6$ volts r.m.s.

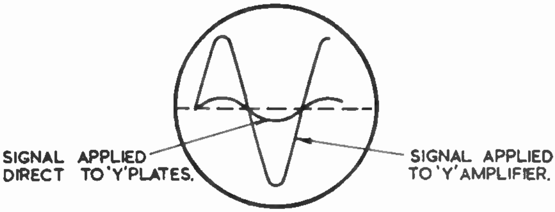


FIG. IX.15.—Use of Amplifier

Fig. IX.15 shows the display of a sine wave on the face of a C.R.O. The sensitivity of the C.R.O. may be increased by passing the signal through an input amplifier (*e.g.*, the " Y " plate amplifier), thus amplifying the signal by the gain of the amplifier and increasing the size of the displayed waveform.

Some C.R.O.s produced commercially provide a feature known as a " double beam ", which provides two oscilloscope displays on one tube. A typical range of frequency covered by a C.R.O. amplifier is from 20 c/s to 1 Mc/s.

The load presented by a C.R.O. is of the order of 2 megohms in parallel with a capacitance of 20–40 pF.

9.7. Tracing Distortion

The C.R.O. is used for the examination of waveforms at various points throughout a piece of equipment. It is thus possible to trace distortion in audio- and radio-frequency amplifiers, detectors,

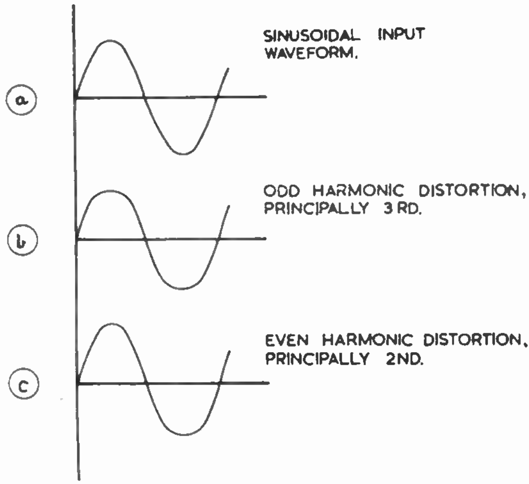


FIG. IX.16.—Harmonic Distortion

oscillators, etc. Systematic checking, stage by stage, from input to output will expose any distortion present, and the offending stage may be further investigated. By the use of a " double-beam C.R.O." the input waveform may be applied to one set of plates whilst the signal to be checked is applied to the other set. Provision is made for the two waveforms to be made coincident, and small amounts of distortion are then easily detected. Fig. IX.16 shows an input waveform and the resultant output waveforms obtained with definite types of distortion. Fig. IX.16(*b*) shows the type of waveform produced by " overloading " a valve, producing " flattening " of the waveform while both halves of the wave are equal in amplitude. Fig. IX.16(*c*) shows the type of waveform produced by overbiased triode valves where the two halves of the waveform are unequal in amplitude. In both these cases the type of distortion referred to is Harmonic Distortion—*i.e.*, the waveform is affected in some way, but the fundamental frequency remains unaltered.

Another form of distortion is known as frequency distortion.

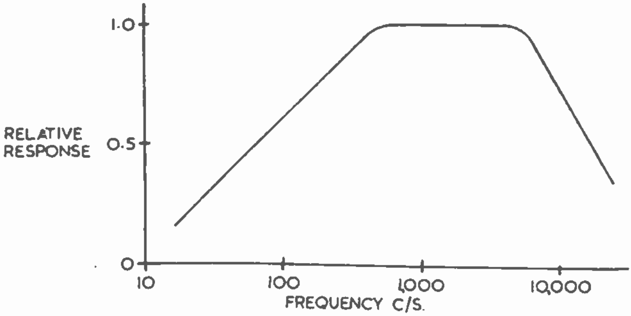


FIG. IX.17.—Frequency Distortion

This type of distortion is characterised by the equipment concerned reproducing frequencies at the output at different levels compared with those at the input. Fig. IX.17 illustrates this type of distortion, which is produced at low frequencies by the falling impedance of transformers, chokes, etc., and at high frequencies by stray capacitance in wiring, valve electrodes, and can be plotted by measuring input and output amplitudes and calculating the gain over the desired range of frequencies.

The additional equipment required to carry out full tests on amplifiers includes a multi-range meter to measure valve potentials and currents, and an audio-frequency oscillator.

Fig. IX.18 shows the manner in which tests on an audio-frequency amplifier may be carried out over a range of frequencies and the points to which the C.R.O. should be connected in the various circuits to examine intermediate stages if the output waveform is observed to be distorted. The waveform is examined by connecting the C.R.O. to the following points in order :

1. Across the primary of the input transformer (this checks the input waveform to the amplifier).
2. Across the input transformer secondary—*i.e.*, the grid input of the first stage.
3. Between anode of V_1 and earth—this checks the first stage and shows the input to V_2.

4. Between grid and earth and anode and earth of the two remaining stages and finally across the output terminals of the output-transformer secondary.

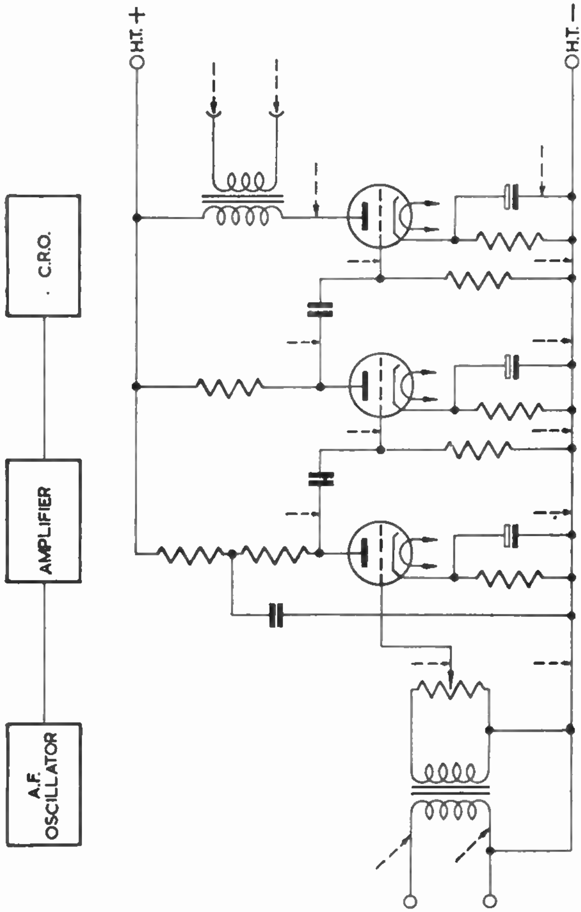
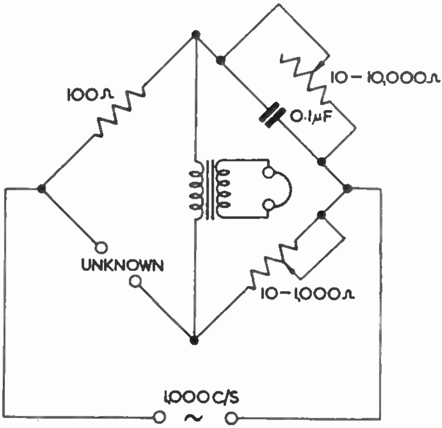


FIG. IX.18.—Investigation of Performance of Audio-frequency Amplifier

The gain of each part of the circuit may be measured with the same arrangement by noting the reading in volts of the signal at each point. Then the gain may be expressed as "n times", where n is the number of times the amplified signal is greater than the input signal and a gain/frequency graph may be plotted.

Questions

1. The circuit diagram shows a Maxwell Bridge for the measurement of inductance. With the values shown, calculate the range



of inductance and Q measurements possible. What feature of such a bridge would make it somewhat difficult to operate? Show how a cathode-ray-tube circuit could be used to replace the headphones as a null indicator.

(Brit. I.R.E., Radio Measurements, May 1946.)

2. You are provided with the following apparatus: a fixed resistor of 1,000 ohms, a variable resistor of 10–1,000 ohms, a variable resistor of 1,000–50,000 ohms, a 1-μF capacitor, a 1,000-c/s oscillator and a pair of telephones. Show how these can be arranged for the measurement of: (*a*) capacitance and power factor, (*b*) inductance and Q. Determine the ranges of measurement in each case.

(Brit. I.R.E., Electronic Measurements, May 1951).

3. What is meant by the Q-factor of a coil?

Describe one method of measuring the Q of a coil at radio frequency. The Q of a coil at a frequency of 1,000 kc/s is 100, and its inductance is 100-μH. What is the series resistance at this frequency? (C. & G., 1947.)

Answer: 6·28 ohms.

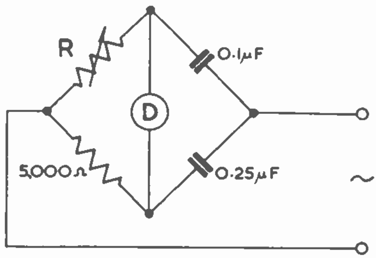
4. Given an audio-frequency oscillator and a cathode-ray oscilloscope, describe how you would determine the presence and source of overloading in a multi-stage audio-frequency amplifier.

(C. & G., 1947.)

5. Explain with the aid of a circuit diagram how a simple a.c. bridge may be used to measure capacitance.

What would be the value of R at balance in the circuit shown, and what would it be if the 0·1-μF capacitor were shunted by one of 0·05 μF?

Answer: 12,500 ohms.



6. How would you measure the self-capacitance and resistance of a coil of about 1,000 μH inductance at a frequency of 500 kc/s?

(C. & G., 1949.)

7. Describe how, with the aid of a signal generator and a cathode-ray oscilloscope, you would locate the source of amplitude distortion in a broadcast receiver. (C. & G., 1949.)

8. Describe how you would measure capacitance (about 300 μμF) at a radio frequency by comparison with a calibrated variable capacitor. (C. & G., 1949.)

9. Describe the tuned-circuit substitution method of measuring inductance or capacitance at radio frequency.

An inductor so measured at a frequency of 1 Mc/s is found to be equivalent to a capacitance of —150 μμF. What is the value of its inductance? (C. & G., 1951.)

Answer: 168 μH.

10. In a Maxwell Bridge the resistance in two opposite arms are 2,000 and 750 ohms respectively. The third arm consists of a capacitance of 0·05 μF in parallel with a resistance of 40,000 ohms. Deduce the conditions for balancing this bridge and calculate the inductance and resistance of the coil under test.

(E. T. A. Rapson, " Problems in Radio Engineering ".)

Answer: 75 mH, 37·5 ohms.

11. What is meant by the "Q-factor" of an inductor? Describe a simple method for measuring the Q-factor of inductors at radio frequency.

An inductor has a Q-factor of 45 at a frequency of 600 kc/s; calculate its Q-factor at 1,000 kc/s, assuming that its resistance is 50 per cent. greater at 1,000 kc/s than at 600 kc/s. (C. & G., 1952.)

Answer: 48·5.

12. Prepare a list of the test equipment you would need for the alignment and maintenance of broadcast receivers, and write brief notes on the uses of each item. (C. & G., 1952.)

13. Given an oscillator calibrated in frequency, a calibrated variable capacitor and a valve-voltmeter, describe how you would determine the inductance L, the Q-factor and the series resistance R of an unknown inductor.

If the inductor resonates at 1·2 Mc/s with 1,010 μμF, and a change of capacitance of 20·2 μμF varies the response from 0·707 volt through a maximum of 1 volt back to 0·707 volt, determine L, R and Q. (The self-capacitance of the coil should be ignored.) (C. & G., 1953.)

Answers: 17·4 μH, 1·31 ohms, 100.

Specimen Answer

Q. Describe a method of measuring the inductance and capacitance of a medium-wave aerial, paying particular attention to the precautions needed to ensure accuracy. (C. & G., 1942.)

A. A suitable method consists in adding a known value of inductance in series with the aerial and measuring the resonant frequency of the aerial plus added inductance. If this is repeated with other values of added inductance, the inductance and capacitance of the aerial may be determined graphically.

Consider the circuit shown in Fig. IX.Q.1. If the coupling between the oscillator and aerial circuits is loose, the current through the coupling coil is independant of the aerial impedance, and may be kept constant for all frequencies of the calibrated variable oscillator, the current being indicated on the thermocouple instrument M_1.

With a known value of added inductance L the frequency of the oscillator is adjusted, keeping M_1 constant, until resonance is indicated by a maximum-current reading on M_2, which is also a thermocouple instrument. The value of L and the resonant frequency f are recorded, and the process repeated with other values of L. From the results obtained a graph is plotted with $\frac{1}{(f_r)^2}$ as ordinates and L as abscissæ, as shown in Fig. IX.Q.2.

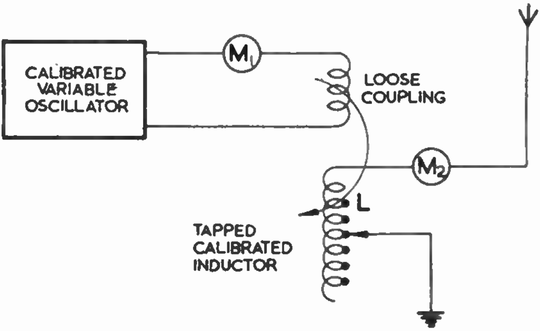


FIG. IX.Q.1

Now if the effective inductance and capacitance of the aerial are L_A and C_A respectively, we have :

$$f_r = \frac{1}{2\pi\sqrt{(L + L_A)C_A}}$$

$$\therefore \quad \frac{1}{(f_r)^2} = 4\pi^2 (L + L_A)C_A$$

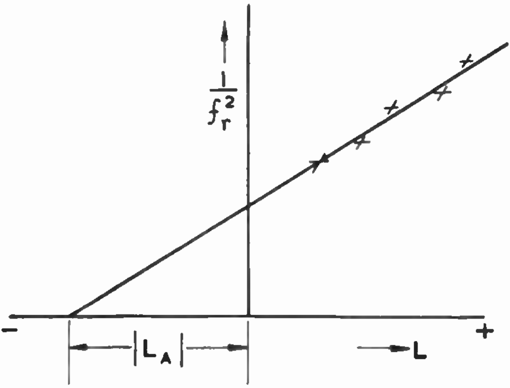


FIG. IX.Q.2

i.e., the graph is a straight line with slope 4^2C_A and by measuring the slope of the plotted graph C_A may be determined. When $\frac{1}{(f_r)^2} = 0$, $(L + L_A) = 0$, thus the value of L_A is given by $-L$, the intercept on the axis of L.

To ensure accurate results the oscillator itself should be screened so that no stray coupling with the aerial circuit can occur. The power output from the oscillator should be large enough for the currents used in the test to swamp any aerial currents induced by atmospherics or distant senders. The resistance introduced into the circuit by the inductor L, the coupling coil, meters M_1 and M_2 and necessary connexions, should be kept as small as possible so that a sharp resonance peak is obtained and f_r can be accurately determined. The series of readings should be taken over a range close to the working frequency of the aerial by choosing a suitable range of variation of L, so that the error due to added self-capacitance is small.

It has been assumed that the aerial impedance is capacitive over the range of test frequencies used. Should it be inductive in this range it will be necessary to connect a capacitor in series with L, the value being chosen to make the aerial plus capacitor a net capacitive reactance in the test range. The value of C_A determined from the slope of the graph will then be the joint capacitance of the aerial and additional capacitor in series, and if the value of the latter is known the capacitance of the aerial may be calculated. If this additional capacitor is required, it must be of low loss, preferably an air-dielectric type.

(P.O. Eng. Dept.)

CHAPTER 10

ACOUSTIC EQUIPMENT

THE apparatus used for converting sound waves into corresponding electrical changes is of great importance in broadcasting systems, where fidelity of transmission and reproduction is essential. The production and reproduction of gramophone records also requires equipment of high quality to obtain realism.

10.1. The Nature of Sound

Sound is caused by the vibration of any object at a frequency within the approximate range 20 c/s–20 kc/s, although the limits of this range vary among different persons. The vibration of an

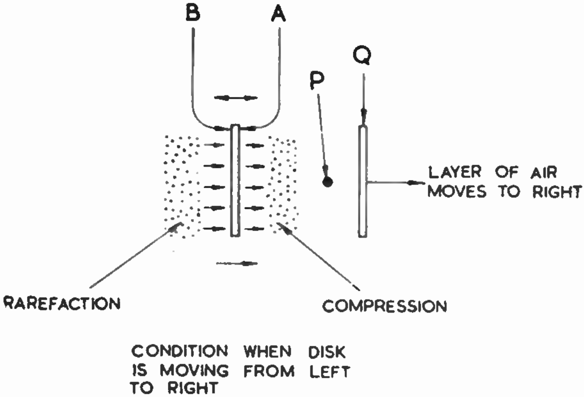


FIG. X.1.—Illustrating Sound Waves

object is transmitted to the surrounding air, and so to the ear of the hearer, who from past experience associates various aural sensations with certain occurrences.

Sound waves travel at a speed of about 1,100 ft. per second in air at normal temperature and pressure, and from the equation $\text{wavelength} = \dfrac{\text{velocity}}{\text{frequency}}$ we can determine the wavelength of a sound of any frequency. Consider a flat disk, moving to and fro in simple harmonic motion at an audible frequency (Fig. X.1).

When the disk is moving from left to right, the air immediately against the face A of the disk is carried with it, thus compressing the air a little farther away. Meanwhile the air against face B is being drawn with the disk and is expanding. This state of affairs is reversed when the disk reverses its direction of motion, with the result that layers of air near to the disk are alternately compressed and rarefied in sympathy with its movement. The compression and rarefaction of layers of air is then transmitted through the air with the velocity of sound, and if a means of measuring the air pressure is placed at a point such as P (Fig. X.1) it will be found that the pressure there varies in the way previously indicated. Consider the movement of a thin layer of air (Q, Fig. X.1) under the influence of a sound wave. It is clear that when pressure to the left of Q is high, the vertical layer passing through Q will move from left to right, whereas when the pressure there is low, the layer will move in the opposite direction. The movement is, of course, small, and is oscillatory except in the case of explosions.

The range of loudness of sounds which are perceptible to the average person is very great; the loudest bearable sound pressure is about 10^{12} or 10^{14} times that of a sound which is just audible. Sound pressure is usually measured in r.m.s. dynes/sq. cm.

The human ear is sensitive to changes in pressure of the order of two to one only, that is, one can tell that the pressure has changed if it is halved or doubled, but one is likely to be uncertain if the change is only to $1\frac{1}{2}$ or $\frac{2}{3}$ of the original pressure.

Telecommunication engineers dealing with the electrical signals in transmission circuits have devised a logarithmic scale to assist them in referring to power levels in different parts of circuits or to ranges of power involved when reproducing different sounds. The logarithmic unit is the " bel " and the difference in bels between two powers W_1 and W_2 is given by the common logarithm of their ratio, that is,

$$\text{bels difference} = \log_{10} \frac{W_1}{W_2}$$

The " decibel " (abbreviated db.), one tenth of the bel, is more commonly used, being a convenient practical unit :

$$\text{decibels (db.) difference} = 10 \log_{10} \frac{W_1}{W_2}$$

Thus powers having a ratio of 10 differ by 10 db. ($\log 10 = 1$), while if the ratio is 100, the difference is 20 db. ($\log 100 = 2$) and so on. Since the sound pressure developed by a reproducer is proportional to the power input, and a change of about $1\frac{1}{2}$–1 is barely noticeable, a change of 1·8 db. in the power (log 1·5 = 0·18 approx.) is also barely noticeable.

If the decibel notation is applied to the e.m.f., p.d. or current in a given circuit, then, because power is proportional to I^2 or E^2, db. $= 20 \log_{10} \frac{E_1}{E_2} \left(\text{or } \frac{I_1}{I_2}\right)$.

The great audible range of intensity (10^{12}–10^{14} at 1,000 c/s or 120–140 db.) cannot be used in practice because of background noise, self-generated noise in equipment, etc., at the low end, and discomfort, disturbance to neighbours and power involved at the high end of the range.

The range of peak volume level transmitted by the B.B.C. is reduced to about 20 db. by manual control of the overall amplification during a programme; this compression enables a satisfactory separation between signals and noise to be maintained while permitting the transmission of reasonable artistic changes of level.

10.2. Principles of Microphones

Microphones are used to convert acoustical energy into electrical energy, and the main requirements may be summarised as follows :

1. Over the audible range the electrical output should be proportional to the acoustical input.
2. The electrical output should not contain harmonics, " overtones ", or sum and difference tones not present in the acoustical input.

Microphones may conveniently be classified thus :

(i) Principle of conversion of acoustical energy into electrical energy.
(ii) The nominal directional characteristics obtained.

Considering the principle of conversion, all microphones involve some moving part called a diaphragm which is set in motion by the sound waves and transmits this motion to the converter. The classification may be carried further by considering the means of producing electrical signals from mechanical motion :

(*a*) Carbon—change of electrical resistance of a conductor by mechanical compression.
(*b*) Condenser—change of capacitance by moving one conductor relative to another.
(*c*) Piezo-electric—generating e.m.f. by stress in special materials (Rochelle salt and other crystalline materials).
(*d*) Electro-magnetic—movement of conductor in magnetic field, including the type in which the conductor itself forms

the diaphragm, known as a ribbon microphone. In another type the conductor is a coil attached to a diaphragm; this is known as a moving-coil microphone.

Many microphones may be required to operate over a frequency range of 50–10,000 c/s, *i.e.*, a range of 200 : 1, and the corresponding wavelengths are from about 20 ft. down to about 1 in. This means that at the low-frequency end of the range the microphone is small compared with the wavelength, and at the high-frequency end the wavelength is small compared with the microphone. Because of this the directional characteristics of microphones at high frequencies are liable to diverge from those obtained at low frequencies. It is therefore convenient to classify microphones by their *nominal directional characteristics*, or the directional characteristic obtained at mid-band frequencies. We have then :

(*a*) Nominally omnidirectional; many moving-coil and condenser and a few ribbon microphones of this type exist.

(*b*) Nominally cosine or " figure-of-eight ". The commonest figure-of-eight microphones are of the ribbon type; condenser microphones are also made with this directional characteristic (Fig.X.2).

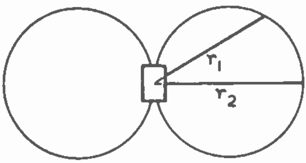


FIG. X.2.—Figure-of-eight Characteristic

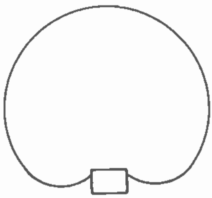


FIG. X.3.—Cardioid Characteristic

(*c*) A combination of (*a*) and (*b*), the commonest example being the cardioid microphone (Fig.X.3). Some commercial cardioid microphones have a ribbon element. Others are a combination of ribbon and moving-coil elements. Condenser-type cardioids and moving-coil microphones having cardioid characteristics also exist.

There are other microphones which depend for their successful operation on their position relative to the source of sound. Two examples in this group are the throat microphone, which is held in contact with the user's larynx, and the lip or close-talking microphone, which is held in close proximity to the user's lips

(Fig. X.4). Some lip microphones are also "noise-cancelling" for use by commentators in noisy surroundings.

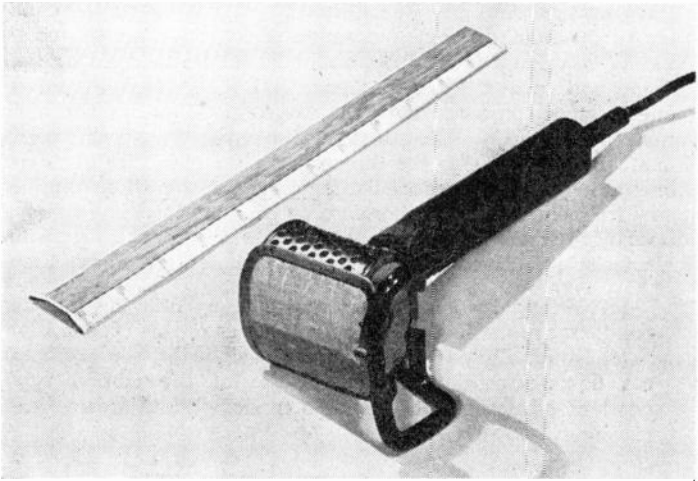


FIG. X.4.—B.B.C. Lip-microphone Type L2

A particular method of converting acoustic to electric power is not always associated with the same directional characteristic.

10.3. Construction of Carbon Microphones

Carbon microphones comprise the first type to be developed to a degree capable of giving reasonably faithful results. Fig. X.5 shows a sketch of the type known as a "Transverse Current" microphone. This consists of a block of insulating material, which may be a plastic moulding, a wood block or a block of marble in one face of which are cut two parallel grooves about ½ in. deep and spaced about 2 in. apart. The grooves are connected by a shallow channel cut in the face of the block, and a carbon rod is fixed into each groove by means of a conducting rod carried through the block and fitted with terminals at the back.

The channel and grooves are sealed by a thin mica sheet (about 0·001 in. thick), and the resulting cavity is filled with dry fine carbon granules through a hole which is then sealed. These granules form an electrical circuit between the two carbon rods, and the resistance of the arrangement is a few hundred ohms. If pressure is applied to the mica sheet or "diaphragm" the resistance will be found to decrease slightly, and if the microphone is

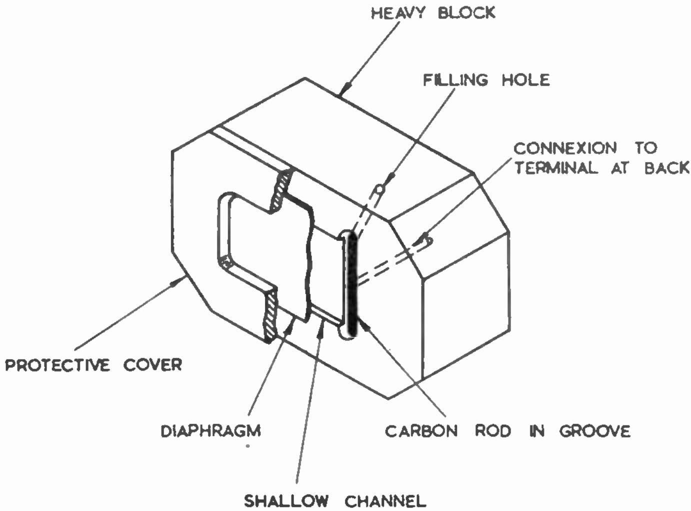


FIG. X.5.—Construction of Carbon Microphone

connected in series with a battery and one winding of a transformer, a current will flow (Fig. X.6). The current in this circuit will depend on the battery p.d. and the resistance, which is mainly that of the microphone.

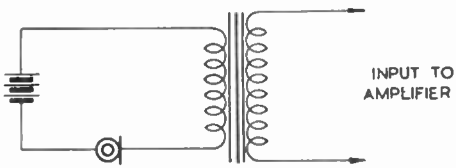


FIG. X.6.—Circuit for Carbon Microphone

When sound waves strike the diaphragm the pressure on the granules is alternately increased and decreased, causing the resistance of the instrument to decrease and increase correspondingly. The change of resistance depends on the amplitude of the movement of the diaphragm, and causes the current in the primary of the transformer to rise and fall. An e.m.f. is thus induced in the secondary, and a p.d. appears across the secondary load resistor which is proportional to the change of current in the primary circuit.

Carbon microphones suffer from the disadvantage of requiring a direct supply of 10–15 volts and producing a " noise " e.m.f. because of the spasmodic flow of current through the carbon granules. This noise is reproduced as a hissing sound. The granules also tend to pack together, causing the instrument to become insensitive; this can be remedied by inverting and shaking it gently. Another disadvantage of carbon microphones is that they give a distorted output when the sound input is high, and they are also very sensitive to mechanical shocks.

In spite of all these drawbacks, the marble-block version due to Reisz was used for many years by the B.B.C.

10.4. Construction of Moving-coil Microphones

The construction is similar to that of the moving-coil loudspeaker on a small scale as shown in Fig. X.7. When sound waves

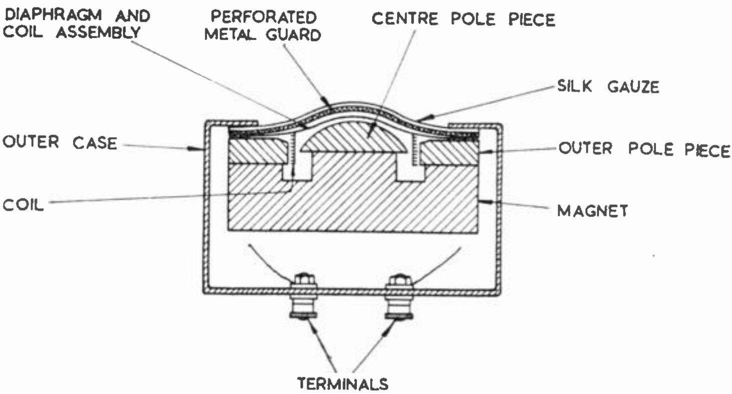


FIG. X.7.—Diagram of Moving-coil Microphone

strike the cone or diaphragm of the instrument, the coil attached to it moves in the gap of the magnet, and an e.m.f. is generated which is proportional to the product of the number of turns on the coil, the magnetic flux density and the velocity of the coil.

The velocity depends on the pressure of the sound wave, the restraint of the suspension of the diaphragm and its mass. The back of the instrument is enclosed so that sound waves cannot reach the back of the diaphragm, where they might cancel those arriving at the front. It is also heavily damped to give a flat frequency response.

Modern materials (especially permanent magnets) and methods of manufacture have enabled moving-coil microphones to be capable of high fidelity, and they are widely used.

No power supply is required, and the instrument is robust and can handle large inputs without distortion.

10.5. Construction of Ribbon Microphones

This type of microphone was invented about 1913, and was developed by Olsen and others to a high degree in the 1930s.

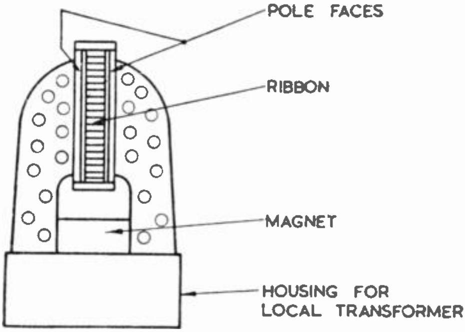


FIG. X.8.—Diagram of Ribbon Microphone

The moving element consists of a thin strip of aluminium foil suspended in the gap of a powerful magnet (Fig. X.8). The foil is anchored at each end by insulated supports, and is corrugated across its width so that it is more flexible. The foil is not otherwise restrained, and being extremely light it vibrates in sympathy with any sound waves in which it is " immersed ". The poles of the magnet are pierced to reduce obstruction to the sound waves. Sounds from the sides produce equal and opposite pressures on both sides of the ribbon, and thus the response is zero at the sides.

The output of this ribbon microphone is fed via a step-up transformer (housed in its base) to the input of the microphone amplifier. The nominal directional characteristic of this type of ribbon microphone is a figure-of-eight. Other types of ribbon microphone may be made to have any intermediate directional characteristic between a circle and a cardioid $\left(i.e., \text{of the form } \frac{1 + A\cos\theta}{1 + A}\right)$, and examples of ribbon microphones giving cardioid, figure-of-eight or characteristics between these two are common.

10.6. Construction of Piezo-electrical Microphones

Crystals of Rochelle salt and ceramics are examples of materials so far discovered which, when cut or polarised in certain ways,

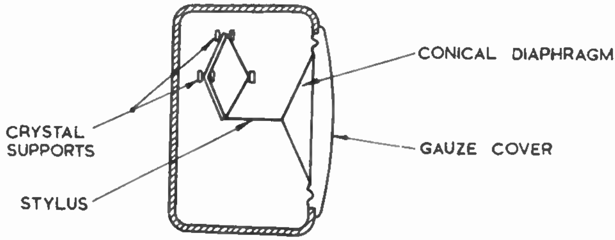


FIG. X.9.—Diaphragm Type of Piezo-electric Microphone

generate electrical charges on opposite faces if they are deformed, and deform themselves if potentials are applied to the same faces. Rochelle salt and ceramics demonstrate this property to a greater extent than quartz, and are therefore preferred. Piezo-electric microphones operate by causing sound waves to deform a suitable crystal, from whose faces potentials are picked up by attaching foils to them. This deformation can be caused in two ways. Fig. X.9 shows a crystal attached by a short stylus to a diaphragm, which is moved by sound waves and twists the crystal as a result.

Another form comprises a pair of crystals forming a cell, which expands or contracts when sound waves strike it. The latter construction is known as the " cell " type. It is capable of high fidelity. For normal purposes a number of cells are used in series to increase the output.

The e.m.f. produced by a crystal is proportional to the displacement, and the impedance of the microphone is high, being effectively a small capacitance in series with the resistance. The high impedance requires the microphone to be connected directly into the grid circuit of a valve (Fig. X.10), and it also means that the capacitance of the cable connecting the microphone to the amplifier will reduce the p.d. available (Fig. X.11). The output of a cell-type microphone depends on the number of cells in series.

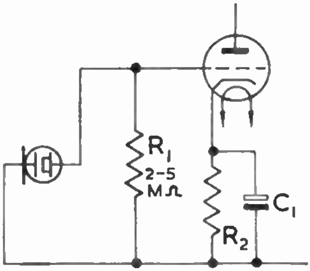


FIG. X.10.—Connexion of Piezo-electric Microphone to Amplifier

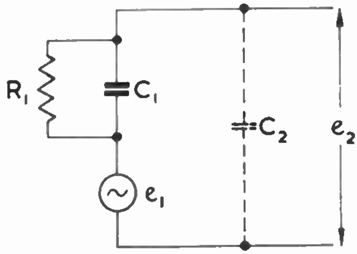


FIG. X.11.—Showing how Capacitance of Microphone C_1 and Capacitance of Cable C_2 give

$$\frac{e_2}{e_1} = \frac{X_{c_2}}{X_{c_1} + X_{c_2}} = \frac{1}{1 + C_2/C_1}$$

10.7. Construction of Condenser Microphones

Consider a variable capacitor connected in series with a p.d. and a resistor as shown in Fig. X.12. The capacitor charges up to the applied p.d. through the resistor, and if its capacitance is C farads and the p.d. is V volts, the charge $Q = CV$ coulombs. If the capacitance is now decreased, say to $C - \delta C$, the p.d. across it

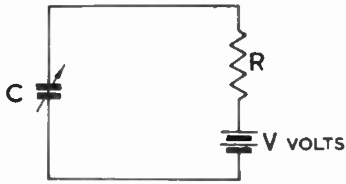


FIG. X.12.—Principle of Condenser Microphone

becomes $V + \delta V = Q/(C - \delta C)$ and the capacitor commences to discharge into the source p.d. If the capacitance is made to alternate above and below a given value, the p.d. across it similarly falls below and rises above the mean value, and if the resistor is high in value, current will not flow from the source sufficiently to alter the charge appreciably.

One of the simplest types of condenser microphone consists of a solid metal backplate and a very thin diaphragm of duralumin (about 0·001 in. thick) or of gold-sputtered plastic spaced apart (also about 0·001 in.) to form a small air-dielectric capacitor.

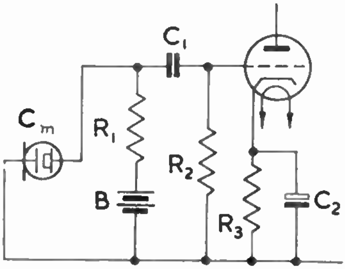


FIG. X.13.—Circuit for Condenser Microphone

This is connected in a circuit as shown in Fig. X.13. Here C_m is the microphone self-capacitance and C_1 is a blocking capacitor to isolate the direct p.d. from the grid of the amplifier. Resistors R_1 and R_2 have values as high as permissible from considerations of leakage, valve-grid current, etc. The microphone amplifier is usually housed in the same casing as the microphone.

10.8. Summary of Principal Features of Microphones

1. *Carbon*

Requires polarising supply, is sensitive to mechanical shocks, cannot handle large inputs without distortion. Commonly used where fidelity is not of first importance, though better types are sometimes used for broadcasting. Has a high output capable of operating headphones without the use of an amplifier.

2. *Moving Coil*

Robust, less sensitive than carbon. No additional supply necessary. Capable of sufficient fidelity for most purposes. Mostly omni-directional at low frequency, becoming increasingly directional with increasing frequency.

3. *Ribbon*

Suitable for the highest fidelity. Insensitive, usually figure-of-eight, sometimes omni-directional or cardioid, or intermediate (Fig. X.14).

4. *Piezo-Electric*

Sensitive, cell-type capable of high fidelity. Less robust than the previous types. High impedance and requires no transformer.

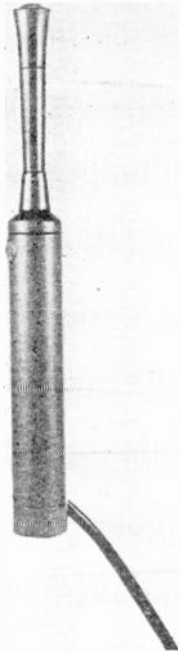


FIG. X.14.—RCA "Starmaker" Ribbon Microphone

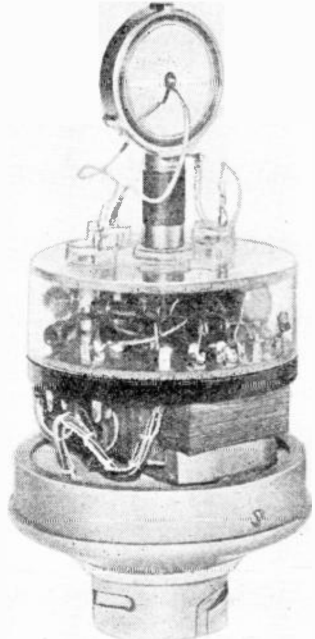


FIG. X.15.—Neumann Condenser Microphone with Adjustable Characteristic. Note local amplifier

Usually omni-directional. Diaphragm type has similar directional characteristics to moving-coil. cell type similar to condenser.

5. *Condenser*

Capable of high fidelity, have a high impedance. Require stable, smooth polarising supply and local amplifier because of

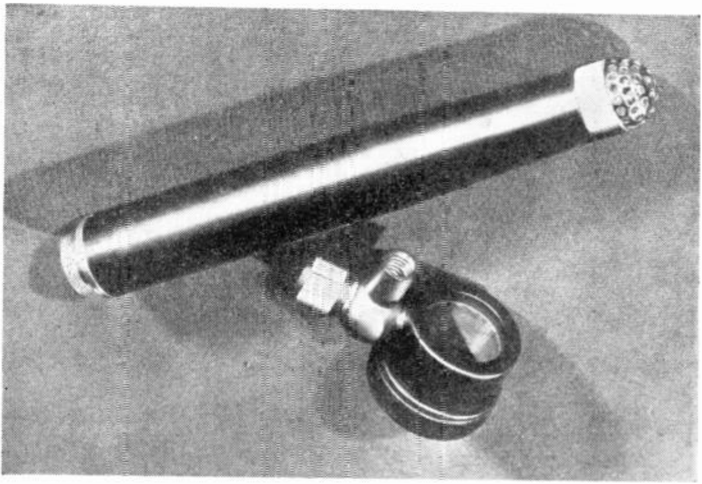


FIG. X.16.—Philips Condenser Microphone

Local amplifier is contained in the tube; the universal clamp below enables the microphone to be directed as required.

high impedance. Characteristics may be made either omni-directional only, adjustable omni-directional/cardioid or intermediate between circle and cardioid. Among the virtues of ribbon and condenser microphones is that by careful design a smooth and flat frequency characteristic is more easily obtained than with moving-coil microphones (Figs. X.15 and X.16).

10.9. Gramophones

The gramophone record is a convenient and very popular method of recording sound. It has been highly developed commercially, in addition to other methods such as sound-on-film and magnetic-tape recording, both of which are also highly developed and much used.

In disk recording the variations of air pressure corresponding to the sound waves are caused to modulate a spiral groove on the surface of the disk, and the deviations from the exact spiral are a plot of the pressure variations. Modern gramophone recordings are made with transverse modulation, but the early "phonograph" of Edison and the modern office dictating machine employ "hill and dale" modulation, in which the depth of the groove is varied. Reproduction of the record is carried out by following the groove with a stylus whose movement actuates a generator of either electro-magnetic or piezo-electric type. This is known as a "reproducing head" or sometimes as a "pick-up".

A reproducing head must convert a mechanical movement into an alternating e.m.f., and the electrical output must be a faithful reproduction of the recorded modulation. Mechanically, it must cause little wear on the record track.

There are three main types of reproducing head:

1. Moving-iron or moving-armature.
2. Moving-coil.
3. Crystal or piezo-electric.

10.10 Moving-iron Type

This consists of a small fixed coil with a soft-iron armature passing through its centre. The armature carries a stylus which runs

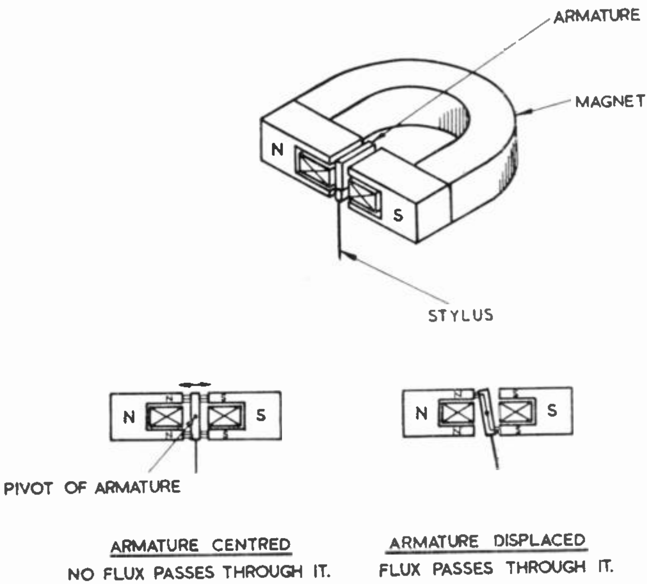


FIG. X.17.—Diagram of Moving-iron Reproducing Head in the groove of the disk and is mounted on rubber. The coil and armature are fitted into a magnet system, as shown in principle in Fig. X.17. When the armature is tilted sideways by the modulation on the groove the reluctance of the magnetic circuit is decreased because the air gap is reduced so that the flux in the armature is increased. The changing flux generates an e.m.f. in the coil.

This type has been highly developed and is capable of good fidelity, by making the size and mass of the armature small and by improvements in the rubber suspension. The exact arrangement of the magnetic circuit is also important, while modern magnets have enabled the weight of the head to be reduced.

10.11. Moving-coil Type

The moving-coil head consists likewise of a small permanent magnet with an air gap in which is mounted a coil attached to the stylus (Fig. X.18). Movement of the coil in the gap generates an

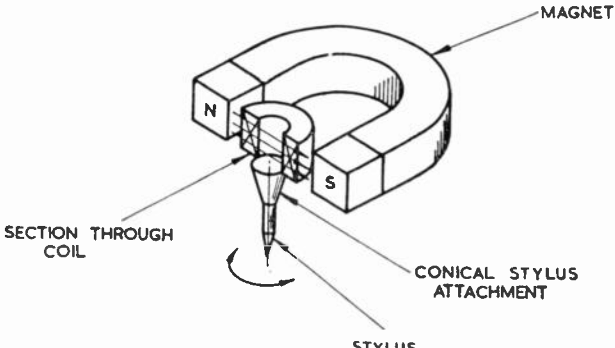


FIG. X.18.—Diagram of one Type of Moving-coil Reproducing Head

e.m.f. in the coil proportional to the flux and the velocity of the coil. In order to follow rapid movements of the needle on high notes, the whole moving structure is made as light as possible and the number of turns on the coil is small. A transformer is used to step up the e.m.f. before feeding it to the amplifier.

This type has also been highly developed, and some models comprise a single conductor only, after the manner of the ribbon microphone.

10.12. Piezo-electric Type

Like some piezo-electric microphones, these heads make use of a Rochelle salt crystal which is twisted by the movement of the stylus (Fig. X.19). The movement causes charges to be built up on the faces of the crystal, and a p.d. is obtained from them by

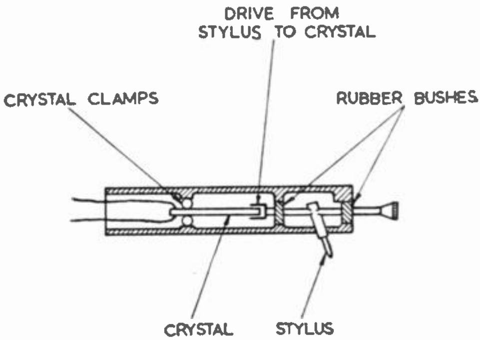


FIG. X.19.—Diagram of Piezo-electric Reproducing Head

means of foils. Like the piezo-electric microphone, the output is proportional to the amplitude of movement, whereas the electro-magnetic heads give an output proportional to the velocity of the armature or coil.

10.13. General

The reproducing head is carried on a light but rigid arm, which must be freely pivoted to enable the reproducing stylus to follow the spiral groove without excessive side force. Vertical movement of the head must also be permitted, so that the head may be placed on the disk and the vertical load on the stylus is controlled by spring or counter-weight. The arm is shaped so that the armature pivot is closely tangential to the groove over the useful part of the disk.

A reproducing head comprises a heavy mass (the head and arm) free to rotate about the pivot and held in place by the stylus resting in the groove of the record. The stylus is resiliently mounted, and the system therefore has resonant frequencies. There are two principal resonances; a low-frequency resonance determined by considering the head pulled to one side against the restoring force on the stylus, and a high-frequency resonance which can be found by lifting the arm from the record and plucking the stylus. The same restoring force is effective in each case, but the mass involved is much less in the second instance. Another possible high-frequency resonance is a torsional (twisting) one in the arm itself, determined by the stiffness of the arm and the mass of the head.

If the record excites any of these resonances, an excessive output is generated, and wear on the record is increased.

The effect of resonances is to make the output a function of frequency which is not desirable, although use was at one time made of the low-frequency resonance to give some sort of inherent correction for the reduction of amplitude at low frequencies when recording.

The modern tendency is to reduce the mass of the moving parts and the restoring force on them, so that the low-frequency resonance is very low. The low mass of moving parts raises the high-frequency resonance of the armature in spite of the reduced restoring force. The torsional high-frequency resonance is made high (above the recorded range) by making the arm torsionally stiff. The result is that the characteristic over the recorded range is a fairly smooth line, and correction may be made in the associated amplifiers when it is required.

Questions

1. Describe with a sketch the construction and principles of operation of a moving-coil microphone. Indicate the approximate form of the frequency response. (C. & G., 1949.)

2. Sketch and describe the construction of a simple type of ribbon or condenser microphone and indicate its directional properties by means of a polar diagram. (C. & G., 1951.)

3. Sketch the essential parts of either (*a*) a permanent-magnet moving-coil loudspeaker, or (*b*) a permanent-magnet gramophone pick-up. Describe the operation of the example chosen. (C. & G., 1938.)

Specimen Answer

Q. Sketch and describe a crystal transmitter. What are the principal advantages and disadvantages of this type compared with other transmitters?

A. A typical crystal transmitter is shown in Fig. X.Q.1(*a*), the crystal assembly being shown in section in Fig. X.Q.1(*b*).

The transmitter depends on the piezo-electric effect for its operation, *i.e.*, when certain crystals are mechanically strained a difference of potential is developed between certain faces of the crystal, the value of the p.d. being proportional to the mechanical strain. Slices of Rochelle salt crystals are used for transmitters because these crystals are more sensitive than piezo-electric quartz, although less stable. Two slices are cut differentially, so that when cemented together to form a bimorph, twisting or bending of the bimorph produces a p.d. between the two adjacent faces, which are covered with metal foil to form one electrode, and the two outer surfaces which are also covered with foil and connected together. Thus the e.m.f.s generated by mechanical strain in the two component slices are in parallel; the very high internal impedance of the generator is thereby reduced. The bimorph shown in the figure is of the bender type, being clamped along two opposite edges with the diaphragm attached to the centre, so that

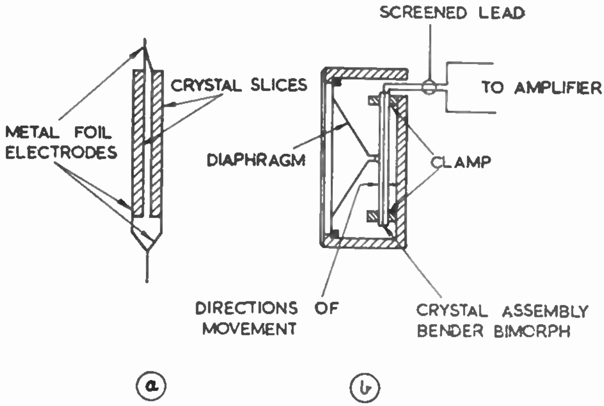


FIG. X.Q.1

vibrations of the diaphragm due to sound waves impinging on it cause the crystal to vibrate in a bending mode and produce e.m.f.s corresponding in waveform to the incident sound waves. The principal advantage of the crystal transmitter is its good frequency response, which is almost flat from 300 to 2,000 c/s, and then rises to a peak at about 6,000 c/s. It is free from the background noise, which is characteristic of carbon-granule transmitters. It has the disadvantages of low sensitivity and a high internal impedance; an associated valve amplifier is therefore required. A further disadvantage is reduction of sensitivity with increase of temperature, the transmitter being practically inoperative at 100° F. (P.O. Eng. Dept.)

CHAPTER 11

DIRECTIONAL AERIALS

In this Chapter we shall examine the properties and construction of the loop aerial and its use in simple direction-finding apparatus. We shall also discuss some of the errors in direction finding.

11.1. Polar Diagrams

We shall first consider a method of indicating the reception or " pick-up " obtained at various positions round any aerial. This type of indication, known as a polar diagram, has axes consisting of concentric circles and radii. In order to produce a polar diagram

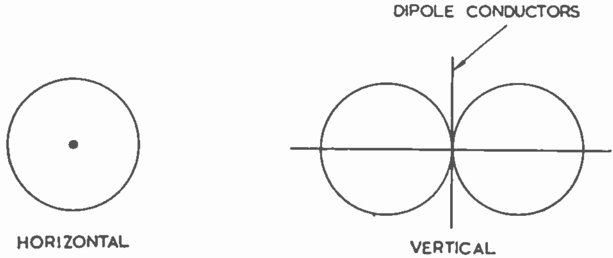


Fig. XI.1.—Polar Diagrams of Dipole Aerial

of field strength, measurements are taken at a fixed distance, and in every direction, from the transmitter, and the results plotted in a particular manner. These results will show a value of field strength for every bearing. When plotting, the signal strength is represented by the length of the radii and the bearing by the angle of the radii from a datum line. For example, if the field strength is of equal magnitude all the way round a transmitter, the resultant polar diagram would be a circle having the location of the transmitter as its centre. This is the diagram obtained from a vertical dipole, the measurements being recorded in a plane normal to the dipole. The important point to remember about polar diagrams is that they are plotted using Polar as distinct from Cartesian co-ordinates. Similar results would be obtained if the measuring set were fixed and the transmitter moved around it, the strength being recorded and plotted as before.

Clearly these measurements may be carried out in either the vertical, horizontal or any inclined plane, giving different results for each plane, Fig. XI.1. If the "complete picture" of the radiation into space from a transmitting aerial is required, a combination of polar diagrams is needed giving a three-dimensional result. It is necessary that the results be recorded near to the transmitter so that the effect of the intermediate ground is negligible.

Another type of graph, not to be confused with a polar diagram, is one on which contours connecting points of equal signal strength are plotted on a map. With this type of graph, the effect of the ground over which the waves pass is clearly indicated. Finally, it must be noted that although we shall be dealing with receiving aerials, the arguments and results obtained would apply equally well if the aerials were used for transmission.

11.2. The Loop Aerial

The loop aerial consists of a single loop of wire usually in the form of a rectangle. Fig. XI.2 shows such a loop having corners A, B, C, D and capable of rotation in a vertical plane about the axis XY.

In Volume I it was shown that for a vertical conductor of effective height h_e or length l_e in a field of strength e volts/metre the induced e.m.f. is simply $E = eh_e$ or el_e. Now sides AC and BD are vertical conductors, and if placed in an electro-magnetic field an e.m.f. will be induced in them. Assume now that the plane of the loop is placed so that it is normal to the direction of propagation of the wave and that the wave is vertically polarised (*i.e.*, its electric field is vertical, and at right angles to this and to the direction of propagation is the magnetic field). In this position the sides of the loop AC and BD are equidistant from the transmitter, and therefore the e.m.f.s induced in both sides will be equal in magnitude and phase. Therefore the resultant e.m.f. around the loop will be zero (Fig. XI.2(*a*)). The horizontal sides AB and CD will have no e.m.f. induced in them because the radiated wave is vertically polarised. Now consider the effect of rotating the loop about its axis XY by 90° so that one of the vertical sides is nearer to the transmitter than the other by the width of the loop (Fig. XI.2(*b*)).

In this condition the magnitudes of the e.m.f.s induced in both vertical sides will again be equal, but the time taken for the radiated wave to reach side AC will be longer than that taken to reach BD, so that there will be a phase difference between the two induced e.m.f.s, and the resultant will be the vector difference between the two e.m.f.s. With the loop in any other position than the first we considered (*i.e.*, with the plane of the loop at right

H

angles to the radiated wave), there will be a resultant e.m.f. acting around the loop.

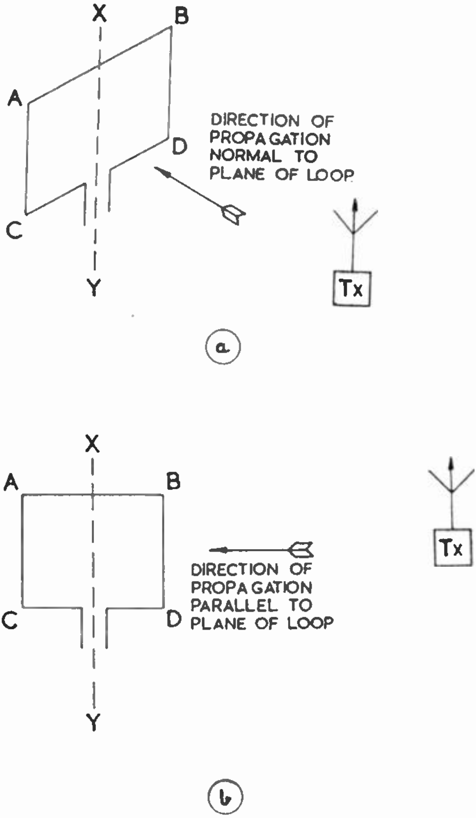


FIG. XI.2.—Reception by Loop Aerial

A more comprehensive result can be obtained by the mathematical analysis of the loop aerial as follows.

Let the width of the loop (AB or CD) be W.

Let the angle made between the plane of the loop and a line joining the centre of the horizontal limb with the transmitter be θ.

When $\theta = 90°$, then the distances travelled by the wave to each vertical limb will be equal; when θ has any other value than 90° there will be a difference in the distance travelled.

Let the distance travelled from front to back of loop be d, then from Fig. XI.3, $d = W \cos \theta$.

The distance "d" may also be represented as a function of the wavelength λ, *i.e.*, $\frac{d}{\lambda}$ or $\frac{W \cos \theta}{\lambda}$.

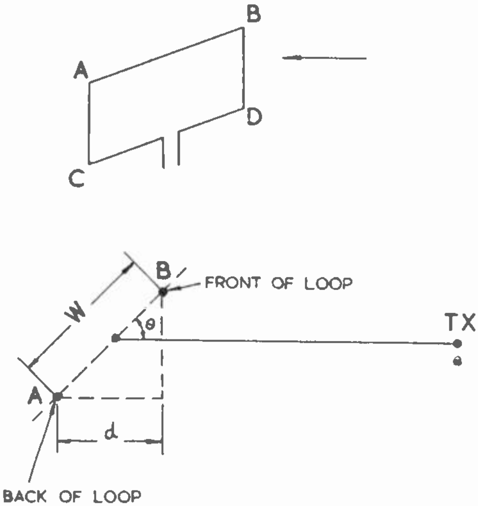


FIG. XI.3.—Loop Aerial

This is shown in Fig. XI.4, where it may also be seen that the phase angle in radians is $2\pi \frac{W \cos \theta}{\lambda}$.

We may now consider the two e.m.f.s equal in magnitude, but differing in phase by an angle $\frac{2\pi W \cos \theta}{\lambda}$ radians. These two e.m.f.s are represented vectorially in Fig. XI.5. The resultant e.m.f. acting in the loop in the direction AB is the vectorial *difference*, and would be given by the e.m.f. in AC minus the e.m.f. in BD. The vector diagram showing this is given in Fig. XI.5(b). The resultant (Fig. XI.5(b)) $OP = 2eh_e \sin \left[\frac{\pi W \cos \theta}{\lambda}\right]$ = the resultant e.m.f. in loop in volts (E_l).

Suppose now that the physical dimensions of the loop are small compared with a wavelength (*i.e.*, W is small relative to λ), then with no appreciable error it is permissible to write $\sin\left(\frac{\pi W \cos\theta}{\lambda}\right) =$

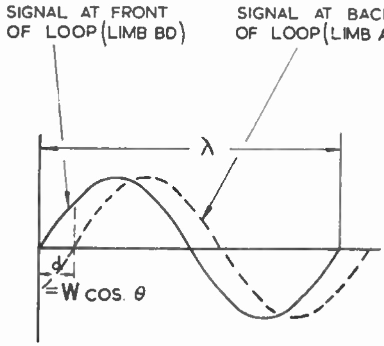


FIG. XI.4.—E.m.f.'s in Loop Aerial

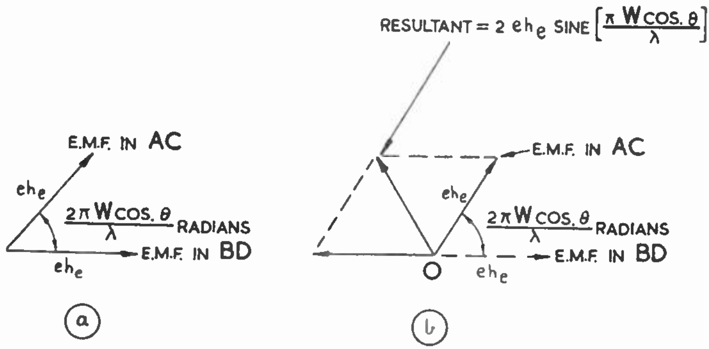


FIG. XI.5.—Resultant e.m.f. in Loop Aerial

$\frac{\pi W \cos\theta}{\lambda}$, since for small angles the sine of the angle is equal to the angle in radians.

$\therefore$ the e.m.f. induced into the loop becomes

$$E_L = \frac{2\, eh_e \pi W \cos\theta}{\lambda}$$

or, since $W \cdot h_e =$ loop area A, then

$$E_L = \frac{2\pi e A \cos\theta}{\lambda} \text{ volts}$$

Note that in the above case, when the dimensions of the loop are small compared with the wavelength, the e.m.f.s induced in the vertical limbs are much greater than the resultant e.m.f. round the loop.

11.3. The Frame Aerial and its Polar Diagram

In order to obtain a larger resultant e.m.f., a loop aerial may consist of many turns. If the number of turns is N, then the e.m.f. induced in one vertical limb will be Neh_e volts and the e.m.f.

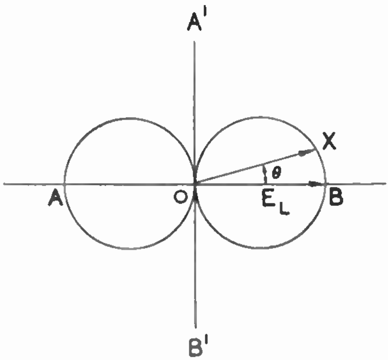


FIG. XI.6.—Figure-of-eight Diagram

(E_L) induced into a frame aerial of N turns will be given by $E_L = \dfrac{2\pi N \cdot e \cdot \times \text{ loop area} \times \cos\theta}{\lambda}$ volts, provided that the dimensions of the loop aerial are small compared with the wavelength concerned.

If E_L is calculated for all values of θ from 0° to 360° a curve may be plotted showing the variation of magnitude and phase of E_L; such a curve is called a Polar Diagram (described above), and for a loop or frame aerial the curve is of the form of two contiguous circles (Fig. XI.6), commonly known as a "figure-of-eight" diagram. When $\theta = 0°$ or 180°, then $\cos\theta$ will be unity and E_L will have its maximum value. Notice that a change of 180° in θ does not alter the numerical value of E_L but only the phase. When $\theta = 90°$ or 270°, then $\cos\theta = 0$ and $E_L = 0$ and at any intermediate angle the value of E_L will depend on the value of $\cos\theta$ (Fig. XI.6). Therefore if AO (or OB) is the maximum value E_L may have, then clearly the value at any other point, say OX in Fig. XI.6, will be given by $E_{L max} \cos\theta$ where θ is angle XOB. When the loop is in the position $A'OB'$, *i.e.*, when $\theta = 90°$ or $270°$ E_L is zero.

11.4. Direction Finding (D.F.)

As has been shown above a loop, coil or frame aerial capable of rotation about its vertical axis will give maximum signal strength when the loop is pointing in the direction of the transmitter and

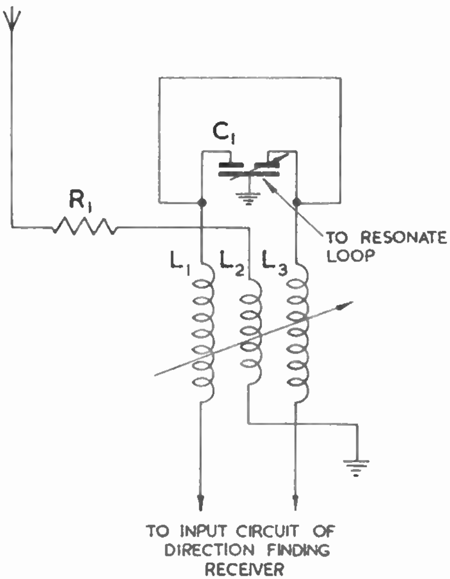


FIG. XI.7.—Addition of Outputs from Vertical and Loop Aerials

minimum signal strength when the plane of the loop is at right angles to the direction of transmission. From this it can be seen that it is possible for the frame aerial to give the direction of the transmission, but not the bearing, *i.e.*, it can tell that the transmitter lies, say, East or West, but it cannot indicate whether to the East or the West. In order that the sense of the bearing may be determined, a vertical aerial (called the sense aerial) is used in conjunction with the frame aerial as described below. It is assumed that all measurements are made several wavelengths from the transmitter, and that the dimensions of the loop are small compared with a wavelength.

The resultant e.m.f. in the loop is approximately 90° out of phase with that induced in the vertical limbs (Fig. XI.5(b)), therefore the e.m.f. from the sense aerial must be changed by 90° so that it is in phase with the resultant e.m.f. in the loop. This may be accomplished by introducing a high resistance in series with the vertical sense aerial, which has the effect of bringing the current into phase with the induced e.m.f. and substantially independent of frequency (Fig. XI.7). The inductive coupling between the sense and loop aerial circuits provides a phase shift of 90° bringing the sense aerial e.m.f. in phase with the resultant e.m.f. in the loop. Fig. XI.7 shows the loop aerial having two equal coils (L_1 and L_3) connected in each leg of its output whilst capacitor C_1 is used to resonate the loop to the desired frequency. Clearly the magnitude

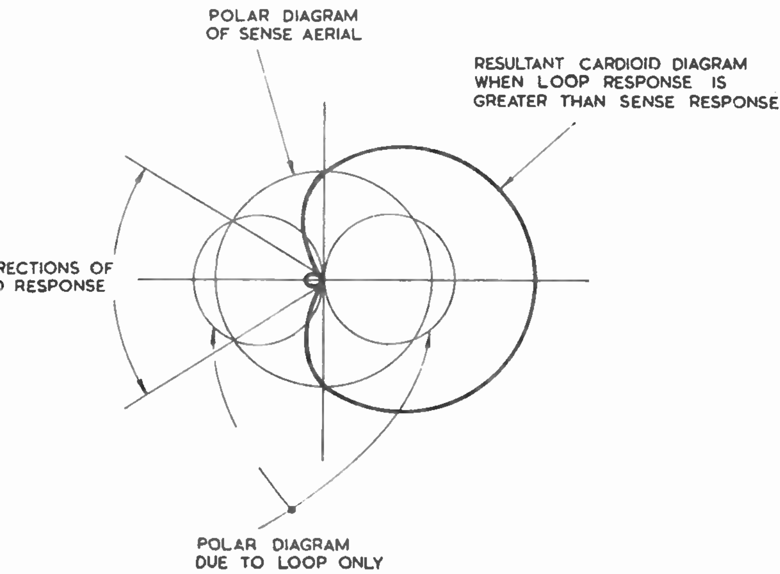


FIG. XI.8(a)

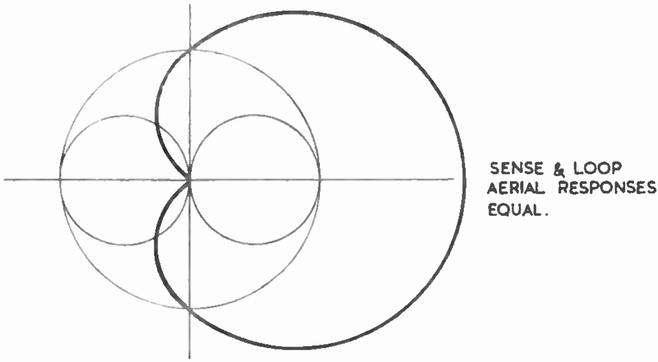


FIG. XI.8(*b*)

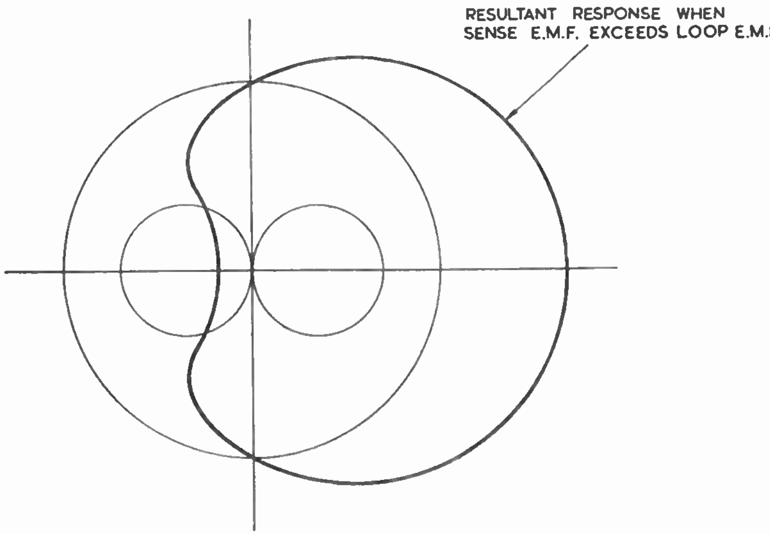


FIG. XI.8(*c*)

FIG. XI.8.—Showing Effect of Various Ratios of Vertical and Loop on Characteristic E.m.f.'s of the sense aerial e.m.f. must not be greater than that of the loop, whilst it must be large enough to produce a noticeable change of pattern. This is achieved by varying the mutual coupling between the coils in Fig. XI.7.

Fig. XI.6 shows that for a given change of angle or rotation the corresponding change of signal strength is greatest around the zero signal position. For this reason it is better to adjust for a minimum than for a maximum, and this is always done in practice.

The effect of adding various amounts of sense aerial e.m.f. to the loop response is shown in Figs. XI.8(*a*), (*b*) and (*c*). It will be noticed that when the two e.m.f.s are equal, the polar diagram is a heart shape, known as a cardioid. After minimum signal strength is determined by the position of the loop, the loop is then rotated 90° to the position of maximum signal and the sense aerial is switched in; by observing whether an increase or decrease of signal occurs, the final bearing may be determined.

11.5. Vertical or Antenna Effect

The accurate operation of a loop for direction finding depends largely upon the ease with which a null reading may be detected. Anything which prevents the accurate detection of this null reading

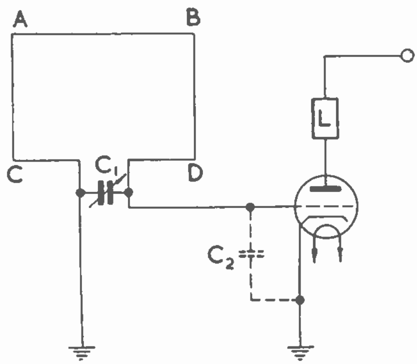


FIG. XI.9.—Simple Input Circuit

tends to reduce the accuracy of the D.F. system. Unwanted signals caused by unequal impedance paths through the two vertical limbs of the loop produce an effect known as " vertical " or " antenna " effect.

Fig. XI.9 shows a simple D.F. receiver; the loop is tuned by a

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capacitor C_1 to the required frequency (this added capacitance is necessary depending on the frequency, Chapter 3, Volume I), and the p.d. developed across C_1 is applied to the grid of the valve. There exists between grid and cathode a capacitance to earth C_2 (because the cathode is earthed), and the point B is not at earth potential. The comparatively large e.m.f.s induced in each vertical limb can cause currents to flow through two unequal paths. (It will be remembered that the resultant e.m.f. in the loop should be zero when the loop is in this position, thereby producing no output from the receiver.)

Elimination of this vertical effect may be achieved in one of two ways :

(1) by balancing the capacitance of each limb to earth ; or

(2) by providing alternative symmetrical paths of lower impedance for the vertical components.

The simplest way to provide symmetrical paths of low impedance to earth is to earth the centre point of the top horizontal limb (Fig.

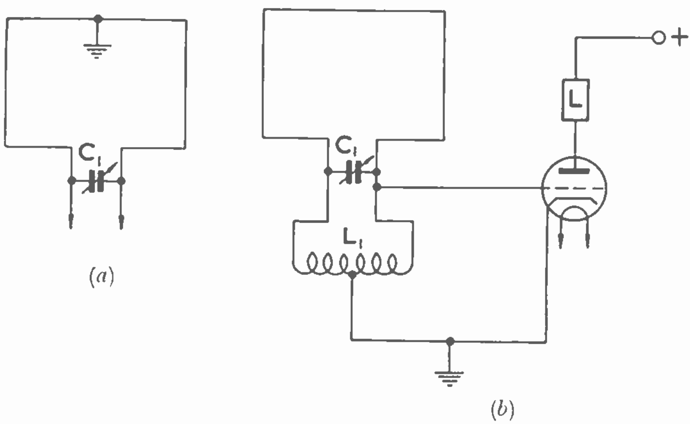

FIG. XI.10.—Balanced Loop Connexions

XI.10(*a*)). This method is satisfactory for frequencies up to approximately 500 kc/s, and the loop is tuned by means of a capacitor connected in parallel with it. At higher frequencies (up to approximately 700–800 kc/s) it is necessary to shunt the loop with an inductor in order to reduce the inductance of the loop, tuning being effected by a parallel capacitor as before. The centre point of the inductor is taken to earth, and so vertical effect is eliminated (Fig. XI.10(*b*)).

At higher frequencies and for more accurate results, care is taken to balance the loop by one of the methods shown in Fig. XI.11.

The use of a transformer as a means of balancing the vertical paths in loop aerials makes it necessary to include an electrostatic screen, otherwise the capacitances existing between the transformer windings may produce unequal paths. Another precaution

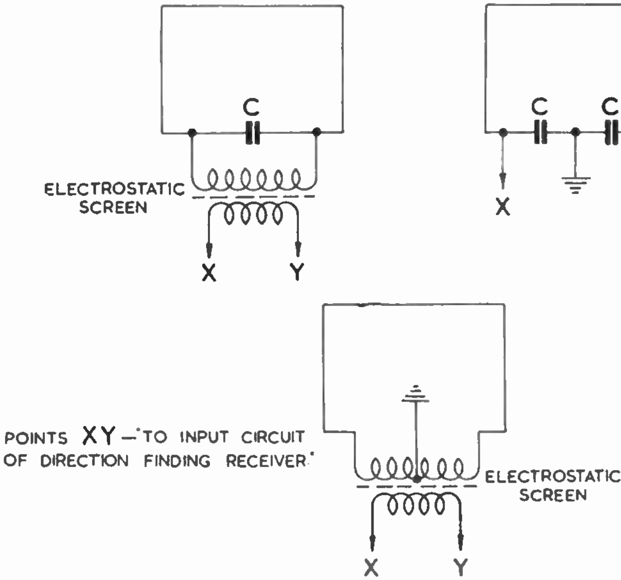

FIG. XI.11.—Balanced Input Circuits

usually taken for the successful operation of the loop aerial when used for low and medium frequencies is the placing of an earthed electrostatic screen around the loop with an insulated joint in it to avoid a short-circuited turn. This ensures constant capacitance to earth from all parts of the loop whatever its position (Fig. XI.12).

11.6. Night or Sky Effect

Since, in general, frame aerials are used at broadcast and lower frequencies (below 600 kc/s), the effect of downward travelling waves is most serious at night, and is often called " night " or " sky " effect.

When a wave, originally vertically polarised, is reflected from the ionosphere, it is no longer polarised in the vertical direction only, but contains horizontal polarisation as well; such a wave when received by a frame aerial will induce e.m.f.s into the horizontal arms, and a zero resultant e.m.f. will not be obtained even when the plane of the loop is perpendicular to the received signal. This night effect sets a limit to the effectiveness of any D.F. system, and when this effect is present, false readings are

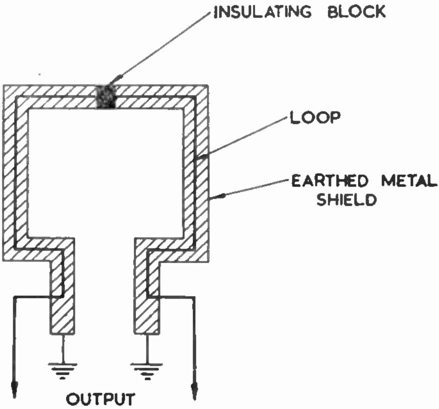

FIG. XI.12.—Screened Loop Aerial

obtained, while it is often not possible to obtain a null reading. It will be appreciated that near a transmitter where the ground wave is stronger than any sky wave that may be received, this night effect is not important. Factors which reduce the ratio of ground wave to sky wave, and hence cause inaccuracies in D.F. measurements, are :

1. As the frequency is increased, the effect produced by the ionosphere is also increased, tending to increase the effect of the sky wave.
2. Measurement at night accentuates the effect of the sky wave compared with day-time measurements.
3. Since the attenuation of electro-magnetic waves is greater over land than water, measurements of ground waves over composite paths cause bearing errors.
4. As the distance from the transmitter increases, so the effect of the sky wave becomes greater.

These effects can be eliminated by replacing the loop by some other aerial arrangement such as an Adcock aerial.

11.7. Adcock Aerial

In its simplest form the Adcock Aerial consists of two crossed aerials as shown in Fig. XI.13. The action is such that for horizontally polarised down-coming waves there is no resultant e.m.f. at the output terminals. Suppose due to a downward-travelling horizontally polarised wave an e.m.f. is induced into the horizontal limb AD so that a current tends to flow through the output

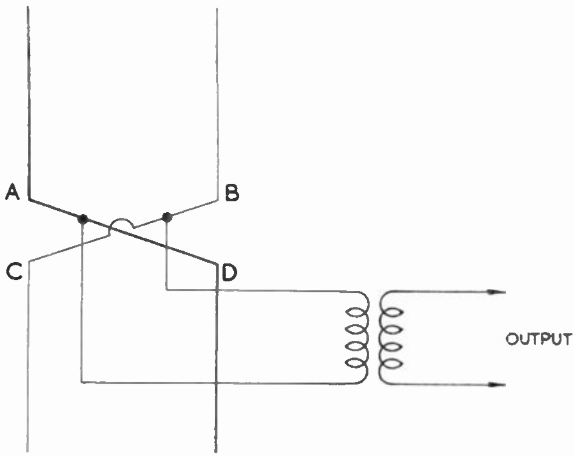

FIG. XI.13.—Adcock Aerial

coil in the direction A to D. At the same time an e.m.f. of the same magnitude and phase will be induced into limb BC, causing a current to flow in the opposite direction producing no output in the secondary.

The use of the Adcock Aerial is limited by the fact that it is essentially a "one-turn loop", and therefore not suitable for many requirements because the output is directly proportional to the effective loop area and the Adcock aerial must be large compared with a multi-turn loop in order to produce a reasonable output.

11.8. The Goniometer

It is not always convenient to rotate a loop or Adcock system, and to avoid this it is possible to obtain the effect of rotation by combining the outputs of two crossed loops or Adcock Aerials in a "Goniometer" (angle measuring instrument) (Fig. XI.14). The radio-goniometer consists of two sets of primary coils at right angles to each other with both sets mutually coupled to a secondary coil which may be rotated through 360°. There is a pair of primary coils associated with each loop aerial. If the radio-goniometer is so constructed that the e.m.f. from the secondary coil is proportional to the cosine of the angle through which it is rotated, then the directional characteristics of the system will be identical

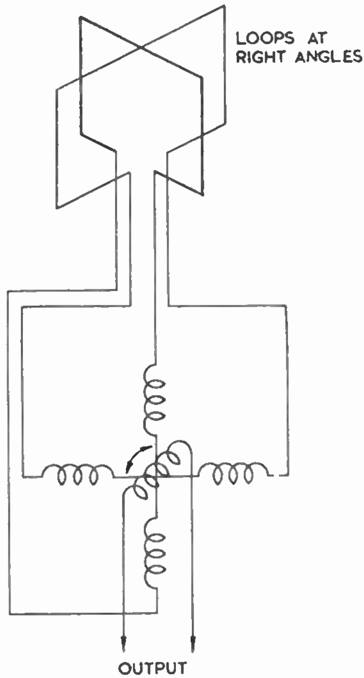

FIG. XI.14.—Loop Aerials with Radio-goniometer

with a rotatable loop or Adcock Aerial. Radio-goniometers and the associated apparatus are usually set up (having a scale marked from 0° to 360° over which a pointer attached to the moving part is free to pass), so that the scale reads 0° when a wave known to be incident from due North produces zero signals in the receiver. Subsequent measurements are then relative to this reading, which may be read directly from the scale and plotted on maps of suitable projection.

Questions

1. Explain why a loop aerial has directional characteristics and describe a practical direction-finding system using loop aerials. (C. & G., 1947.)

2. Derive an expression to give the e.m.f. acting around a rectangular loop aerial whose dimensions are small compared with the wavelength received.

Use your result to calculate the e.m.f. acting in a twenty-turn loop 50 cm. high and 30 cm. wide and whose plane is inclined at 60° to the direction from which a 600-kc/s signal arrives. The signal strength at the loop is 2 mV/metre.

Answer: 37·7 μV.

3. A vertical aerial without a top has the same length as the vertical members of a rectangular single-turn loop aerial. It is found that the signal from a transmitting station is stronger on the vertical aerial than on the loop when the latter is in a position for maximum reception. Explain the reasons. (C. & G., 1946.)

4. A loop aerial consists of twenty turns of wire on a square of 50 cm. side. A transmission on 400 metres wavelength produces a series e.m.f. of 1·5 mV in the loop when the latter is in the direction of maximum response. What is the field-strength of the transmission? (C. & G., 1949.)

Answer: 19·1 mV/metre.

5. Given a loop aerial of ten turns whose dimensions are 40 cm. × 70 cm. inclined at 30° to the direction from which a 1-Mc/s signal arrives; find the e.m.f. acting in the loop if the signal strength at the loop is 30 μV/metre.

Answer: 1·53 μV.

6. Explain how a frame aerial can be used for direction finding. A frame aerial consists of ten turns, each 1 metre square, is situated in a field of 10 mV/metre and wavelength 300 metres. Find the maximum e.m.f. that can be induced in the frame. (I.E.E., May 1936.)

Answer: 2·09 mV.

7. Find the maximum e.m.f. induced in a frame aerial 2 ft. high by 1 ft. wide and having twenty turns of wire by a signal of wavelength 800 metres and electric field strength of 3 mV/metre. Prove any formula used. (L.U., 1938.)

Answer: 87·6 μV.

8. Explain the principle of direction finding and the cause of night errors.

Explain how an open aerial and a loop can be combined to give a heart-shaped polar diagram for reception. (I.E.E., May 1931.)

9. Describe a method whereby the direction and sense of a distant transmitting station can be determined. (I.E.E., May 1933.)

10. Find an expression for the e.m.f. induced in a square vertical frame aerial as it is rotated about a vertical axis in a plane electro-magnetic wave. Explain how direction-finding by wireless is achieved. Discuss briefly the errors which may occur and the methods of minimising them. (I.E.E., November 1935.)

11. A tuned loop aerial encloses an area of 0·5 sq. metre and consists of thirty turns, the Q-factor being 20 at 2 Mc/s. Calculate the e.m.f. developed across the tuning capacitor when the aerial is adjusted for maximum response at 2 Mc/s in a field of strength 0·1 mV/metre.

Answer: 1·26 mV.

12. State and derive a formula for the e.m.f. E induced in a vertical rectangular loop aerial, length of sides a and b, number of turns N, by a plane vertically polarised, electro-magnetic wave of field strength F, the angle between the plane of the loop and the wave direction being θ. (C. & G., 1952.)

13. Describe a simple direction-finder using a rotating loop aerial. What is meant by " sense ambiguity " in such a direction finder, and how can it be avoided? (C. & G., 1953.)

Specimen Answer

Q. A vertical aerial without a top has the same length as the vertical members of a rectangular single-turn loop aerial. It is found that the signal received from a distant sender produces a greater output in the receiver with the vertical aerial than with the loop aerial in its position of maximum response. Explain why this occurs.

A. Fig. XI.Q.1. shows a vertical aerial of height h metres and a loop aerial with vertical sides h metres and horizontal sides d metres in length. Let both aerials be located in the same electric field of strength ξ volts/metre. Assuming this field to be vertically polarised, the e.m.f./E_v induced in the vertical aerial is given by :

$$E_v = h\xi$$

Since the electric field is vertical, no e.m.f. will be induced in the horizontal sides of the loop aerial. In each vertical side of the loop an e.m.f. will be induced of r.m.s. value $h\xi$. These e.m.f.s will in general be out of phase because the two sides of the loop are at different distances from the sending station. Assuming a medium-wave sender, the wavelength of the signal will be much greater than the width d metres of the loop, and the position of

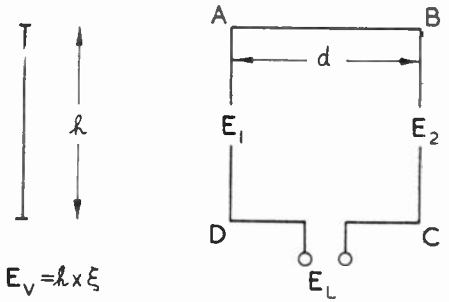

FIG. XI.Q.1

maximum response will be that with the plane of the loop parallel to the direction of propagation. This condition is illustrated by Fig. XI.Q.2.

Let the e.m.f. induced in the vertical sides be E_1 and E_2 r.m.s. volts. Since the same field passes through each side of the loop as the wave progresses, these e.m.f.s will have the same r.m.s.

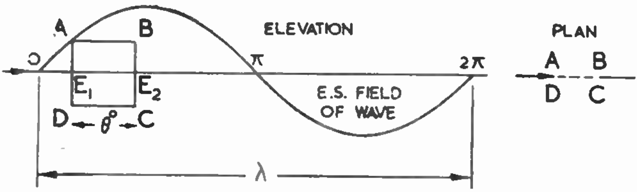

FIG. XI.Q.2

value $E = h\xi$, but will be out of phase by an angle $\theta = \frac{d}{\lambda}(2\pi)$ radians as shown in Fig. XI.Q.3. The instantaneous directions of the e.m.f.s will be the same in space, but will be opposing round the loop, thus the resultant e.m.f. E_L is given by the vector difference of the two components.

Referring to the vector diagram (Fig. XI.Q.3),

$$E_L = 2E \sin \theta/2$$

$$= 2h\xi \sin \frac{\pi d}{\lambda}$$

For a medium-wave sender and a loop of practical dimensions, $\lambda >> \pi d$ and $\frac{\pi d}{\lambda}$ is very small.

$$\therefore \quad \sin \frac{\pi d}{\lambda} = \frac{\pi d}{\lambda} \text{ approximately}$$

$$\therefore \quad E_L = \frac{2h\xi\pi d}{\lambda}$$

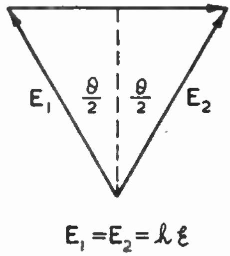

FIG. XI.Q.3

Thus the ratio of the signals received by the two aerials is $\frac{E_L}{E_v} = \frac{2h\xi\pi d}{h\xi . \lambda}$ or $E_L = \frac{2\pi d}{\lambda} . E_v$, and since $\frac{\pi d}{\lambda}$ is very small the signal received on the loop aerial will be only a small fraction of that received on the vertical aerial. (P.O. Eng. Dept.)

CHAPTER 12

RADIO COMMUNICATION

MUCH has been accomplished in the realm of radio communication since Marconi's early experiments with transatlantic telegraphy in 1901. About fourteen years later speech was first transmitted by radio from Virginia, U.S.A., to Paris, France. However, it was not until 1927 that the first public radio-telephone service was opened between London and New York. The carrier frequency was 60 kc/s (5,000 metres), and this circuit is still in use today. This was followed a year later by a short-wave circuit, and the development and improvement of both types has constantly engaged the attention of radio engineers.

In this chapter we shall consider the basic requirements of a two-way radio-telephone link and discuss means of overcoming difficulties involved in its operation. A typical Post Office circuit and terminal will be described.

12.1. Radio-telephone Circuits, Basic Requirements

At each end of a circuit there must be a radio transmitter and receiver; the two are usually geographically separated to prevent the transmitter from interfering with the receiver at the same end of the circuit. The two radio links provide the means of two-way communication, each path being uni-directional. The circuit is completed by connecting the transmitter and receiver by means of one-way land lines to a common point. At this common point the terminal apparatus must be capable of combining the two paths and connecting them to the normal two-way telephone system of the country. Apart from this function, facilities must be provided for signalling and for testing and lining up the radio circuit.

The carrier frequency for two-way operation of radio circuits may be the same for both send and receive directions (as in the long-wave system between London and New York), or separate bands may be employed as in most short-wave circuits in use today. A short-wave circuit in its simplest form employing two separate frequency bands is shown in Fig. XII.1. This shows that the complete system consists of a send amplifier and radio transmitter, a receiver and receiver amplifier and a hybrid coil, this equipment being duplicated at the distant end. Before we go on to describe the actual operation of a system, we must consider the hybrid coil in greater detail.

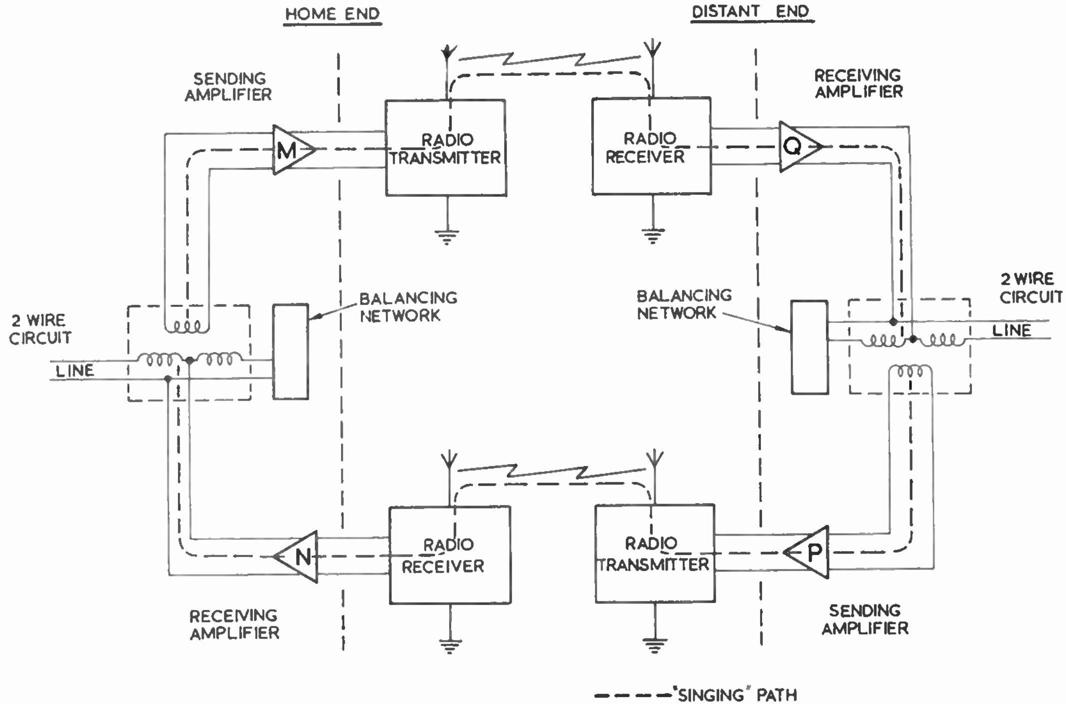

Fig. XII.1.—Diagram of Simple Two-way Radio Telephone

12.2. The Hybrid Coil

In this Volume a brief analysis of the hybrid coil and its use in radio-telephone terminal equipment is given. Fig. XII.2 shows the circuit of the hybrid coil. When $Z_1 = Z_2$ and the two wind-

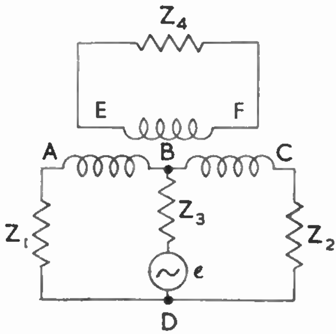

FIG. XII.2.—The Hybrid Coil

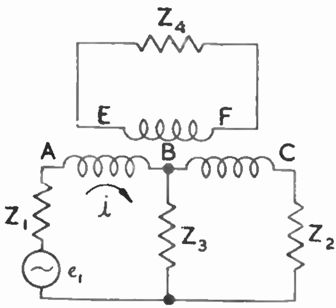

FIG. XII.3.—Hybrid Coil

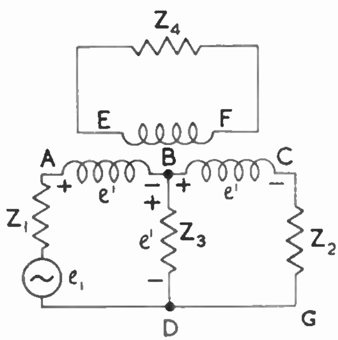

FIG. XII.4.—Potentials in Hybrid Coil

ings AB and BC are equal, currents flowing as a result of the generator e will produce zero resultant flux in the core. Hence there is no induced e.m.f. in the winding EF. This means that there is no interaction or transfer of energy between Z_3 and Z_4.

Another condition we have to consider is an e.m.f. in series with Z_1 (Fig. XII.3). A current will be established around the circuit Z_1, Z_3 and winding AB of the transformer. The current flowing in winding AB of the transformer will induce into winding BC an e.m.f. equal in magnitude and phase to the induced or " back " e.m.f. in winding AB. Now if the impedance Z_4 seen through the impedance transformation $AB : EF$ of the transformer equals the impedance Z_3 then equal p.d.s will appear across the points AB and BD in the network.

Taking instantaneous values to show the direction of the e.m.f.s and p.d.s, they appear as shown in Fig. XII.4. It will be seen that in the circuit $BCGD$ two equal and opposing p.d.s exist so that no current flows in Z_2 due to the generator e_1. This condition is a result of the equality (and direction of winding) of the windings AB and BC, and equality (after the transformer impedance transformation) of Z_3 and Z_4. With this condition it means that no transfer of energy occurs between Z_1 and Z_2. It will be noted that the impedance transformation is dependent on the turns ratio between windings AB and EF (or BC and EF).

12.3. Practical Use of the Hybrid Coil

The hybrid coil is used to combine two outputs into one or vice versa. This is also known as converting a four-wire system into a two-wire.

Fig. XII.1 shows that it is undesirable to have transmission (without considerable attenuation) round the circuit Sending Amplifier M, Receiving Amplifier Q, Sending Amplifier P, Receiving Amplifier N, Sending Amplifier M. In other words, there should be negligible transfer of energy between the output of the receiving amplifier and the input to the sending amplifier. This can be achieved if the impedance of the line (denoted by Z_1 in Fig. XII.2) is simulated by the balancing network (denoted by Z_2 in Fig. XII.2). Under these ideal conditions half the energy from the receiving amplifier will be dissipated in the balancing network and half usefully employed in being sent to the two-wire circuit. If the output impedance of the receiving amplifier is made equal to the input impedance of the sending amplifier (after allowing for the impedance transformation of the hybrid coil), then half the energy received from the two-wire circuit will be passed to the input circuit of the sending amplifier while the other half will be dissipated in the output circuit of the receiving amplifier. If these conditions are not met, some of the energy from the two-wire circuit will appear in the balancing network.

We are particularly interested here in the use of the hybrid coil in radio-telephone terminal equipment ; it has many other applications in the field of Telecommunications.

12.4. Transmission Characteristics

Consider a subscriber speaking from the " Home " end of the two-way radio-telephone system shown in Fig. XII.1. The speaker's voice produces electric currents which pass along a land line to the radio terminal equipment. Here the speech currents pass into a hybrid coil dividing (ideally) equally, between the send-amplifier input (useful energy) and the output of the receiver amplifier (wasted energy dissipated in the output impedance of the amplifier). The send amplifier amplifies the speech currents and passes them via a land line to the transmitting station. Here the speech currents modulate the radio-frequency amplifier and are radiated from the aerial. The distant receiver picks up the transmitted wave and demodulates it back into speech currents, which are passed along a land line to the distant terminal equipment. Here the speech currents pass through a receiving amplifier, which acts as a buffer stage, and then to a hybrid coil, half the energy being lost in the balancing network and half usefully employed in going to the distant subscriber via a land line in the normal telephone network of the country. Because there cannot be perfect balance between the balancing network and the subscriber's line, part of the energy will be returned via the lower transmission path to the " Home "-end subscriber as echo. Similar conditions apply when the " Distant "-end subscriber speaks to the " Home "-end subscriber.

Since the long-wave band is very crowded it is usual to operate both paths of a long-wave radio-telephone system on the same carrier frequency, rather than on different carrier frequencies as is usual in the short-wave band. This means that there are common paths for transmission of signals via transmitter and receiver at both ends of the circuit. Although the transmitter and receiver are geographically remote from each other at each end, the coupling obtained is large enough to cause undesirable effects which will be discussed later.

Considering the short-wave system again, providing the radio circuit is stable and free from noise, the balance provided by the hybrid and balancing network will prevent instability around the circuit, and " singing " will not occur. " Singing " is the term given to the condition of a circuit when the total gain round a loop exceeds the attenuation and the excess gain causes instability rather like the oscillations set up in amplifiers due to inadequate decoupling (Volume I, Chapter 5). If, however, the radio circuits (or land lines) are long, then delayed " echoes " set up as described above will become increasingly annoying. To overcome this, voice-operated echo suppressors are used which disable the return path taken by the delayed echoes while the other path is transmitting.

Another consideration is the different speech levels to which the circuit will be subjected by subscribers. Since amplitude-modulated systems are used, it is desirable for the level at the input to the transmitter to be as constant and as high as possible, maintaining full modulation at all times. This gives a high ratio of desired signal to unwanted noise from the radio circuit. Constant volume level at the transmitter input for all types of subscriber may be accomplished manually (by operators) or electronically. The early installations employed the former method, whilst the trend nowadays is to use the latter.

Short-wave circuits are subject to fading, which may be of two kinds. The entire band of frequencies may fade simultaneously; the automatic gain control (A.G.C.) of the receiver will compensate for this and maintain the received signal substantially constant. Other types of fading, however, cause certain frequencies in the transmitted band to fade independently of other frequencies within the same band. This " selective fading " is not fully compensated by A.G.C. action, and over short periods considerable variations of received level may be experienced.

Because of these comparatively rapid adjustments to the gain of radio-telephone systems, and despite the action of the hybrid coils and some types of " echo suppressors ", " singing " may frequently occur. One way of reducing the risk of " singing " would be to introduce attenuation in the receiving path whenever the gain of the transmitting path is increased. This, if done manually, would set an almost impossible task, even for the most highly skilled operators, apart from causing inconvenience to the receiving subscriber by the changes of received volume. On common frequency working, " singing " may also occur over the local radio transmitter and receiving paths, so that it is imperative that other means are provided to prevent this happening and so maintain optimum conditions for transmission.

12.5. Voice-operated Device Anti-singing (V.O.D.A.S.)

Fig. XII.5 shows one end of a two-way radio-telephone system employing a simple form of " vodas " at each end. The vodas consists of transmitting and receiving delay circuits with detectors and associated relays. The condition of " no speech being transmitted " is given in Fig. XII.5, and the operation of the vodases is as follows. Relay A open-circuits the transmitting path so that no " singing " may occur around the whole circuit, or round the local radio circuit and terminal. Relay B is closed, and speech may pass its contacts. When a subscriber from the " Home " end speaks, the voice currents pass through the terminal equipment to the Transmission Detector and Delay Circuit. Whilst the voice currents are traversing the Delay Circuit, relays A and B operate (provided relay C has not been operated previously). When relay A operates the voice currents are passed on to the transmitter, while the operation of relay B opens the receive circuit, thus preventing " singing " and echoes which would otherwise occur when relay A is operated. When the voice currents arrive at the distant end they pass through the Reception Detector at that end operating relay C, which avoids operation of relays A

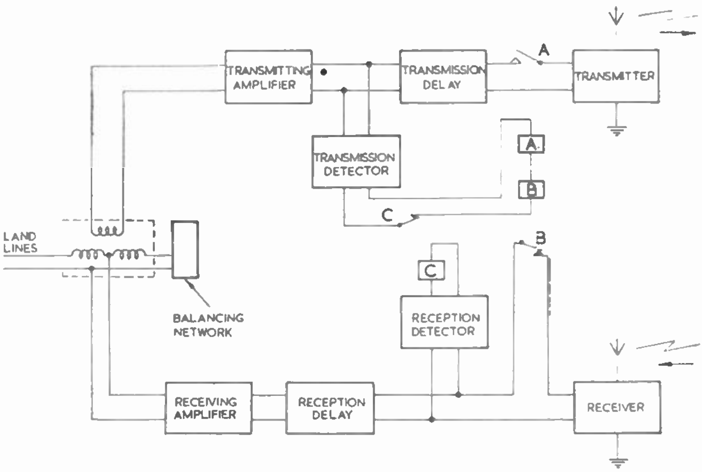

FIG. XII.5.—Operation of Anti-singing Device

and B by echoes of received-speech currents. Delay of the received speech is made long enough by the Reception Delay Circuit to ensure complete operation of relay C before any part of the received signal can reach the Transmission Detector via the Hybrid Coil and Transmitting Amplifier. The relays return to normal when the subscriber stops speaking.

In practice, it is necessary to prevent the operation of a vodas by noise. This is achieved by the use of delay circuits, amplitude discrimination (of speech relative to noise) and frequency discrimination. The latter is achieved by filters in t[illegible]tor circuits which attenuate frequencies not essential [illegible]le speech.

12.6. A British Radio-telephone System

We shall now describe a typical modern two-way radio-telephone system operated by the British Post Office. Fig. XII.6 is a block schematic showing the terminal equipment used for operation of a two-way radio-telephone system. It has already been noted that speech signals from subscribers arrive at the radio terminal at widely different average levels. It is desirable to maintain the modulation depth of the transmitted A.M. signal as high and as constant as possible. This is achieved by electronic methods using in the transmitting path an automatic variable-gain amplifier which maintains a constant output over a wide range of input volumes. This amplifier will "reduce" an input range of 100 to 1 in amplitude to an output range of 2 to 1, at a higher level and is labelled "transmit constant volume amplifier" (T.C.V.A.) in Fig. XII.6. The T.C.V.A. has different characteristics for increasing and decreasing gain. When the speech input increases the output stabilises in about 20 milliseconds, but when the mean speech level decreases the gain remains practically constant for about a second, then rises slowly until the output stabilises in about 5 seconds. These characteristics produce an output substantially constant during passages of speech.

The counterpart of the T.C.V.A. on the receive side of the equipment is the "receive constant volume amplifier" (R.C.V.A.), and this corrects for short-term variations in received level not compensated by the A.G.C. The Automatic Gain Control in the radio receiver corrects for the slower general fading. The R.C.V.A. gives an output range of about $1\frac{1}{2}$ to 1 for an input range of 10 to 1.

Speech from the subscriber at the "Home" end passes via the International Exchange to the Hybrid Coil at the Radio Terminal building via a land line. Here it is amplified by the T.C.V.A. so that the output volume is sensibly constant. The speech then passes via static relays (T.S.R. 1 and 2) to another land line connected to the radio transmitter. The static relays in the send and receive path consist of metal rectifiers and form part of the "singing suppressor". Their resistance, and hence loss, is determined by the direction and amount of direct current flowing which in turn is determined by the relative levels at the points A and B on Fig. XII.6. (The use of two relays is necessary to provide Privacy facilities, which will not be described here as they are not a fundamental requirement.)

In t ite direction, signals received by the radio receiver a land line to the Radio Terminal, where they pass V.A., which provides a nearly constant volume at

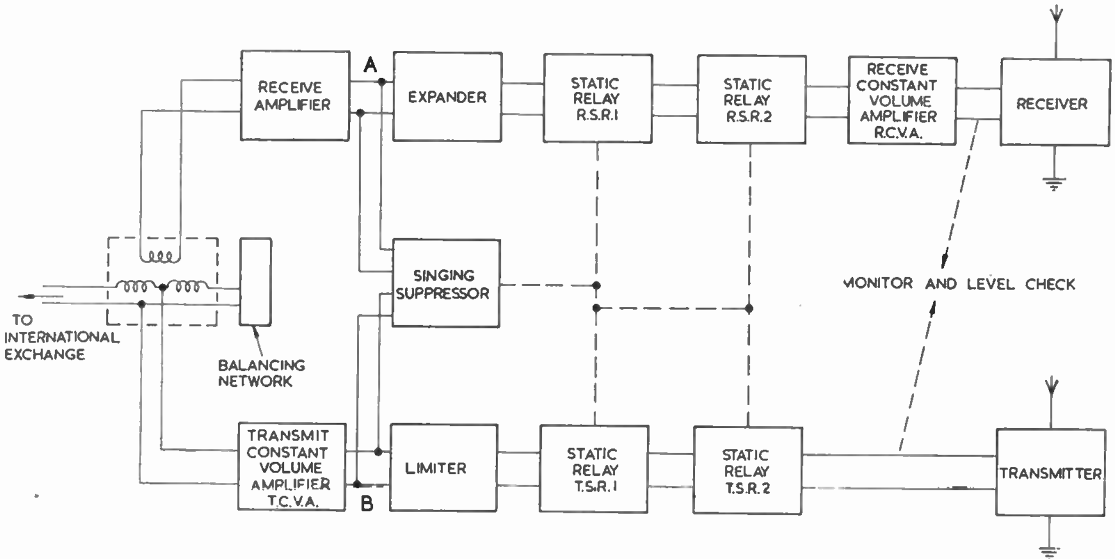
RECEIVE AMPLIFIER
A
EXPANDER
STATIC RELAY R.S.R.1
STATIC RELAY R.S.R.2
RECEIVE CONSTANT VOLUME AMPLIFIER R.C.V.A.
RECEIVER
SINGING SUPPRESSOR
MONITOR AND LEVEL CHECK
TO INTERNATIONAL EXCHANGE
BALANCING NETWORK
TRANSMIT CONSTANT VOLUME AMPLIFIER T.C.V.A.
B
LIMITER
STATIC RELAY T.S.R.1
STATIC RELAY T.S.R.2
TRANSMITTER

Fig. XII.6

its output, having removed the fluctuations in volume due to rapid fading. From here the speech is passed by two static relays in the receive path (R.S.R. 1 and 2) and thence on to a noise reducer or limited range expander. The expander is essentially an amplifier with a 3·5 : 1 expansion characteristic below a predetermined datum level; signals above this level are reproduced by the amplifier linearly. By arranging that speech signals are above datum, which itself is several times the worst expected noise level, the performance of the circuit is greatly improved under poor transmission conditions. The time taken for the output level to stabilise after a change in input is made to be about 20 milliseconds, so that the expander is able to cater for changes at syllabic rate. The expander is followed by a receive amplifier, whose main function is to prevent signals from the International Exchange reaching the expander output. These signals are present due to the imperfect balance of the hybrid coil noted earlier. In the condition of " no speech transmitted " it is arranged that the receive static relays have low loss and the ones in the transmit path have a high loss. When a subscriber speaks from the " Home " end, some of this speech is amplified by the singing suppressor amplifier and then rectified. The rectified current operates a thermionic relay which reverses the direct current through the static relays, reducing the attenuation in the transmit path and increasing it in the receive path. This operation is completed in about 2 milliseconds, and there is negligible " clipping " of the initial passage of speech. The singing suppressor takes about 180 milliseconds to restore itself to the original condition after the cessation of the signal. When speech is being received in the receive path, a feed is taken from the output of the expander, amplified, rectified and applied as direct current to reduce the gain of the singing suppressor amplifier. The receive path has a hang-over time for a period of 60 milliseconds, which prevents false operations due to echo signals from the land line. Because noise on the radio circuit may also reduce the gain of the singing suppressor amplifier and so make it impossible for normal-volume speech signals on the transmit side to operate the thermionic relay, the sensitivity of the control amplifier is made adjustable to suit the conditions existing on the radio circuit at any given time.

Summary

For the successful operation of a two-way radio-telephone system anti-singing devices are required at both ends of the circuit to :

1. Suppress echoes and prevent singing which would result from instability.

2. To enable adjustments to be made at one end without reference to the other end.

12.7. Noise in Radio Links

Noise in radio links of any kind is of many types :

1. *Noise in Transmitters.* This may be hum, caused by insufficient smoothing of high-tension supplies or by a.c. heating of valve cathodes. Mechanical vibration of components may also cause sympathetic variations of frequency or output. Both these types of noise may be made negligible by careful design.

2. *Noise in Receivers.* Hum (as in transmitters) and microphony of valves and components are significant sources of noise. In superheterodyne receivers mixer noise is significant at low

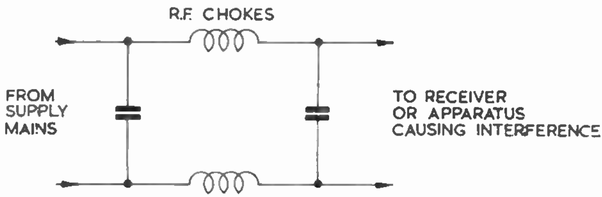

FIG. XII.7.—Simple Interference Filter

signal levels, and is overcome by using signal-frequency amplification preceding the mixer. The layout and construction of the receiver must also be good when small signals are to be reliably received. Circuit noise caused by thermal agitation of conductors becomes an important factor when handling small signals, and the best solution is to use a tuned circuit of high Q at the receiver input. The use of a "T.C.V.A." at the transmitter to maintain a high level of modulation assists in overcoming the latter type of noise.

3. *External Sources of Noise.* These may be natural or artificial. Natural sources are principally electrical storms, while artificial noises are generated by any electrical equipment where sparks occur, such as commutator-type motors, vibrators, petrol-engine ignition systems, flashing signs, etc. The effects of natural noise can be minimised by reducing the bandwidth of the receiver to the smallest acceptable on the basis of intelligibility of speech and by using limiting circuits which prevent impulsive signals from exceeding the normal maximum level of desired signals. Artificial noise is best suppressed at the source by resistor–capacitor spark

quenchers or choke–capacitor filters in the supply leads to the offending equipment (Fig. XII.7). In the case of mains-borne interference a choke–capacitor filter at the power input to the receiver is frequently effective. In all cases, the receiving aerial should be situated as high as possible and away from local sources of noise.

Questions

1. Explain, with the aid of a block schematic diagram, the general arrangement of a two-way radio-telephone system. (C. & G., 1947.)

2. Describe the causes of the various classes of noise encountered in a radio receiver. (C. & G., 1946.)

3. Give an outline description with block diagrams, of a medium-frequency amplitude-modulated radio-telephony transmitter. (C. & G., 1948.)

4. Describe, with the aid of a block schematic diagram, a simple radio-telephone equipment for communication between two vehicles over a distance of up to 20 miles. (C. & G., 1949.)

5. Explain briefly how the ignition system of a car may cause interference with radio reception and state how the interference can be minimised. (C. & G., 1949.)

6. The input to a radio-telephone transmitter modulator has a speech bandwidth of 6,000 c/s. Reception is made by a super-heterodyne receiver in which the intermediate-frequency amplifier has a total bandwidth of 6,000 c/s. Would you expect to obtain faithful reproduction of the original modulation? Give reasons for your answer. (C. & G., 1943.)

7. What are the chief sources of interference with radio broadcast reception in Great Britain? State briefly how they can be minimised. (C. & G., 1938.)

8. How could an appreciable reduction be effected in the interference with radio reception due to: (*a*) mains hum; (*b*) medical appliances; (*c*) therapy apparatus. (C. & G., 1945.)

9. Discuss the ways in which electrical apparatus can produce interference in a receiving aerial. Indicate the methods by means of which such interference can be minimised at the source, and state what requirements the components must satisfy.

10. Describe, in outline, the arrangement of a long-distance radio-telephone circuit which is connected into the international telephone network. (I.E.E., April 1949.)

11. Explain the purpose of each of the following items used in a two-way radio-telephone link: (*a*) hybrid transformer; (*b*) line-balancing network; (*c*) singing suppressor. (C. & G., 1952.)

12. State the sources of noise that may affect a long-distance point-to-point short-wave radio link, and say what can be done to minimise the interference produced. (C. & G., 1952.)

13. Enumerate the main sources of noise in the reception of long- and medium-wave broadcast signals, and say what can be done at the transmitter and the receiver to minimise the interference so produced. (C. & G., 1953.)

Specimen Answer

Q. Discuss the ways in which electrical apparatus can produce interference in a receiving aerial. Indicate the methods by means of which such interference can be minimised at the source, and state what requirements the components must satisfy.

A. Interference from electrical apparatus may be propagated in two ways. In the first case radio-frequency oscillations occurring in parts of the circuit may cause energy to be directly radiated so that it is accepted by a receiving aerial; alternatively, the radiated energy may be picked up by neighbouring conductors and re-radiated. In the second case radio-frequency currents may be propagated along the supply mains and radiated by the house wiring to the receiving aerial. Propagation along the supply mains may occur either as symmetrical currents which flow round the loop or as asymmetrical currents which flow outwards over one or both conductors and return to earth via the capacitance of the mains. Neutral conductors which are earthed at the supply station cannot be regarded as being at earth potential throughout their lengths at radio-frequency voltages.

Interference can be minimised at the source by screening the components and by fitting radio-frequency filters as near to the source as practicable. Where the propagation is effected over the supply mains the interference can be minimised by the insertion of the appropriate type of radio-frequency filter in the mains as close to the source of interference as possible. If the interfering wave is propagated asymmetrically, suppression can be achieved by fitting a low-pass filter between the source and the mains. Low-frequency currents pass through this filter with little attenuation, but radio-frequency currents are greatly attenuated and so prevented from reaching the mains.

Fig. XII.Q.1 illustrates a simple low-pass filter. In order, however, to suppress radio-frequency interference propagated both symmetrically and asymmetrically it is necessary to use a symmetrically designed filter as shown in Fig. XII.Q.2. The inductors present a high series impedance and the capacitors a low shunt impedance to radio-frequency currents.

The method of connexion depends upon the relative magnitudes of the source impedance and the mains impedance. For maximum suppression the shunt path must be connected directly across the higher of these two. If connected across the lower, the shunting effect on the radio-frequency currents will necessarily be

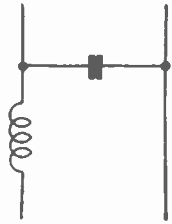

FIG. XII.Q.1

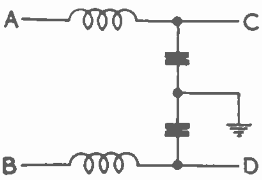

FIG. XII.Q.2

smaller. Hence if the source impedance is greater than the mains impedance, the source is connected to C.D. and the mains to A.B. and vice versa.

If both mains and source impedances are high, then adequate suppression is generally obtained from the shunt path alone and the inductors may be omitted.

If both impedances are low, then the shunt path contributes little to the total suppression, and adequate suppression is obtained by use of the inductors alone.

The capacitors used must be capable of withstanding the voltages applied to them, and this is ensured by specifying minimum " test " voltages to be applied by the manufacturer. Each capacitor must have a low self-inductance and be fitted with a fuse having a rating not greater than 3 amps.

The inductors must be constructed to carry the full-load current without over-heating. They must have low self-capacitance and be adequately insulated from earth.

(P.O. Eng. Dept.)

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