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ATOMIC ENERGY FOR INDUSTRIAL USES

THE Trueman-Wood Lecture given by Professor M. L. Oliphant, F.R.S., to the Royal Society of Arts last month on the Industrial Applications of Atomic Energy has done much to bring this aspect of atomic energy into a bright perspective in contrast to the gloom caused by the recent announcement of the new hydrogen bomb and its military consequences. At the same time it dispels many popular illusions held by the layman regarding the non-military applications of atomic energy.

Professor Oliphant gives us a glimpse of the not-too-distant future when all our electrical energy will be generated from nuclear reactors and when, he says, the present-day restrictions on the consumption of electricity will be replaced by a positive urge to use it, and electric clocks will really tell the time.

Power engineers may be inclined to disagree with Professor Oliphant's statement that the cost of coal in the present day power station is about one-third of the charge to the consumer and his inference that if the fuel were free, the cost of electrical energy would be reduced by 30 per cent. In point of fact, the cost of coal for a modern efficient power station is about 10 per cent of the unit charge, while the remaining 90 per cent is absorbed in distribution costs and depreciation charges.

It is true that when nuclear power stations have replaced the present power stations the unsightly coal dumps, coal- and ash-handling equipment will disappear for ever, but it is unlikely that electricity will be appreciably cheaper.

The change-over is by no means the simple one of replacing the coal-fired boiler by a reactor, and, as Professor Oliphant points out, there are many problems still to be solved.

By no means the least of these is the satisfactory disposal of the fission products. These, according to him, are equivalent in activity to many tons of radium, and if buried or thrown down a disused mine, for example, they may reappear in underground water supplies with disastrous results. Moreover, the uncertainty of ocean currents renders it hazardous to dump the fission products into the sea, even into deep sea when sealed in containers.

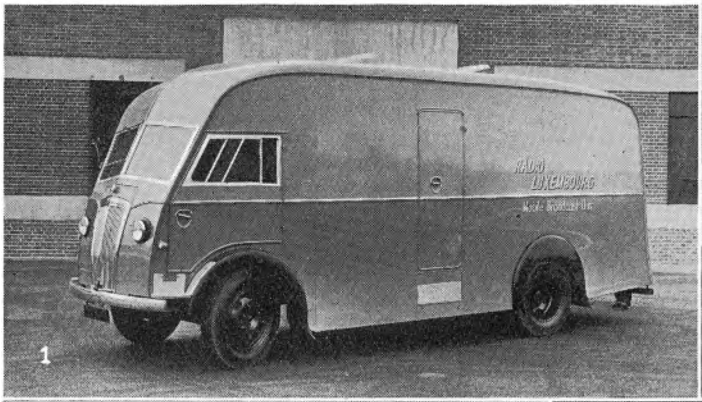
However, these are technological problems which are not declared insuperable, and already the construction of two reactors in the U.S.A. has been announced by the Atomic Energy Commission for experimental purposes.

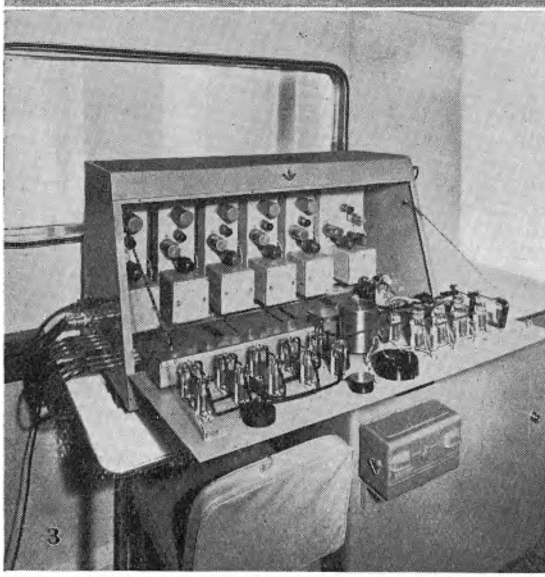
Ultimately, of course, the possibilities of hydrogen as a nuclear fuel must be considered, for when the methods of producing a controllable reaction have been developed, the ultimate substitute for coal will have been found.

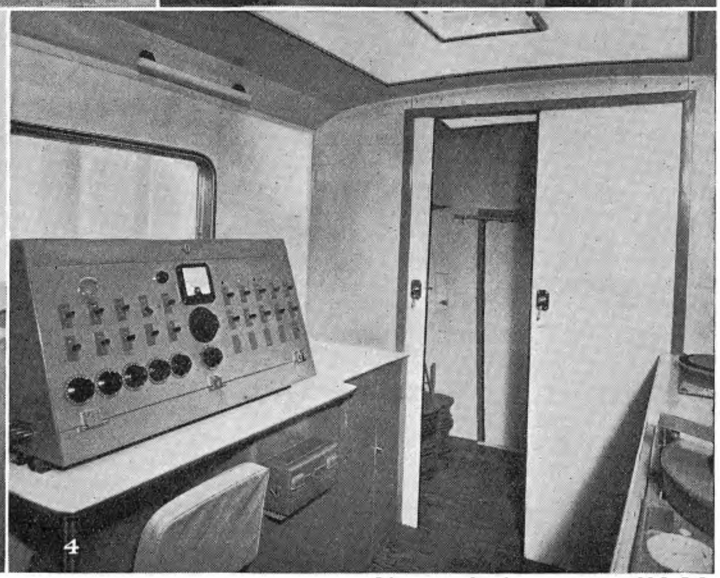
Some conception of the enormous amount of energy available can be gathered from the statement that 1 lb. of hydrogen if transformed into helium would produce 100 million kilowatt hours of heat energy; and calculations show that there is enough hydrogen in the sea to provide energy for 1,000 million million years!

However, this is a remote possibility in the light of existing knowledge and the more immediate and urgent problem is that of the military aspects of atomic energy. This rests with mankind in general and not with the scientists alone.

A New Mobile Recording Van for Radio-Luxembourg







[*Photos supplied by courtesy of E.M.I. Ltd.*

A recent noteworthy export is a new type of Broadcast Recording Van which has been supplied by E.M.I., Ltd., to the order of Radio Luxembourg.

The main items of equipment comprise two E.M.I. magnetic tape recorders and a disk recorder and replay desk for operation at either 78 or $33\frac{1}{3}$ r.p.m.

The main control panel comprises a five channel microphone pre-amplifier with individual fading, a programme meter and telephone or loudspeaker monitoring facilities. This control console is designed to give the operator complete flexibility of control and interconnexion of the various items of equipment; full dubbing, mixing and cross recording facilities are available, enabling any type of studio work to be carried out. In addition, the complete console can be lifted out and carried to the recording site when the distances involved make it undesirable to run long microphone cables.

Two high quality loudspeaker units are provided for remote replay purposes on the recording site. Communication between the van and the recording site is by a "talk-back" system. A.C. power supplies for operational, heating and lighting requirements are taken on site but stand-by lighting may be operated from the van's electrical system.

The chassis is a 5 ton long wheel base Morris, the bodywork of which was constructed by O. G. Lywood, Ltd.

Details of the van may be seen in the accompanying photographs :—

1. An exterior view of the Mobile Recording Van.

2. A general view of the Recording Room. On the right is seen the Engineer's control panel and behind, the two E.M.I. Tape Recorders, and two of the five microphones. On the left is the disk recorder and replay desk. The separate entrance to this room is forward of the disk recorder desk.

3. An inside view of the control panel.

4. A close-up of the Engineer's control panel. The overall level control and meter is seen at the centre and the five separate microphone fading controls are at the left. The rear compartment of the van houses the cables, cable entries and drums.

Microwave Lenses

By J. BROWN, M.A., and S. S. D. JONES, M.A.*

(Part I)

This is the first of a series of articles on Microwave Lenses in which a survey of the various types will be presented. Particular attention will be paid to the theory of operation of the artificial dielectrics which have been developed for such lenses. Design equations for lens contours will be given and the various aberrations to which lenses are subject will be considered.

Analogies with Optics

WITH the development of centimetric waves, it has become possible to apply the techniques of optics to the field of radio. The first practical application of such techniques has been the wide-spread use, particularly in radar systems, of reflectors such as paraboloids and parabolic cylinders (cheeses) to produce narrow beams. In fact, however, the possibility of focusing radio waves by means of lenses has been realised since 1889, when Sir Oliver Lodge[1] demonstrated the beaming effect of a lens constructed of pitch on the radiation from a spark-gap oscillator. No serious attempt to use lenses as practical aerial systems was, however, possible until convenient sources of power at centimetric wavelengths became available about 1940. The first such attempt was that of Rust,[2] who in 1942 made use of a lens to correct the wave fronts emerging from an electromagnetic horn. Since then a variety of types of lens has been developed both in this country and in the United States. Several of these types of lens are now widely used in the United States, particularly as the aerial systems of repeater stations in centimetric radio links.

Although the analogy with optics has been of the greatest value in developing microwave lenses, there are important differences which must never be overlooked. Firstly, even the smallest optical lens has an aperture whose dimensions are approximately one million times the wavelength of light, whereas the largest radio lens so far constructed has an aperture of about five hundred times the wavelength at which it is used. It follows that the effects of diffraction will be of the greatest importance in determining the behaviour of radio lenses, whereas such effects may usually be neglected as far as optical lenses are concerned. On the other hand, microwave lenses are at present, at any rate, only being used for the production of narrow beams—this is equivalent to saying that either the object or the image is at infinity. The significance of this is that the question of obtaining depth of focus, a severe limitation in the design of optical lenses does not arise. In optics this limitation is upon the numerical aperture or F number, i.e., the ratio of the focal length of the lens to the diameter of its aperture, and it is seldom that the F number of an optical lens is less than 2. Microwave lenses of F equals 1 are, however, quite common and indeed the F number of such lenses has never to the knowledge of the authors exceeded 1.5.

* Radar Research and Development Establishment, Ministry of Supply.

The most important parameter in optical lens design is the refractive index of the material from which the lens is constructed. The usual definition[3] of this quantity is in terms of the angles of refraction (r) and incidence (i) when a beam is refracted at the interface between free space and the lens material (Fig. 1).

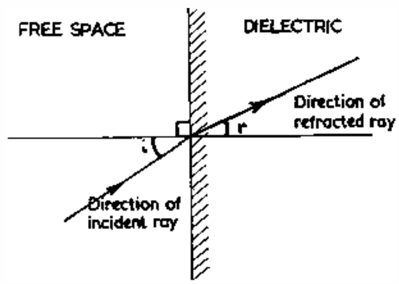


Fig. 1. Refraction at a Dielectric Surface

The refractive index (n) of the lens material is given by Snell's Law which states that

$$n = \sin i / \sin r \qquad (1)$$

A more general definition[3] of refractive index may be given in terms of the phase velocities in free space and in the lens material v_0 and v respectively. It is

$$n = v_0 / v \qquad (2)$$

This definition is the more useful for radio work as the phase velocity of all the lens materials to be considered may be readily calculated.

Classification of Lenses

There are two convenient methods of classifying lenses, one being according to the function which they are designed to perform and the second being according to the nature of the lens material. The first classification of lenses depends on their geometrical properties, and for the purpose of this series of articles is not so fundamental as the second. Most of the lenses so far constructed have been designed to produce a pencil beam from either a point source such as a dipole or a line source such as a slotted waveguide. The simplest design equations for these cases will be given later. More complicated cases will be dealt with in a later article.

There are six convenient groups under which the refracting media used in the construction of radio lenses may be classified. These are:—

1. Solid dielectrics.
2. Metallic delay dielectrics.
3. Metal plate dielectrics.
4. Rodded dielectrics.
5. Path length dielectrics.
6. Miscellaneous dielectrics.

A brief description of each of these types will be given below and fuller details will follow in later articles.

Examples of some of these types of lenses are illustrated on this page.

Solid dielectric lenses are exactly analogous to optical lenses, the dielectric of the radio lens behaving in the same way as the glass of an optical lens. It may be shown from Maxwell's equations that for dielectrics the refractive index is related to the dielectric constant (K) by the equation

$$n = \sqrt{K} \quad \text{..........} \quad (3)$$

Suitable materials for use in making dielectric lenses are listed in Table 1 together with their dielectric constants, specific gravities and loss factors. The loss factors give a simple indication of the amount of energy absorbed in the passage of the radiation through the lens. It may be seen from the magnitudes of these loss factors that this source of loss is negligible.

The principal advantage of dielectric lenses is their simplicity. They may be readily made by machining a solid block of dielectric to the required profile. The chief disadvantage of such lenses is the weight and the consequent difficulty of supporting the structure. The principal use of such lenses is in the correction of sectoral horns, where only a small depth of the dielectric is required.

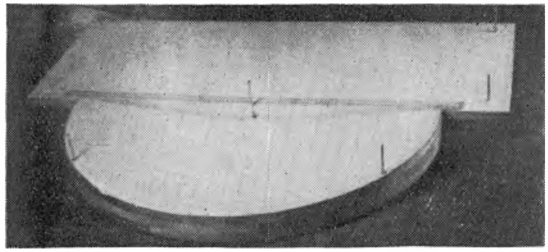
The properties of a normal dielectric arise from the polarisation of the molecules in the presence of an electrical field. Metallic delay dielectrics[4] are basically large-scale models of a dielectric structure, the molecules being simulated by metallic elements arranged in a lattice structure. The dimensions of the metallic elements are less than the wavelength at which the structure is to be used. This type of dielectric was developed by Kock and is known as a delay dielectric because it has a refractive index greater than unity. It has also been referred to as an artificial dielectric, but this term applies equally well to any of the groups listed above except the first.

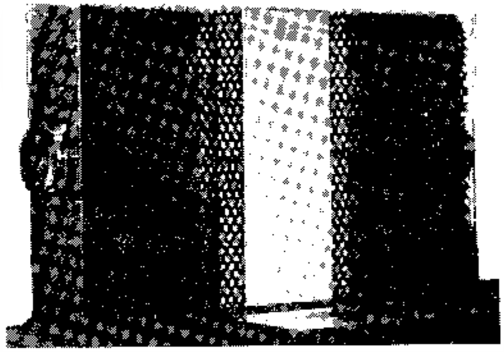
TABLE I

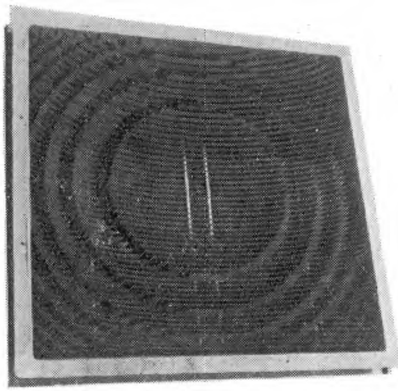
DIELECTRICS SUITABLE FOR LENS CONSTRUCTION*

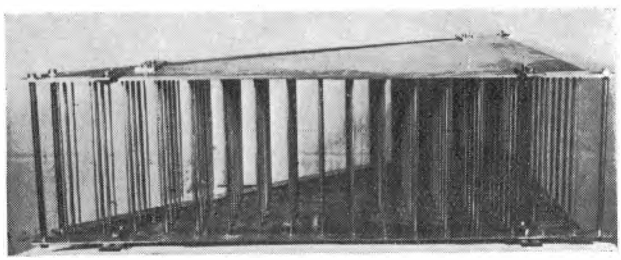
Dielectric	Refractive Index	Specific Gravity	Loss Factor Tan δ
Ethyl Cellulose	1.4	1.10-1.20	0.0135
Polyethlyene (e.g., Alkathene) ...	1.5	0.95	0.0003
Polystyrene (e.g., Distrene) ...	1.6	1.05-1.07	0.0002
Polyisobutylene	1.6	0.95	0.0005
Methyl Methacrylate (e.g., Perspex)	1.7	1.18	0.02
Mica	2.7	—	0.0001

* The data contained in this table is obtained from "Plastics in the Radio Industry" by Couzens and Wearmouth. ("Electronic Engineering" Monograph)









Some examples of microwave lenses

(Top) **Variable refractive index type**

(Above) **Delay lens**

(Left) **Metal plate lens**

(Below) **Rodded prism**

Metal plate lenses[5] depend for their operation on the fact that the phase velocity of a wave travelling between parallel plates is higher than in free space, provided that the electric vector is parallel to the plates. The mode of propagation between the plates is similar to that in a waveguide, and for this reason these lenses are often referred to as waveguide lenses. Since the phase velocity is higher than in free space it follows that the refractive index is less than unity (*c.f.* Equation (2)). Such a possibility does not arise in optics and is another of the ways in which radio and optical lenses differ.

Rodded dielectrics are similar to delay lenses in that they consist of a lattice structure, the elements of the lattice being metal rods. The structure is such, however, as to produce a refractive index less than unity, and it is dispersive, *i.e.*, the refractive index varies with frequency. This is a disadvantage compared with the delay lens whose refractive index is constant over a band of frequencies, but, on the other hand, rodded dielectrics do not require the presence of a supporting medium which, as will be seen later, is an undesirable feature of the delay dielectrics.

Path length lenses[6] employ a somewhat different principle to the types already described. It will be shown below that the function of a lens is to alter the phase fronts of radiated beams. In all the above types this alteration of phase front is achieved by making use of media in which the phase velocity is either greater or less than it is in free space. Path length lenses, however, change the phase fronts by causing different portions of the wave to travel different distances, the velocity of propagation within the lens remaining constant at the free space value.

The miscellaneous types of lens in the last group have one feature in common, viz., that the refractive index within the lens varies from point to point. That such a variation of refractive index will be advantageous in lens design has long been known to workers in optics, and has been used by workers in electron optics. No really practical method of achieving the variation in optics has yet been developed. Because of the much larger wavelength involved in the case of radio lenses, various methods of obtaining media with non-homogeneous refractive indices are possible. These methods are being developed and already some interesting results have been achieved.

The Case for Lenses

In all the aperture aerial systems used at microwave frequencies—stacks of slotted wave guides, dipoles feeding paraboloid mirrors, electro-magnetic horns, etc.—the gain of the system is determined by

(i) the area of the aperture,

(ii) the energy distribution over the aperture,

(iii) the phase front emerging from the aperture.

The greatest practicable gain and narrowest beam for a given aperture is achieved when the energy distribution is uniform and the phase front plane. (It has been shown by Woodward and Lawson[7] that super gain aerials, i.e., aerials with a gain of more than that of a constant phase and amplitude distribution, are a theoretical possibility but not likely to be realisable in practice. Such systems will not therefore be considered.) Gain is, however, not the sole criterion and it may be desirable to adjust the energy distribution over the aperture to reduce side-lobe intensity. Even so, the phase front should be as nearly plane as possible if the aperture size is to be exploited to the full.

The purpose of the microwave lenses described in this and the subsequent articles is to produce radiating apertures in which the phase fronts of the emerging rays are plane and thus to achieve a performance limited only by the size of the system. It is well known that the maximum gain which can be achieved with an aperture of area A square wavelengths is $4\pi A$. Practical systems with a reasonable degree of side-lobe suppression have a gain of just over half this value.

In the case of large aperture systems, the paraboloid mirror is the chief competitor of the lens as a focusing device. It has the great advantage over the lens of simplicity. However, the inherent disadvantage of the mirror is that the primary feed which may be a slot radiator, open-ended waveguide or dipole is in the path of the emergent radiation. This introduces two difficulties: firstly, obscuration by the feed distorts the wavefront of the radiation emerging from the aperture, with consequent increase in side-lobe levels: secondly, the emergent radiation intercepted by the primary radiator sets up a standing wave in the feeder, with consequent difficulties in matching to the oscillator.

If a lens is used as in Fig. 2 the first of these difficulties is solved immediately. The second is mitigated to a large extent, but is still present. Most practical lenses have a reflexion coefficient of about

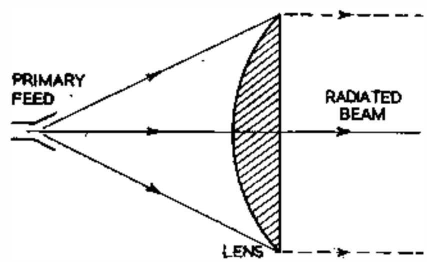


Fig. 2. Method of using a microwave lens

5 per cent which causes some energy to be thrown back into the feed. The avoidance of beam distortion by the primary feed is of particular importance in scanning systems, in which the supporting structure is frequently quite massive.

Lens Profiles

A great variety of lens profiles is possible and a discussion of the more useful will be given in a later article. One, however, is by far the most common and it will be considered here. It is the case of a lens designed to produce a narrow beam and which has the face remote from the feed (hereafter referred to as the front face) plane. The equation of the rear face will be derived for a lens made from a solid dielectric and the modifications required for the various other types of dielectric will be discussed later.

Consider, then, a spherical lens of focal length f constructed from a solid homogeneous dielectric of refractive index n and fed by a point source F. A cross-section of this lens through the axis is shown in Fig. 3.

The lens contour is most simply derived if the polar co-ordinates r, θ (θ is the angle OFA in Fig. 3). Let 2a be the aperture of the lens, and d be its maximum thickness.

As the front surface CBD of the lens is plane the beam must be formed entirely by refraction at the rear surface $CAOD$. Every ray path through the lens such as AB must therefore be parallel to the lens axis OG. In order to obtain constant phase

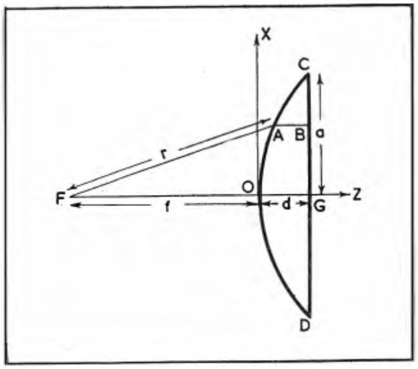


Fig. 3. Calculations of a lens profile

over the surface CBD the time taken by a wave to travel from F to this surface must be the same for all paths. The algebraic expression of this condition is

$$FA/v_0 + AB/v = FO/v_0 + OG/v \quad \ldots\ldots \quad (4)$$

where v_0 is the phase velocity in free space and v is the phase velocity within the lens.

From Equation (2), $n = v_0/v$.

When this is substituted in Equation (4) and the various lengths expressed in terms of r, θ, f, d, there results $r + n\ (d + f - r \cos \theta) = f + nd$ which simplifies to

$$r = \frac{(n-1)f}{n \cos \theta - 1} \quad \ldots\ldots \quad (5)$$

For a dielectric lens, n always exceeds unity and Equation (5) is therefore the polar equation of a hyperbola with focus at F and eccentricity n.

The refractive indices n of the dielectrics listed in Table I are practically independent of the frequencies at which the lens will be used, and it follows from Equation (5) that the lens contour will be correct for all these frequencies. In optical terminology it may be said that dielectric lenses are achromatic: the radio engineer would prefer to say that such lenses have wide-band properties. It will be seen later that the band-width of a dielectric lens is limited by other considerations.

Equation (5) does not involve either the semi-aperture a, or the maximum thickness d of the lens. It may, however, be readily shown that, since $FC = FO + nOG$, the quantities a and d are related by the equation

$$(n^2 - 1)\ d^2 + 2(n-1)fd - a^2 = 0 \quad \ldots\ldots \quad (6)$$

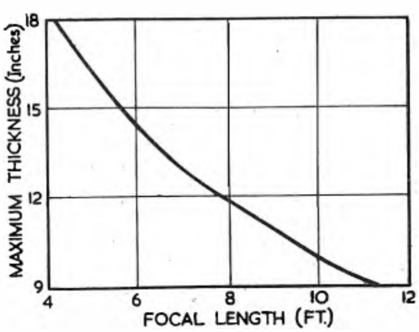
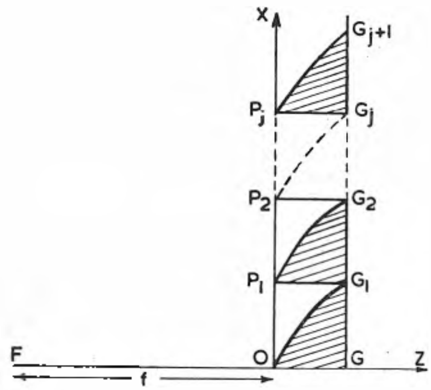


Fig. 4. Maximum thickness of a lens as a function of focal length. (Lens aperture 6ft. refractive index 1·5)

Substitution of numerical values in this equation shows that for constant a and n, the maximum thickness d increases rapidly as f is decreased: this is illustrated in Fig. 4 where d is plotted against f for a 6 ft. aperture and a refractive index of 1.5. The weight of the lens will therefore also increase rapidly as the focal length is reduced and so there will be a lower limit to the focal length which may be used. This limitation may be avoided at the expense of reduced band-width by a process known as "stepping." The principle of this stepping process is that the phase front of the front surface of the lens may contain any number of discontinuities of 360° without appreciably altering the radiation pattern.

A stepped lens consists of a number of zones as shown in Fig. 5, the phase of the waves leaving the zones being successfully increased by steps of 360°. The equation of the curve OG_1 is, of course, given by Equation (5) and the equations for the other zones may be readily derived. For the jth zone, i.e., the

Fig. 5. Profile of a stepped lens



section of the lens $P_{j-1}\,G_{j-1}\,G_j$, the optical path is less than that for the first zone (OGG_1) by $(j-1)\lambda$ where λ is the free space wavelength, and so it may be shown that the equation of the curve $P_{j-1}\,G_j$ is

$$r_j = \frac{(n-1)f - (j-1)\lambda}{n\cos\theta - 1} \quad \ldots\ldots \quad (7)$$

This is the equation of a hyperbola with the same focus F, and the same eccentricity n as the hyperbola for the first zone.

The outer radius OP_j of the jth zone is given by

$$OP_j^2 = FP_j^2 - FO^2$$
$$= (f + j\lambda)^2 - f^2 \quad \ldots\ldots \quad (8)$$

since $FP_j = f + j\lambda$

Simplifying Equation (8) gives

$$OP_j = \sqrt{j\lambda(2f + j\lambda)} \quad \ldots\ldots \quad (9)$$

The thickness of each of the zones of the stepped lens may be calculated from the equations of the contours. The algebra involved is straightforward but tedious. If the focal length is much larger than the wavelength as will usually be the case, the thickness of the zones is approximately $\lambda/(n-1)$.

The student of optics will recognise the similarity between stepped lenses and Fresnel zone plates. From Equation (8) it may be seen that each step of the lens contains two Fresnel zones. A stepped lens may, in fact, be regarded as a phase-corrected zone plate.

The equations of the contours for all the zones except the first involve the free space wavelength and so the design of a stepped lens is only correct for one particular frequency. In practice, however, the lens may be used over a band-width of a few per cent and this limitation on the band-width is an acceptable price to pay for the considerable reduction in the weight of a large aperture lens. A further disadvantake of stepping the lens is the so-called shadow effect[8]: it will be seen from Fig. 5 that the energy radiated by the primary feeds and incident on each of the sections $P_1 G_1$ of the lens will not be phase corrected by the lens and so will produce side-lobes far removed from the main beam. The level of these side-lobes, however, is not usually sufficiently high to cause any serious trouble. This shadow effect may be avoided by stepping the lens on the plane surface as in Fig. 6, where each step is set back $\lambda / n - 1$ from the previous one, but the lens then assumes a most inconvenient shape and one which is mechanically weak.

For most purposes it is more convenient to have the equations of the lens contours in Cartesian coordinates. If the system of axes with origin O shown in Figs. 3 and 5 is used then the Cartesian equations are for the unstepped lens.

$$x^2 \quad (n^2-1)z^2-2fz(n \quad 1) = 0 \qquad \ldots\ldots\ldots\ldots\ldots \quad (10)$$

and for the j[th] zone of the stepped lens

$$x^2 \quad (n^2-1)z^2 \quad 2fz(n \quad 1) + 2(j-1)n\lambda z$$
$$-2(j-1)f\lambda-(j-1)^2\lambda^2 = 0 \qquad \ldots\ldots\ldots\ldots\ldots \quad (11)$$

These equations give the contours of any cross section of the lens through the axis—the complete spherical lens is the solid of revolution obtained by rotating Figs. 3 and 5 about their z-axes. The shape of a cylindrical lens may also be obtained from these equations—it is the cylinder obtained by moving the shapes of Figs. 3 and 5 parallel to the y-axis, i.e., the axis at right angles to the plane of the diagram.

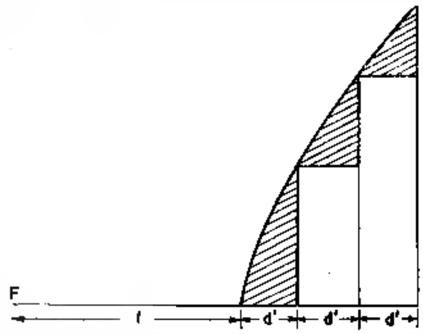


Fig. 6 Alternative method of stepping a lens

The above derivation of lens profiles is not entirely rigorous, since it has been assumed that the ray paths through the lens will be parallel to the axis. This may be proved by an application of Fermat's least path principle. An alternative method of deriving the lens contours is to apply Snell's law at each point of the lens surface—this results in a differential equation for the surface, the mathematics involved being rather more complex than in the above derivation.

(To be continued)

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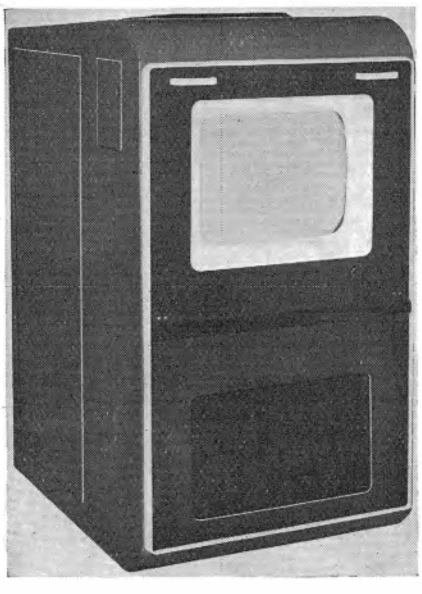
"Cintel" 20-Inch Television Receivers



TWO monitor receivers, manufactured by Cinema-Television, Ltd., have recently been delivered to the B.B.C. Ordered to help their drive for better quality television, these receivers will provide B.B.C. monitoring engineers with as faithful a reproduction as it is reasonably possible to obtain from the received signals. The engineers will know that any defect in the picture should be the signal transmitted and not in the receiver.

The exacting specifications necessitated the production of a 20-in. cathode ray tube (the largest glass tube being made in this country) giving the biggest direct viewing picture, namely, 16 in. by 12 in. The equipment employs about three times the number of valves, for the same function as are used in a normal home television receiver. It is designed to reproduce a television picture and the associated sound programme from a combined "picture plus synch." signal and an audio signal supplied to it by cable. Each apparatus is otherwise self-contained, being operated from A.C. mains.

Provision is made for accepting signals in the ranges 1.0 volt overall (picture plus synch.) ±6 db. and 10 volts overall (picture plus synch.) ±6 db., and for terminating the signal line or leaving it unterminated, as desired.

The signal amplifier has a level frequency response up to 3 Mc/s. and is phase-corrected. The scan generator is designed to give deflexions which, in addition to all normal errors, are corrected for those introduced by the geometry of the tube; as well as assuring perfect interlacing, even in the presence of interference in the applied signal. Controls are few and comparatively simple to operate.

Permanent Magnet Lenses for Television Tubes

By

D. HADFIELD, B.Sc., A.M.I.E.E., A.I.M.*

THE focusing of cathode-ray tubes for television equipment is receiving increasing attention due to the phenomenal rate of progress in the development of commercial television receivers during the last few years. Designers are quickly realising the economic advantage of using permanent magnets in electron lenses for focusing the cathode-ray beam to ensure low-priced A.C./D.C. receivers, where H.T. voltage can ill be spared to supply electromagnetic focusing arrangements.

From the practical point of view the permanent magnet lens is very stable in itself, but the focusing of the cathode-ray tube is dependent upon other factors, such as the H.T. supply, which will, unless stabilised, vary with a fluctuating mains voltage. Ideally, the focus should be under control by the viewer in order to correct for such variations. In the case of permanent magnets this is not so easy unless the field strength, and therefore the focal length, can be adjusted manually by means of a control knob, as in the purely electromagnetic system. This disadvantage has been overcome in some designs by combining the two systems and using an electro-permanent magnet lens, which probably constitutes the ideal focusing arrangement.

From the economic aspect there is little to choose between the cost of a permanent magnet focusing unit, and an electro-magnet focusing assembly, including the control system. Permanent magnet lenses are coming into favour chiefly because of the present trend in receiver design, which is towards cheaper and hence transformer-less television sets.

Requirements of Electron Lenses

The physical principle involved in the focusing of a cathode-ray beam is to cause divergent electrons to converge during their journey from the cathode to the screen where their presence should result in a fluorescent spot of the smallest possible diameter. This desired effect may be realised in several way

If the tube is "soft" and contains a small amount of inert gas, then focusing is automatically produced by ionisation of the gas particles by the electron stream, providing the anode voltage is not too high and thus the velocity of the stream is maintained reasonably low. However, this early type of tube is not used today in commercial television receivers.

Another method of focusing is electrostatic, but applies only to "hard" tubes. This is effected by applying certain voltages to auxiliary anodes mounted coaxially inside the tube. Again, electrostatic focusing in cathode-ray tubes is not generally found in modern television receivers.

The third type of focusing is produced by the action of a magnetic field on the electron stream. "Magnetic focusing" is almost universally used for cathode-ray tubes in commercial television receivers. The focusing effect is produced as follows: Fig. 1 shows diagrammatically a stream of electrified particles (electrons) carrying a charge e and moving with a velocity v along the axis of a cathode-ray tube. As the stream is virtually continuous, a current flows along it in the direction opposite to its motion. It can be shown that the force exerted on each particle as it travels through the magnetic field H is equal to $Hev \sin \Theta$, where Θ is the angle between the direction of the field and that of the cathode current. This force acts perpendicularly to the plane containing the path of the particle and the direction of the field. Such a force is brought to bear on any particle which, after leaving the cathode, follows a divergent path away from the axis of the tube, and is produced in such a direction as to cause any such particle to be bent back towards the axis. Apart from the influence of this force on the particle while it is travelling through the effective region of the magnetic field, there is also the accelerating force due to the high anode potential above the cathode, which acts in a mutually perpendicular direction, that is, along the axis of the tube, thus causing the particle to travel in a helical path of ever-decreasing radius, so that such particles will all converge at some point on the axis of the tube at a certain distance from the region of the magnetic field.

Suppose, for simplicity, that the particle is projected at right angles to the magnetic field (the force in this case is equal to Hev). Since the force is always at right angles to the path of the particle the speed will remain unaltered. Let r be the radius of curvature of the path of the particle under the action of the field, and m its mass. The centrifugal force is equal to mv^2/r and for equilibrium this must be balanced by the force due to the field. Hence:

$$mv^2/r = Hev,$$

$$\therefore r = mv/eH.$$

As v is constant, the radius of curvature is constant, and the particle describes a circle of radius mv/eH in a plane at right angles to the magnetic field.

If the velocity of the particle has a component parallel to the magnetic field this component will remain unaltered since there is no force acting on the particle along this direction, and the particle thus moves forward through equal distances in equal

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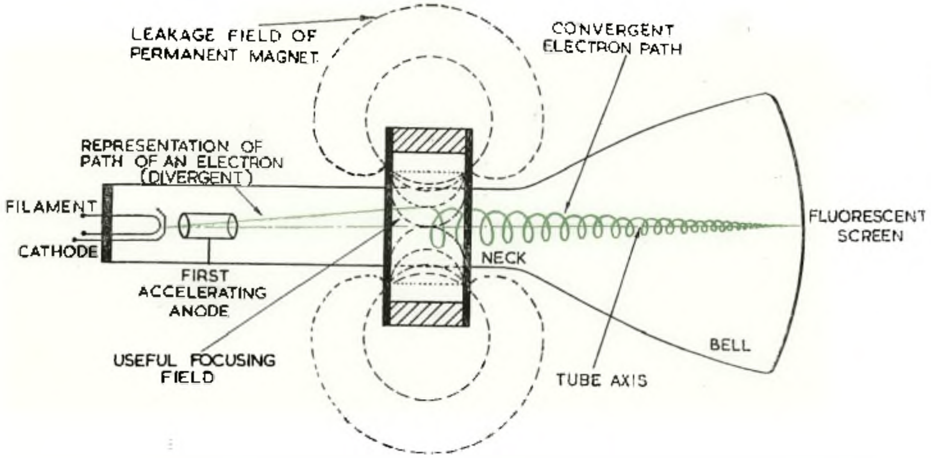


Fig. 1. Diagrammatic representation of magnetic focusing of cathode-ray tube

times along this line, while at the same time describing a curved path around it. Its path is thus a helix, the axis of which is parallel to the field. If Θ is the angle between the direction of projection of the particle and the field, then:

$$mv^2/r = Hev \sin\Theta \text{ ...... (Laplace)}$$

$$\therefore r = \frac{mv}{eH \sin\Theta}$$

Since v and Θ are constant, r is constant and the path of the particle is a helix wound on a circular cylinder, with its axis parallel to the magnetic field, and of radius $r \sin^2\Theta = mv \sin\Theta / eH$. When $\Theta = \pi/2$, then the helix contracts into a circle of radius $r = mv/eH$.

When $\Theta = 0$, then the particle is projected parallel to the field, and continues in a straight line, i.e., $r = 0$.

If the field is non-uniform, the particles describe helices of varying amplitudes about the lines of force of the magnetic field.

The flux distribution over the effective volume of the magnetic field is very important in order to obtain optimum focusing devoid of astigmatism. The standardised cathode-ray tube frame size for commercial television in this country is 405 lines making up the whole picture, and with a 9 in. diameter tube, having a picture size of about 7½ in. by 6 in., this means that they are 69 lines per inch. Thus a spot diameter of about .01 in. is required, allowing for non-overlapping or touching of the lines in the complete raster. As regards the position of a focusing magnet, it must be such that the magnetic field does not interfere with the deflexion system or the electron gun. The field transverse to the axis should be a function only of the distance from the axis, that is, the lines of constant field strength should be circles concentric with the circumference of the tube neck, otherwise astigmatism will result. If the transverse field is slightly displaced from its symmetrical position there will be a permanent deflexion of the spot from the centre of the screen. This is known as excessive shift. It is also advisable for optimum performance that the field should be constant over a small area (about 1 cm. in diameter), in the centre of the magnet, occupied by the electron stream.

The axial distribution of the magnetic flux is also a factor which must be considered. For optimum focusing the magnetic field distribution along the axis of the tube, with which the magnet should be coaxial, is given by the formula:—

$$H_z = H_0 / (1 + (z/a)^2)$$

where z is the distance from the centre measured along the axis,

H_z is the field strength at distance z,

H_0 is the field strength at $z = 0$,

a is the distance along the axis at which the field strength is $H_0/2$.

Fig. 2 shows the ideal distribution plotted from the above equation.

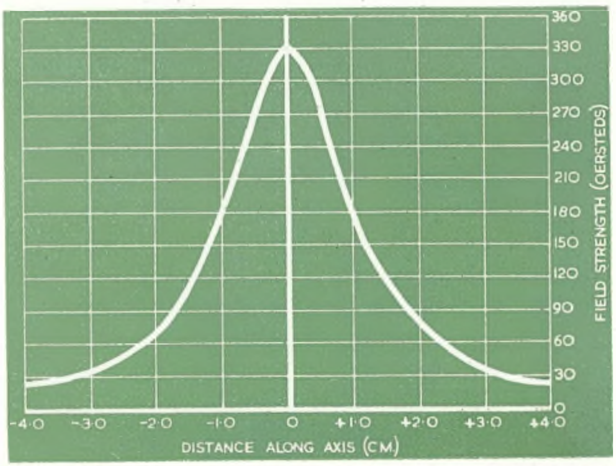


Fig. 2. Ideal distribution of axial flux

The order of field strength at the magnetic centre of the magnet, which should lie on the axis of the tube, is, of course, dependent on the value of the anode voltage. High voltage tubes require stronger magnetic focusing fields, but generally the field strength lies between 200/400 oersted. It is necessary during production and installation of a television receiver to be able to adjust the focusing slightly, and in order to do this it must be possible to vary the magnetic field by a small amount about a certain value. This is easily done in permanent magnet focusing by changing the reluctance of the magnetic circuit by some form of variable magnetic shunt. It is also essential that the magnetic field be absolutely constant in service and this is ensured in the first instance by fully saturating the magnet during magnetisation and then stabilising by the application of a small alternating current field, which may reduce the field strength of the magnet by a few per cent, or, alternatively, by allowing a certain period of time for natural ageing to occur after carrying out a few cycles of magnetic shunting.

Development of Focusing Magnets

Early forms of permanent magnets for focusing were simple cast rings of magnet alloy, usually Alnico, ground on the end faces only, and having soft iron annular end plates whose internal diameter was somewhat less than the internal diameter of the magnet ring as shown in Fig. 3. In effect, the internal peripheries of the iron pole pieces formed two circular poles so that the flux path was of the form shown in Fig. 4.

Another type was in the form of a " doughnut," again in Alnico, having inwardly-directed poles of the same material, and made in the form of a single casting. In order to render the flux distribution more uniform, soft iron "liners" were fitted inside the two circular poles in intimate contact and coaxial with the cathode-ray tube, as shown in Fig. 5.

A somewhat unusual experimental design was as shown in Fig. 6. It is a simple cast annular disc of Alnico alloy magnetised radially, that is, with one pole forming the inner periphery and the other pole forming the outer periphery. The form of the flux path in this magnet is shown in Fig. 7.

The axial distribution of field strength and the equi-field strength contours for the magnet in Fig. 5 and axial distribution only for the magnet in Fig. 6 are shown in Figs. 8, 9 and 10. It is interesting to note that in the case of the type shown in Fig. 6 there is a neutral point in the geometric centre of the magnet and the peak value of field strength occurs at two points ± 1.25 cm. on either side of the geometric centre of the air gap. In other words, this particular magnet acts as a double electron lens. The other types have the same field strength distribution as shown in Fig. 8, which corresponds to the ideal curve shown in Fig. 2. The approximation of the equi-field strength contours to concentric circles is clearly seen in Fig. 9.

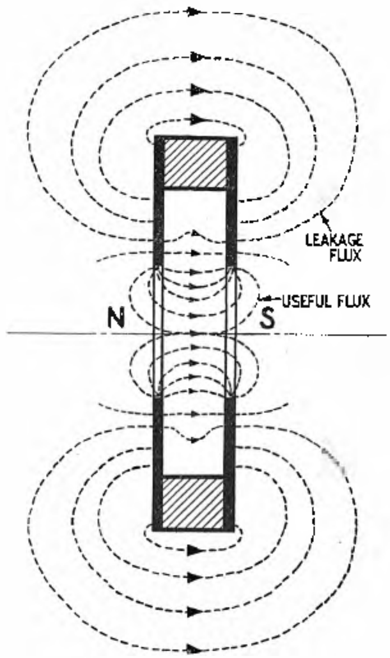
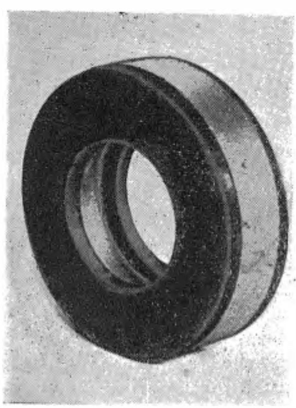


Fig. 4. Flux paths of magnet shown in Fig. 3

For the purposes of investigating the field characteristics, a circular search coil of small dimensions was used in conjunction with a brass jig as shown in Fig. 11. By this means the search coil could be located accurately at various points in the region of the central air gap of the magnet. By rotating the upper part of the jig, which carries the magnet, in the groove in the graduated base into different positions, the variation in field strength around the circumference of a circle concentric with the magnet periphery can be determined. Also, by altering the position of the circular search coil on the vertical brass carrier rod, the field strength distribution along the axis of the magnet can be explored.

To prove the electron optical qualities of the magnet it is necessary to observe the results on the cathode-ray tube for which the magnet is designed. Firstly, the magnet is used with a stationary beam to obtain the best possible focusing of the spot. The size of the spot can then be determined by having a complete scan of 405 lines on the tube and closing the raster until the lines are just touching. By measuring the overall width of the raster and dividing by 405, the spot diameter can be obtained. Astigmatism can be detected by rotating the magnet coaxially with the tube, when its presence

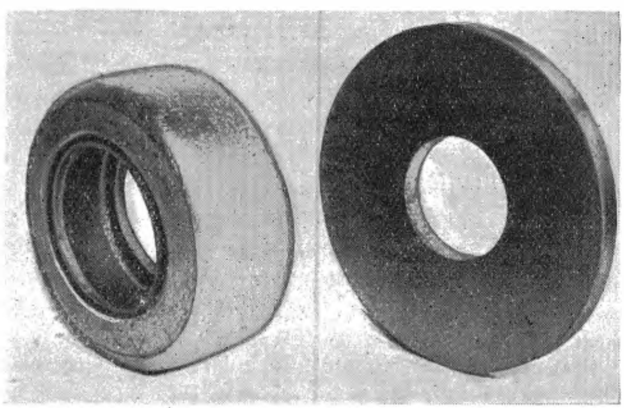


New types of focusing rings:

Fig. 3 (left). Ring type assembly

Fig. 5 (centre). " Doughnut " type

Fig. 6 (right). Plain disk type



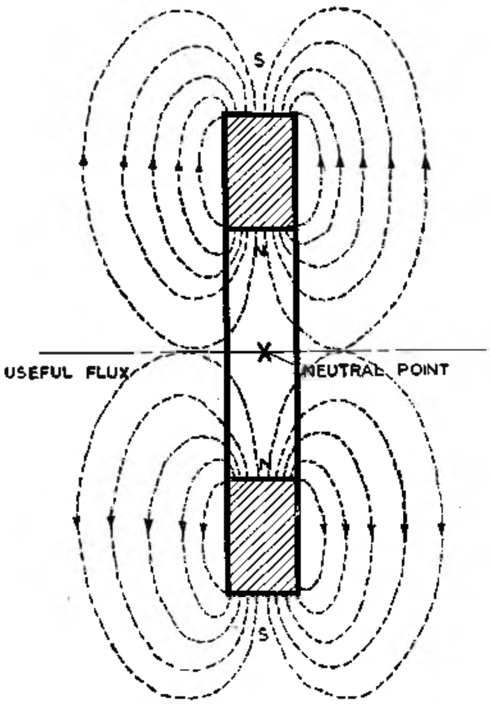


Fig. 7. Flux paths of magnet shown in Fig. 6

will be indicated by the spot becoming oval in shape instead of circular. Excessive shift can be detected by similar means, when the spot will move in a circular path indicating that the beam does not lie along the tube axis.

The early designs of focusing systems suffered from the disadvantage that the field strength was fixed and could only be altered in steps in the downward direction by ageing the magnet in an alternating field. The focus of the spot could be adjusted by moving the magnet along the axis of the tube, but this is not good practice as in a well-designed compact receiver there is very little space for such adjustments to be made. Post-war designs have recently been provided with a magnetic shunt to vary the field strength about a certain value by slightly changing the reluctance of the magnetic circuit.

Modern types comprise a plain ring of magnetic alloy with stamped pole pieces of soft iron or mild steel abutting the ground faces of the ring, but having an additional movable portion in order to alter the length of the air gap or shunt the leakage flux. A close-fitting iron sheet sleeve moves within one of the pole pieces and is constrained by a spiral slot and pin giving an overall change in internal gap length of about ½ in. (Fig. 12). Another type, having a constant gap, has an external form of shunt. The shunting effect is a maximum, and therefore the field strength is at its lowest value in one extreme setting, while the reluctance is at a maximum and therefore the highest field strength is obtained in the other extreme setting. By these means a variation can be obtained in field strength which means that focusing can be achieved without moving the magnet along the tube.

Other means of producing field variation, and control, have been contemplated and examined. One system is the electro-permanent magnet which has been successfully applied to commercial receivers, particularly in the U.S.A. and to a limited extent in Great Britain. In it a small magnetising coil is placed inside the magnet to boost or reduce by a small amount the main field due to the permanent magnet. It is not ideal to use the auxiliary coil to reduce the field, as this causes the working point of the magnet to move down the demagnetisation curve, and thus tends to induce permanent demagnetisation, unless it is operating about a point on a recoil loop or subsidiary hysteresis loop. Preferably, the direction of the field due to the coil should be such as to reinforce the permanent magnet field.

An experimental coil consisted of 50 turns of 32 gauge wire, and when a current of 1 amp. was passed through it a change of ±5 per cent was obtained in the field strength at the centre of the gap. It is also possible to calculate the effect of the superimposed field on the gap field strength. The M.M.F. produced by the coil can be obtained from the formula:—

$$\text{M.M.F.} = 0.4\pi NI,$$

where N is the number of turns and I the current (amp.). If the magnet is working at a point such that the working value of H is H_w, the M.M.F. due to the magnet is H_wL, where L is the mean length of magnet material.

Thus the total M.M.F. $= H_wL \pm 0.4\pi NI$, in accordance with the direction of the current. The new field strength

$$H_g = H_o\left(\frac{H_wL \pm 0.4\pi NI}{H_wL}\right)$$

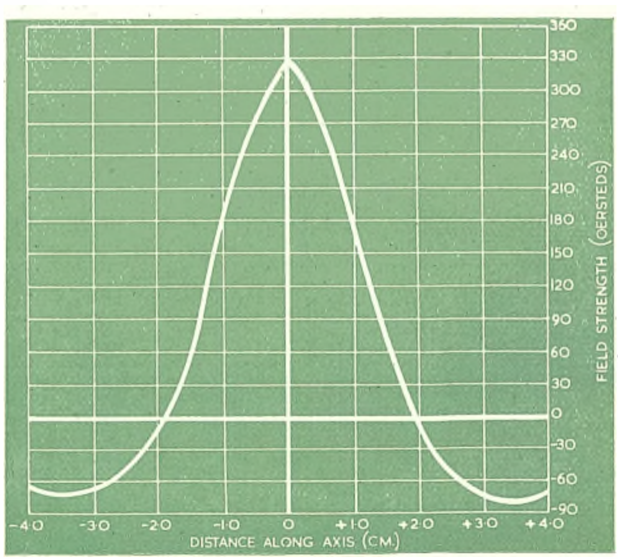
$$= H_o \pm \frac{0.4\pi NI H_o}{H_wL}$$

thus the change in field strength is

$$(H_g - H_o) = \pm \frac{0.4\pi NI}{H_wL} \times H_o$$

Attempts to vary the field strength by using toothed pole pieces inside the magnet and varying the gap

Fig. 8. Axial distribution of field strength of magnet shown in Fig. 5



by rotating one pole piece with respect to the other have met with little success.

Permanent Magnet Design and Materials

Permanent magnet design, at best, is a combination of the application of fundamental principles associated with experience of leakage factors, and a detailed knowledge of permanent magnet materials is essential. It is usually difficult to derive formulae which can be generally applied. Certain assumptions are involved in fixing volume and length, and hence the cross-sectional area of a permanent magnet component part.

The main difficulty lies in the estimation of the magnetic leakage flux, as distinct from the useful flux which the magnet is required to give. In the case of a magnetic lens this may be as high as 95 per cent. This leakage flux is estimated within reasonable limits from existing experimentally determined data. The basic principles involved are that the cross-sectional area of the magnet must be sufficient to supply the desired flux in the air gap, and that the length of the magnet must be sufficiently great for the magnetomotive force to be high enough to overcome the reluctance of the complete magnetic circuit.

A decision must first be taken as to the material in which the magnet is to be made, the governing factors being the type of magnet and the manner in which it has to be used. In the case of focusing magnets, there are certain restrictions due to the diameter of the cathode-ray tube neck over which the complete unit must pass, and thus, as the useful field is required over the very small central volume only, such a magnet is rather large and has a long working air-gap. The most suitable permanent magnet materials to use in this instance are the nickel-aluminium-cobalt-copper-iron types such as the isotropic Alnico or the anisotropic Alcomax alloys. Owing to the brittle nature of the newer alloys, it is necessary to ensure that there are no sharp changes of section in the magnet, and that the thickness at all points is adequate. The most usual shapes of alloy castings are simple rings, and recent developments in foundry technique have made possible the production of thin-walled rings in Alcomax, about $\frac{1}{8}$ in. thick, and 2 in. diameter, which are specially suitable for focusing rings.

Having selected a magnet material, the next step is to fix the working point, i.e., the position on the demagnetisation curve at which the magnet shall operate. Considerations of using minimum magnet alloy require that the working point should coincide with the (BH) max. point (see Fig. 13), and, when possible, this is done. However, due to geometrical and space limitations, this is usually not possible when designing focusing magnets. The working angle ψ is always rather large (about $\tan^{-1} 0.23 = 13°$). Now, $\tan \psi = (H_w / B_w)$, and $\tan \psi = (\text{area}/\text{length})$, thus the length to cross-sectional area ratio (or dimension ratio, as it is called) is low. The working point is now fixed and the corresponding values of B_w and H_w are read off the demagnetisation curve.

Certain formulæ may be deduced, from which the dimensions of the magnet may be obtained. The length of magnetic material required to drive the flux across the air gap is derived from considerations of the magneto-motive force, which may be defined as the product of the length of the magnet and the value of the demagnetising force H_w at the selected working point. Also, in a fixed short air gap, the magneto-motive force required to drive the flux across the air gap is the product of the field strength and the air gap length. Thus:

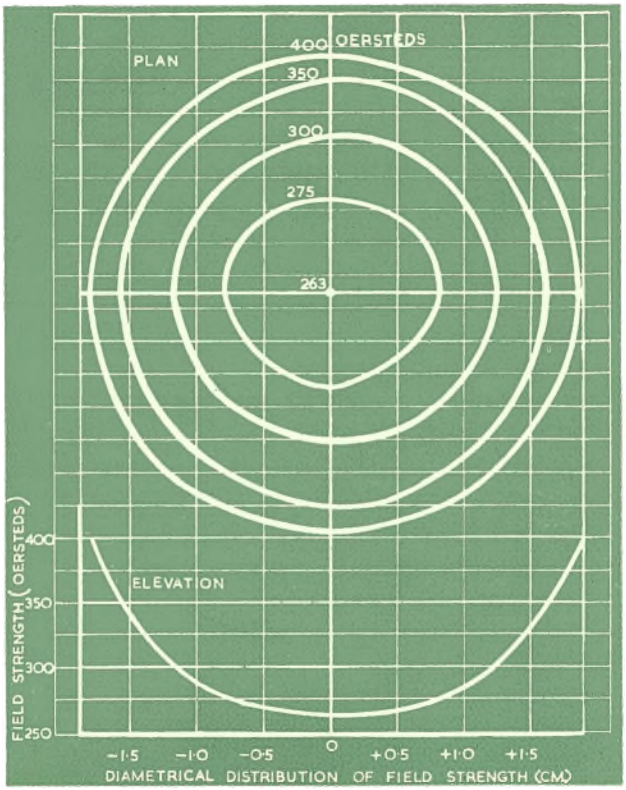


Fig. 9. Equi-field strength contours of magnet shown in Fig. 5

$$L_m H_w = l_g H_g k$$

$$\therefore \quad L_m = \frac{l_g H_g k}{H_w}$$

where L_m is mean length of magnetic material in cm.

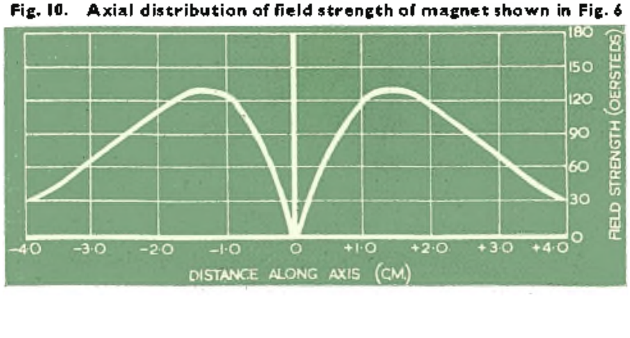
l_g is the mean length of the air gap in cm.

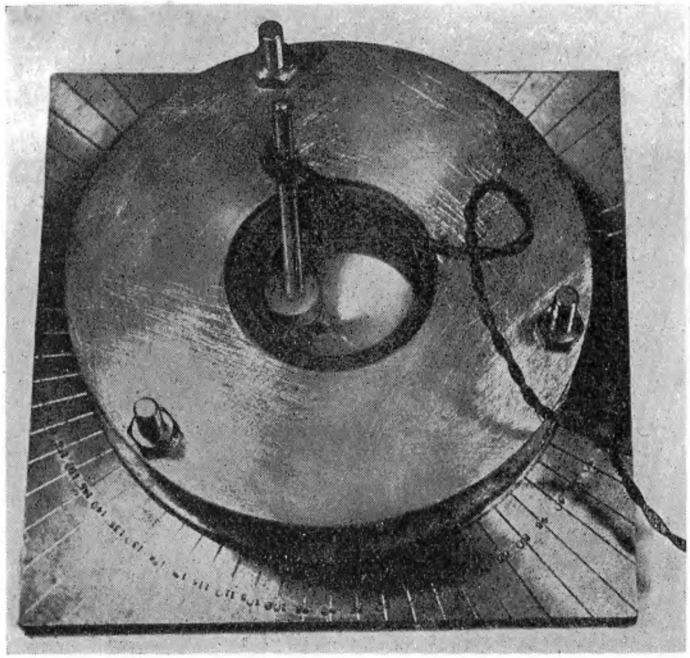
H_g is the gap field strength in oersted.

H_w is the value of H at the working point in oersted.

k is the reluctance factor for the circuit (usually about 1.2).

Fig. 10. Axial distribution of field strength of magnet shown in Fig. 6





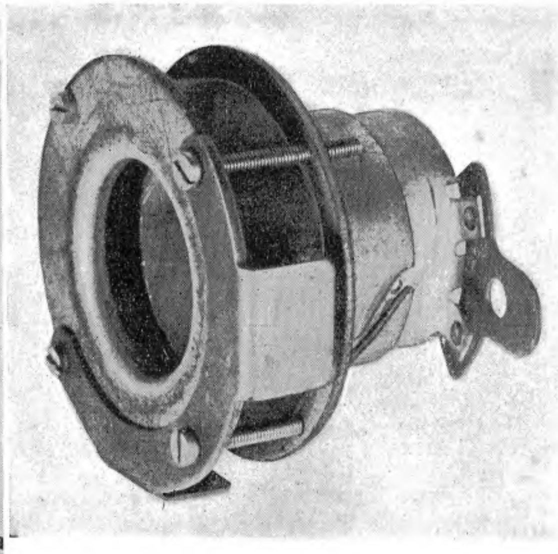


Fig. 11 (*left*). Field strength exploration jig

Fig. 12 (*above*). Focusing unit

Having thus determined the mean length of magnetic material, the next step is to calculate the cross-sectional area. If there were no leakage, the total gap flux would be equal to the total flux in the magnet, when the following relationship would hold:

$$H_g a_g = B_w A_m$$

$$\therefore \quad A_m = \frac{H_g a_g}{B_w}$$

where A_m is the cross-sectional area of the magnet in sq. cm.

a_g is the mean cross-sectional area of the air gap in sq. cm.

H_g is the gap field strength in oersted.

B_w is the value, at the working point, of B in gauss.

The effect of leakage is predominant, however, and the cross-sectional area of the magnet must be increased several times to compensate for it. In order to do so, the leakage coefficient K, i.e., the ratio of total flux (leakage-flux plus flux in the air-gap) to the useful flux (flux in the air-gap only) must be known. This factor is about 20 for focusing magnets, but a more accurate estimate can be obtained by comparison of the total permeance of the leakage paths and gap to the permeance of the air-gap.

The formula for determining the cross-sectional area now becomes:

$$A_m = \frac{H_g a_g K}{B_w}$$

The major difficulty in calculating the leakage is that the leakage paths are not of uniform section; thus it is necessary to introduce empirical relationships involving assumptions which give approximations sufficiently close for most practical purposes.

The total volume of magnet material V_m is given by:

$$V_m = L_m A_m = \frac{l_g H_g}{H_w} k \cdot \frac{a_g H_g}{B_w} K = \frac{H_g^2 v_g K k}{B_w H_w}$$

where v_g is gap volume in cc.

In many cases the higher energy anisotropic permanent magnet materials cannot be employed to full advantage, and this can be better explained by reference to Fig. 13 which shows demagnetisation curves of high-coercivity isotropic British Alnico III and also the anisotropic Alcomax II and the latest Alcomax III and IV alloys developed by the Permanent Magnet Association. On this diagram is also marked the permeance or working line corresponding to angle ψ for a magnet in which both similar Alnico III and Alcomax II magnet alloy rings give approximately the same gap field strength. This is largely the result of the low working point, but may be, in part, due to the saturation of the rather thin end plates. It will also be noted that substituting Alcomax III or IV would not improve the field strength by much more than about 20 per cent.

In order to achieve the high degree of uniformity required in the field strength distribution, it is important that the complete assembly is magnetised correctly to saturation and, further, that the permanent magnet ring is uniform throughout. This condition of uniformity is fulfilled with dimensionally accurate isotropic magnets of the Alnico type, but with anisotropic alloys extreme care has to be taken in production heat treatment whereby the permanent magnet ring is air-cooled uniformly from a high temperature while under the influence of a uniform strong magnetic field. If the cooling rate differs from one part of the ring to another, or the field in

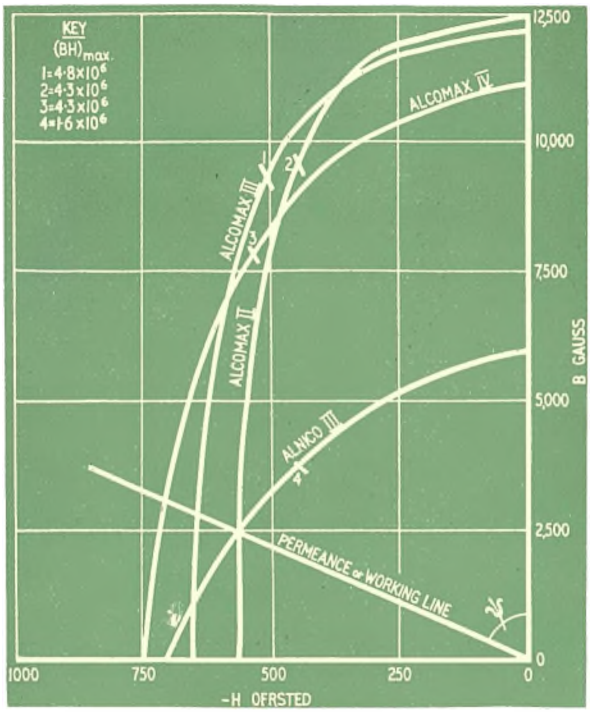


Fig. 13. Demagnetisation curves for Alcomax and Alnico alloys

which the magnet is immersed during cooling is non-uniform, then the final magnetic properties of the ring will vary from point to point, which will produce both astigmatism and excessive shift. The focusing of an electron stream is a most sensitive final test as regards the uniformity of the magnetic properties in a magnet.

In the case of magnets having slight non-uniformity, this can sometimes be compensated for by mounting the magnet on the tube either slightly eccentrically or non-coaxially or by providing a very weak corrector field by the addition of a thin strip of magnetised material placed in a suitable position.

Magnetising

To achieve both stability and a high degree of uniformity of field strength from any permanent magnet assembly it is necessary that the final magnetisation should be carefully carried out. First of all, the magnetising jig should have a coil which will give the necessary magnetomotive force to saturate completely the alloy of which the permanent magnet has been made. The magnetising forces for one or two materials are given in Table 1, and refer to completely closed magnetic circuits, which are possible during magnetisation.

Fig. 14. Magnetising jig

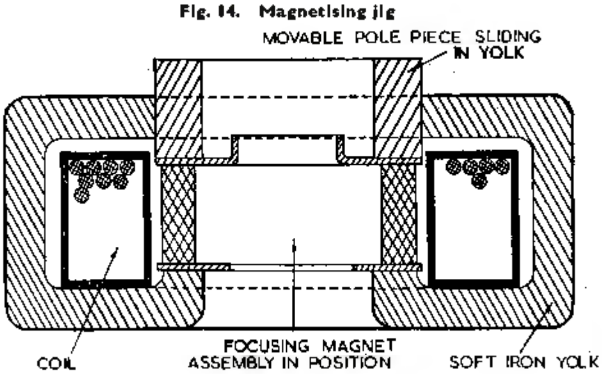


TABLE I

Permanent Magnet Alloy	Magnetising Force	
	Oersted	Ampere Turns per cm.
Alnico	2,000	1,600
Alcomax II	1,900	1,500
Alcomax III	2,000	1,600
Alcomax IV	2,400	1,900

The most efficient type of magnetising jig for most focusing magnet assemblies is as shown in Fig. 14. This comprises a coil which will develop the necessary magnetising force, in accordance with Table 1 above, into which the magnet assembly is placed. The magnetic circuit consists of a soft iron yoke which may be divided into sections in order that annular pole pieces may abutt on to the end faces of the magnet assembly, and the magnetic circuit completed by a ring of soft iron shrouding the coil. The mechanical arrangements of such a jig may be designed for quick action, where magnetisation is carried out on a television receiver production line. It is advisable to give the magnet several surges during magnetisation from the demagnetised condition to ensure that the material is fully saturated.

An indication of insufficient magnetisation is likely to be given by an excessive decrease in gap field strength soon after magnetising. However, such a condition is unsatisfactory, and with modern magnet materials the "natural" ageing should be of the order of only 1 or 2 per cent and should also occur within the first 24 hours or so after magnetisation has taken place, the magnet thereafter remaining sensibly stable to a few parts in a thousand for the rest of its life, providing that it is not mishandled in any way or demagnetised by any ferromagnetic bodies such as steel tools.

If absolute stability is required immediately after magnetisation then it is advisable to age artificially the magnet by a few per cent of gap field strength by immersing the assembly in an alternating field of suitable strength and frequency (say at 50 c/s.) for a short period, up to 15 seconds. Alternatively, artificial ageing may be carried out by "recoiling" the magnet along a subsidiary hysteresis loop. This is effected by operating the variable magnetic shunt a few times through the range of its extreme adjustments. If these precautions are carried out, then the magnet should remain stable and uniform.

In conclusion, it may be remarked that these new permanent magnets may possibly be applied to the focusing of television camera iconoscopes, and as magnetic lenses in the electron microscope. Electronic engineers, however, are showing a great interest in these new applications and their fields of use are expected to extend.

Acknowledgments

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Balanced Rectifier Modulators Without Transformers

Some simple rectifier modulators are described which give balance against carrier leak but avoid the use of transformers

By D. G. TUCKER, D.Sc., A.M.I.E.E.*

BALANCED rectifier modulators conventionally include transformers, since it is not possible to use bridge circuits with all three external circuits (i.e., signal input, carrier or "switching" input, and output) earthed, and some sort of bridge connexion of the rectifiers is essential. In many applications the use of transformers may be no disadvantage; and if it is desired to eliminate them, valve modulators can be used. But in applications where the input signal and switching signal are of the same frequency, so that the difference component is of zero frequency, valve modulators involve the difficulty of separating this component from the steady anode current. Such applications, where the modulator is generally known as a "phase-conscious rectifier," occur in many kinds of electronic instrument where the zero-frequency output is used for automatic control purposes, and it is always important that phase relationships at the modulator are accurate. Thus the use of modulators with transformers also gives difficulty, since the phase-shift of the transformers will require special compensation, and it is almost impracticable to produce a large number of transformers with the same phase-shift within close limits; thus each equipment requires individual phase-adjustment, which is a far from simple matter. Transformers also often lead to trouble due to winding-to-earth capacitance and reactance unbalances.

These difficulties are overcome by the use of a transformerless balanced rectifier modulator of one of the types described below. An approximate analysis of their performance is given on the assumptions that all rectifiers switch suddenly from a constant high resistance to a constant low resistance when the carrier or switching voltage is reversed, and that the input signal is small enough to have no effect on the switching of the rectifiers. In all cases the modulators are of the "half-wave" type, i.e., like the Cowan and not the ring modulator, and a valve is used to feed the carrier voltage to the rectifiers.

Shunt Modulator

This is shown in Fig. 1. Here the valve feeds the carrier voltage across the bridge network, which is connected in shunt with the signal circuit. The equivalent circuit of the latter is thus as shown in Fig. 2. The resistance $1/g_m$ appears across R_V and R_M, in series with one rectifier, because the cathode circuit of the valve offers approximately this internal resistance. The internal resistance from anode to earth is usually quite high compared with R_S and so can be ignored. The modulator thus behaves as a single "effective rectifier" with a forward resistance

$$r_f' = \frac{(r_f + R_S)[r_f(1 + g_m R_S) + R]}{2r_f(1 + g_m R_S) + R_S(2 + g_m R_S)} \quad \ldots\ldots \text{(1a)}$$

and a backward resistance.

$$r_b' \approx \tfrac{1}{2} r_b \quad \ldots\ldots \text{(1b)}$$

where r_f and r_b are the forward and backward resistances of the actual rectifiers, and it is assumed that $r_b >> R_S$.

The modulator is then exactly equivalent to a Cowan modulator, and the condition for maximum efficiency[1] (where efficiency is used in the sense of the reciprocal of the insertion loss of the modulator†) is thus,

$$R_o = R_S \ (= R, \text{ say}) \quad \ldots\ldots \text{(2a)}$$

$$\text{and } R = \sqrt{r_f' r_b'} \quad \ldots\ldots \text{(2b)}$$

and if in any condition we put $R = m\sqrt{r_f' r_b'}$ then the efficiency is

$$\eta = \frac{n^2 - 1}{(1 + mn)(1 + n/m)} \quad \ldots\ldots \text{(3)}$$

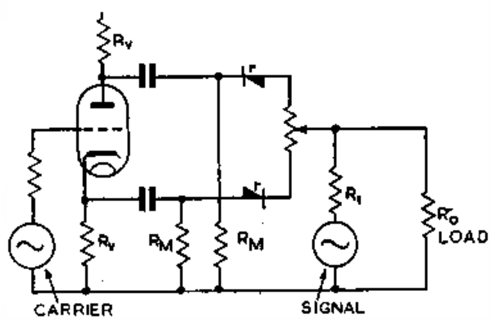
where $n^2 = r_b'/r_f'$

Graphs of the variation of this efficiency with variation of n and m are shown in Fig. 3.

It is evident that the mutual conductance of the valve feeding the carrier enters into the calculation

† This actually requires further definition. More precisely, the efficiency as used here is given by $\eta = (1/L)f \sim (1/L)b$, where L is the insertion loss of the network on one half-cycle of carrier, and f and b refer to the forward and backward direction of switching; η is therefore the ratio of the actual output to that which would be obtained if the effective rectifier had zero forward and infinite backward resistance. In the latter case, the *conversion loss* is 9.9 db in the optimum condition.

Fig. 1. Shunt modulator



* Post Office Research Station.

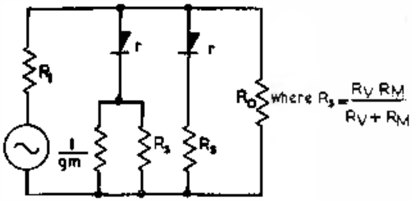


Fig. 2. Equivalent signal circuit of shunt modulator

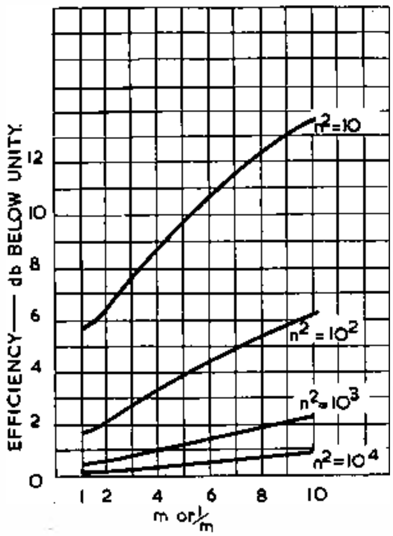


Fig. 3. Variation of efficiency of shunt modulator with rectifier ratio and circuit impedance

of efficiency, and it is worth while, therefore, to notice that its effect is small, and that changes in the valve due to ageing are likely to be imperceptible in the efficiency. For instance, take the numerical example below:—

$$r_f = 400\Omega,\ r_b = 40\ k\Omega,\ R_M = 500\Omega,$$
$$R_V = 2\ k\Omega, \text{ and } \therefore R_s = 400 \text{ ohms.}$$

If $g_m = 2.5$ mA/V, then with R_1 and R_0 chosen to satisfy (2), we obtain

$$\eta = 0.769 \text{ (i.e., 2.28 db loss).}$$

If subsequently g_m falls to 1.25 mA/V, then η becomes 0.764 (i.e., 2.34 db) . . . a change of only 0.06 db. If the effect of g_m is ignored altogether (i.e., put $g_m = 0$ in (1a)), then we obtain $\eta = 0.753$, i.e., 2.48 db. It can be seen, therefore, that in many cases the effect of g_m can be ignored in design calculations, and that unless g_m is very large, it is not likely to cause any serious instability in the efficiency of the modulator.

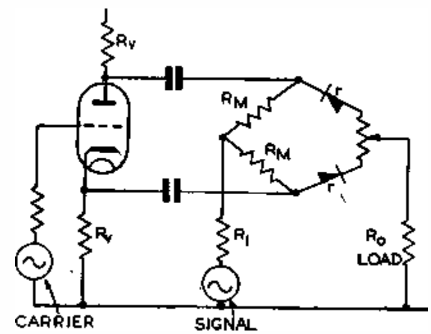
It is possible to use four rectifiers in the bridge by replacing R_M by a rectifier. This is not usually very useful, because each of the additional rectifiers is shunted by R_V and one by $1/g_m$. The improvement in efficiency is thus small, and in the numerical example taken above, the efficiency is increased by no more than 0.17 db.

Another disadvantage of the 4-rectifier circuit is that a D.C. load must be provided for the rectifiers, since they do not form a ring, but must be arranged as shown in Fig. 7. The load R_D does not enter into the signal circuit, of course, but it does fall across the carrier circuit, adding $\frac{1}{2}\ R_D$ in parallel with anode and cathode loads of the valve. Since for satisfactory operation of the rectifiers the value of R_D cannot be high, it does form an undesirable restriction on the operation of the valve.

An alternative arrangement, which removes this restriction, is to use a rectifier in place of R_D, so arranged as to pass the same current as the others, but on the opposite half-cycle of carrier. There is then no resultant current to build up a charge on the capacitors.

It should be noted that the value of R_M in the 2-rectifier circuit should not be high if the rectifiers are to operate efficiently, since when their D.C. load is of too high resistance a bias voltage is built up across the capacitors which tends to keep the

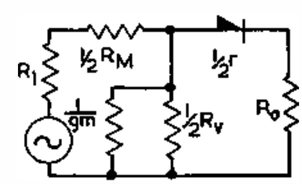
Fig. 4. Series modulator with two rectifiers



rectifiers in the backward condition. Experience shows that when R_M does not exceed about twice the forward resistance of the rectifiers, the bias causes less than 0.1 db reduction in efficiency; and although the effect is not independent of the value of the coupling capacitances, it appears that any value which is satisfactory for coupling purposes will not cause any appreciable reduction of efficiency. Some effects of bias (which include the reduction of certain higher-order modulation products) are discussed elsewhere.[1]

Adjustment of the balance for carrier leak can be made by means of the potentiometer shown in Fig. 1.

Fig. 5. Equivalent signal circuit of Fig. 4



Series Modulator

A series modulator is shown in Fig. 4; it uses two rectifiers, and is superficially the inverse form of Fig. 1. It is different, however, in one important respect; the anode and cathode loads and the internal resistance of the cathode circuit, remaining as before, now cause the equivalent signal circuit to appear as a T-pad, as shown in Fig. 5. This very seriously complicates the determination of the efficiency, and there is now no single optimum set of values of R_1, R_M, R_V and R_0.

The efficiency, defined as before, is the difference

$$\eta = [1/L]_f - [1/L]_b$$

and it can easily be shown that, if we put $\frac{1}{2}R_P$ equal to $\frac{1}{2}R_V$ in parallel with $1/g_m$,

$$1/L = R_P(R_1 + R_0) \Big/ [\tfrac{1}{2}R_P(2R_1 + R_M) + (R_0 + \tfrac{1}{2}r)(R_P + 2R_1 + R_M)] \quad \ldots (4)$$

where $r = r_b$ or r_f as the case may be.

Generally it will be found convenient to make $R_M = r_f$ and $R_1 = R_0 = R$, and if we put $R_P = 2kR$, $R = \frac{1}{2}m\sqrt{r_b r_f}$ and $n^2 = r_b/r_f$, then we obtain

$$[1/L]_f = 2k/[k(1+1/mn) + (1+1/mn)(1+k+1/mn)]$$

and

$$[1/L]_b = 2k/[k(1+1/mn) + (1+n/m)(1+k+1/mn)] \quad \ldots\ldots (5)$$

which is an easy form for computation.

Fig. 6 shows the effect of varying k for three values of m, taking $n^2 = 100$ in all cases. Since, in practice, k cannot be very large (because of difficulties in the valve circuit), it is evident that this modulator is less efficient than the shunt type.

The circuit can be rearranged for four rectifiers as shown in Fig. 7, and the equivalent signal circuit of Fig. 8 is obtained. The remarks made in Section 2 about the D.C. load R_D apply equally here. It is clear that the conditions for maximum efficiency are again irrelevant, since they require $R_P = \infty$. Again taking $R_1 = R_0 = R$ and $R_P = 2kR$, the efficiency can easily be shown to be $\eta =$

$$\frac{2k[(1+n/m)(1+2k+n/m) - (1+1/nm)(1+2k+1/nm)]}{(1+n/m)(1+1/nm)(1+2k+n/m)(1+2k+1/nm)} \quad \ldots\ldots (6)$$

Fig. 9 shows how the efficiency varies with k and m, taking $n^2 = 100$ again. It can be seen that, in general, no particular advantage is obtained from the use of four rectifiers instead of two. In specified cases one or other arrangement may prove to be more efficient; for instance, taking $m = 0.5$ as a requirement of the external circuits, then if R_P is restricted so that $k = 1$ in both cases, two rectifiers give 6.41 db loss (i.e., $\eta = 0.478$) and four give 5.73 db loss (i.e., $\eta = 0.517$), so that the four-rectifier circuit gains by nearly 0.7 db.

The Carrier (or Switching) Circuit

The valve which feeds the carrier to the rectifiers is a conventional phase-splitting arrangement. It functions as a triode, and if a pentode is used for convenience, its screen grid must be strapped to the anode. The effective anode and cathode loads comprise R_V, R_M and the rectifier resistance in parallel. These loads are therefore small. The valve stage introduces a loss, so that the grid amplitude has to be large to produce an adequate switching voltage across the rectifiers; therefore care has to be taken to see that grid current is avoided, because this would flow through the cathode resistor and unbalance the modulator. If grid current is avoided, then the carrier-leak balance and the efficiency are both independent of variations (within reasonable limits) of the carrier voltage, but a grid current of more than a few microamps removes this desirable independence.

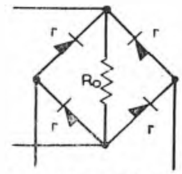


Fig. 7. Modification to Fig. 4 for four rectifiers

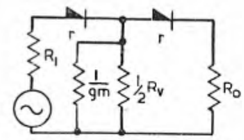


Fig. 8. Equivalent signal circuit for modified arrangement

Reference

[1] As shown, *e.g.* in D. G. Tucker, *J.Inst.Elect.Engrs.*, *95*, Part III. p.116 (1948).

Fig. 9. Variations of efficiency of series modulator (using four rectifiers) with circuit parameters.

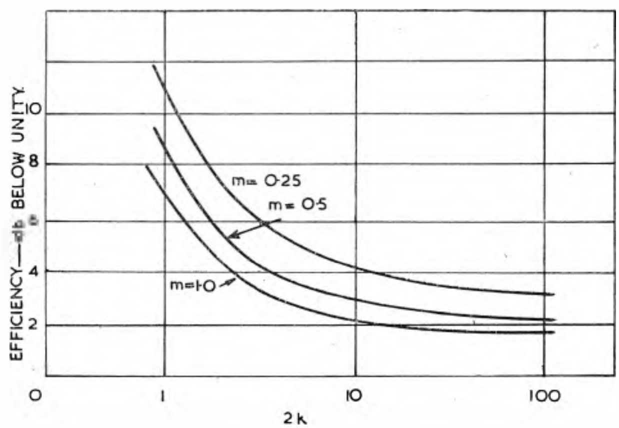
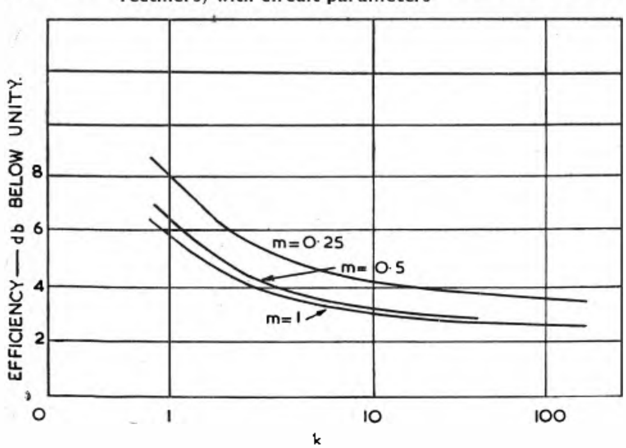


Fig. 6. Variation of efficiency of series modulator (using two rectifiers) with circuit parameters



ERRATA

We regret that some errors occurred in the article entitled "The Miller Circuit as a Low Speed Precision Integrator" in the February, 1950 issue. Corrections are shown below:

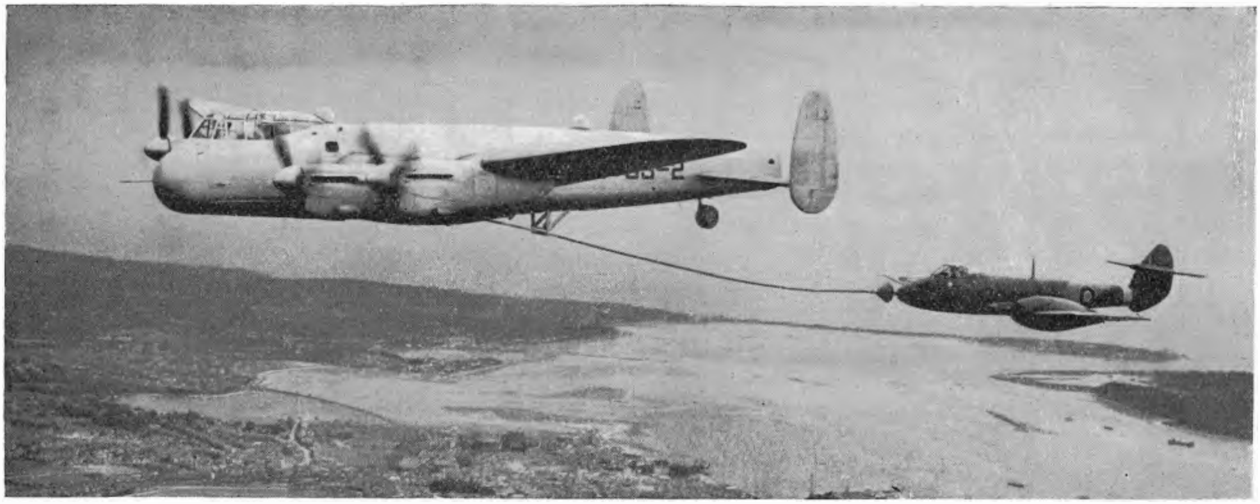
Page 67, Equation (2): *for* V in the denominator *read* V in the numerator

Page 67, line 17: *for* $C1\ C2 \geqslant C$; *read* $C1\ C2 \leqslant C$;

Page 67, Equation (5): *for* V/KT as an added term *read* V/KT as a factor

Page 67, Equation (7): *for* t_1 as upper limit of inner integral *read* t'

Page 68, column 2, line 9 of Table: *for* t_{max} *read* t_{1max}



Refuelling in Flight

The Use of Radar for Interception

By C. F. RAWNSLEY, A.M.I.E.E.*

EVERYONE now knows the story of the development of radar for the interception of bombers during the war. The peaceful application of this technique, to enable trans-ocean airliners to be intercepted and refuelled in flight, is perhaps not so widely known.

Messrs. Flight Refuelling, Ltd., developed such a system of interception for refuelling a series of trans-Atlantic flights. First, in conjunction with the British South American Airways Corporation, operating Lancastrians non-stop between London and Bermuda, with tanker aircraft based at Santa Maria, in the Azores. Secondly, with B.O.A.C. Liberators flying non-stop between London and Montreal, with the company's tankers based this time at Shannon (Eire), Gander (Newfoundland), and Goose Bay (Labrador). This second series of flights was designed to test operations under North Atlantic winter weather conditions.

To those who had any experience of war-time interceptions, the problem under consideration would appear at first sight to be very simple, since the "target," instead of taking evasion, flies straight and level, signals its position and progress to the intercepting tanker, and carries full lighting after dark. But against these advantages it should be considered that interception often has to be made far from land, in areas poorly covered by navigational aids, and without the aid of ground control radar. The heavily-loaded tanker has no speed advantage over the "receiver" aircraft, so that stern chases are out of the question. The interception must be smooth, sure, and on time. Moreover, on the principle that "the customer is always right," the receiver aircraft must be allowed to proceed unhindered on its course throughout the operation, and the radar of the intercepting tanker must have sufficient range to allow for the combined track errors, which may be considerable after an ocean flight of many hours duration.

The solution of the problem depends, therefore, primarily on two things. Firstly, on good communications, so that the tanker base and the tanker aircraft have ample warning of the receiver's movements. The tanker takes off with sufficient time in hand to reach a pre-arranged rendezvous point ahead of the receiver, which keeps the tanker informed of its progress, and passes any information required. Secondly, a device is needed which will give the earliest possible indication to the tanker of the range and relative bearing of the receiver. Knowing the speeds and headings of the two aircraft, and the relative bearing and range, the tanker navigator then has all the data to solve the interception problem. The effect of drift is eliminated by assuming a common wind and considering air positions only. Relative height is arranged by comparison of altimeter settings. The tanker usually maintains a vertical separation of 1,000 ft. above the receiver until ready to formate.

The obvious choice of equipment for this purpose is some form of secondary radar. Consideration has to be given to reliability, availability, simplicity of operation, and any extra equipment carried by the receiver aircraft must be reduced to a minimum in weight and bulk. It was therefore decided to fit the tanker aircraft with Rebecca Mark II, and to install a light weight 200 Mc/s. responder beacon in the

* Flight Refuelling, Ltd.

customer's receiver aircraft. Similar beacons were also set up at tanker bases where this facility was not already provided for homing.

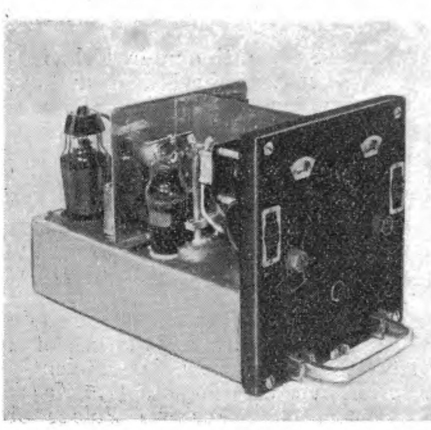
For the first series of flights a standard Eureka beacon was used. Satisfactory interceptions were made, but it became obvious, from the considerable variation in maximum ranges obtained, and this piece of equipment was not robust enough nor sufficiently stable for operation over long periods.

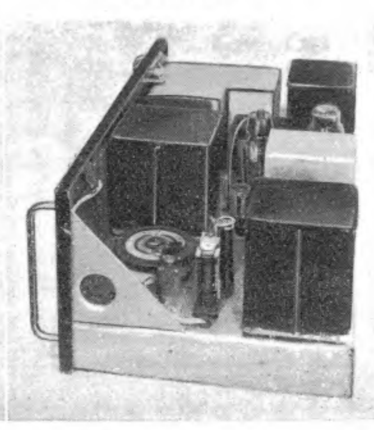
For the second series, Eureka beacons were modified in Flight Refuelling radio workshops so as to give a higher output, and to maintain correctly their pre-set frequencies. Duplicate sets were installed in the B.O.A.C. receiver aircraft as an additional precaution. The tanker's interception equipment was also improved by the provision of a duplicate set of Rebecca antennae on the tail. The change-over from nose to tail arrays could be effected by a switch mounted on the indicator unit. The reason for this modification was that, owing to the relative geographical positions of the tanker bases and the probable refuelling points, it was apparent that the tanker would have to fly out ahead of the receiver, sometimes for considerable distances and on almost identical tracks. Considerable benefit was, in fact, derived from the ability to get radar contact at long range during the flight out, without having to turn the tanker about.

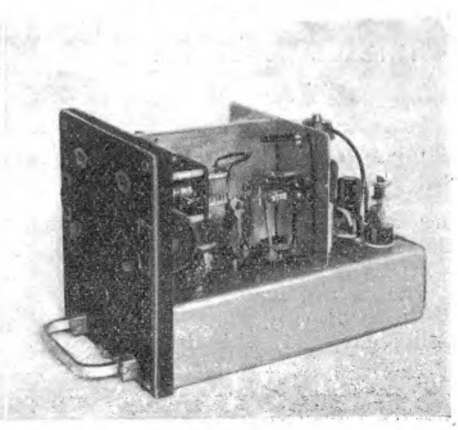
It is, in fact, this limitation of the Rebecca-Eureka combination that has governed the tactics evolved for tanker interception. Since a reasonably accurate bearing can only be obtained when the other aircraft is dead ahead (or astern), the tanker is usually manoeuvred so that the two aircraft approach one another on reciprocal headings. From the initial data obtained, the tanker navigator computes the course to steer and the time required to bring his aircraft by the shortest route on to this reciprocal course. Considerable use has also been made of radio compass bearings on transmissions from the receiver aircraft, to get additional or check bearings and to avoid the repeated turns necessary when taking Rebecca bearings. As the two aircraft approach one another, the remaining displacement errors become apparent, and "dog-leg" corrections are made. The final turn on to the receiver's course is started at a pre-determined range, according to the air speeds of the aircraft, so as to bring the tanker into loose formation, whether or not visual contact has been made before or during the turn. Safety height separation is, of course, maintained. The Rebecca has occasionally been used as primary radar to obtain more accurate information at close range, but in general this has not been found necessary.

As a result of these flights it was clear that to ensure reliable and satisfactory interception, the paramount need (apart from communications) was for a really reliable, compact, light weight responder beacon. Consideration was given to the centimetric bands, but in view of the development work that would be involved, and the cost and doubtful availability of suitable interrogator equipment, Flight Refuelling decided to develop in their radio workshops at Tarrant Rushton a responder beacon working in the 200 Mc/s. band. This is made up in two units, a transmitter-receiver and a power unit, both fitting standard S.B.A.C. racking. The T/R unit measures 8 in. wide, $8\frac{1}{2}$ in. high, and $12\frac{1}{2}$ in. deep, and weighs 11 lb. The airborne power unit is the same size, and weighs about 15 lb. The beacon is triggered by an input of less than 20 microvolts, and gives a peak power output of over 300 watts. The T/R unit can be continuously tuned by thumb knobs over a range of 212/240 Mc/s., and the tuning locked in any position. Well-defined width or gap coding can be quickly reset to any desired combination by means of an ingenious coding device incorporated in the power unit. An alternative power unit for ground use, suitable for universal power supplies, has also been produced. This measures 21 in. wide, $8\frac{1}{2}$ in. high, and $12\frac{1}{2}$ in. deep, and weighs 67 lb.

Although the advent of near sonic speeds and other factors may eventually necessitate a change over to the higher frequencies, it is felt that this 200 Mc/s. beacon will fulfil the needs of the company for some time to come. It has already given very satisfactory results in air-to-air and ground-to-air tests. It would also prove to be a remarkably compact and easily transportable homing beacon for ground use elsewhere. The arrival of Rebecca Mark 4 and the further delay in the introduction of 1,000 Mc/s. D.M.E. will tend to prolong still further the demand for an efficient beacon working in the 200 Mc/s. band.


F.R. radar responder beacon-T/R unit transmitter side


Ground power unit showing coding device


T/R unit—receiver side

Pre-amplifiers for the Televisor

By W. I. FLACK, M.I.R.E.

In response to appeals from many Televisor enthusiasts in the fringe areas, Mr. Flack has designed the two pre-amplifier circuits described.

WITH the opening of the Sutton Coldfield station and the ensuing reports of long distance reception, there has arisen quite a demand among Televisor constructors, particularly towards the Holme Moss area, for a pre-amplifier unit. Obviously, no circuit, however sensitive, will give perfect reception in some areas, but in view of the fact that the Midlands Televisor will be very easily convertible to receive Holme Moss, it is felt that even a unit which may at the moment allow only five or so days' out of seven worthwhile reception in extreme cases will be considered by many enthusiasts to be worth

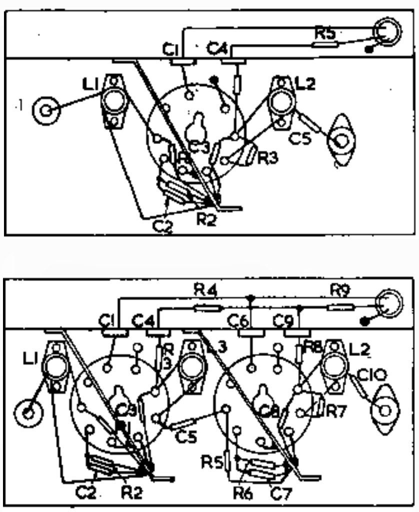


Fig. 1. Layout of both types of amplifier

construction. It need not be emphasised that in all areas of low signal strength where reception is poor a pre-amplifying circuit is a necessity.

Two types of pre-amplifier have therefore been developed, one having a single R.F. stage while the second has two R.F. stages. Each of these pre-amplifiers may, by suitable choice of coils and damping resistors, be used for reception of Alexandra Palace or Sutton Coldfield, details of the modifications to the coils and damping resistors being given in the lists of components.

The choice of pre-amplifier must be left to the individual and will depend on whether a large increase in signal input is required or whether it is only necessary to increase the signal so as to improve picture contrast.

The pre-amplifiers are both built on small chassis and have layouts similar to the Televisor vision receivers, this method of construction having been found to be extremely stable and trouble free; see Fig. 1. Dealing with the single-valve version first, Fig. 2, it will be noted that the aerial is coupled directly into the bottom of the grid coil, which is damped by a parallel resistor. The anode coil is also damped by a resistor of low value, the output of

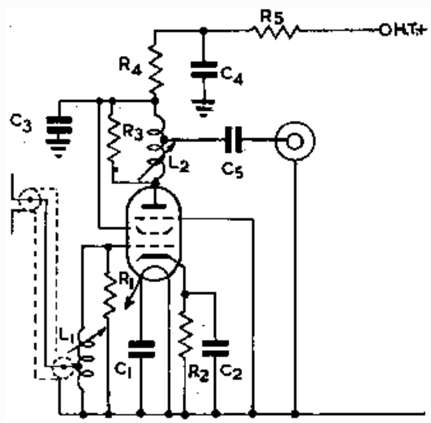


Fig. 2. Schematic of single valve amplifier.

For A.P.— R1 = 3.3kΩ, R3 = 2.2kΩ
For S.C.— R1 = 6.8kΩ, R3 = 4.7kΩ

R2 = 220Ω
R4 = 6.8kΩ
R5 = 2.2kΩ
V1 EF50

C1 500 pF. Micadisc T.C.C.
C2 0.001 μF CM20. Mica T.C.C.
C3 0.001 μF CM20. Mica T.C.C.
C4 500 pF Micadisc T.C.C.
C5 0.001 μF CM20. Mica T.C.C.

Note changes in values of R1, R3 for S.C.

the amplifier being taken from a tapping at the low potential end of this coil. A .001 μF isolating capacitor is connected from this tap to a Belling-Lee coaxial socket.

As the unit will be fed from the main Televisor supply, decoupling of the H.T. is effected by making the value of the decoupling resistor 6.8k. For convenience in use, the pre-amplifier may be clamped to the gantry close to *V*1 so that the short length of coaxial feeder connected to the grid of *V*1 will now plug in to the pre-amplifier output socket and the feeder lead from the pre-amplifier will be taken direct to the aerial. The coaxial feeder coupling the receiver to the pre-amplifier may, of course, be shortened suitably.

Adjustment to the iron dust cores is extremely simple, and does not normally require the use of a signal generator since the band width is sufficiently wide to permit reception over the range of 41.5 Mc/s. to 48 Mc/s. for reception of Alexandra Palace or

58 Mc/s. to 62.5 Mc/s. for reception of Sutton Coldfield, provided that the two tuned circuits are staggered very slightly.

It is sufficient when tuning, to adjust the aerial circuit for maximum response, then screw in the iron dust core very slightly, after which the anode circuit should be tuned for maximum response and then the iron dust core unscrewed very slightly.

The gain of this single-valve unit should be not less than 14 to 15 db and should increase the sensitivity of the receiver by a worthwhile amount.

Where conditions are exceptionally bad the two-stage pre-amplifier is advised. This will not, however, give a gain double that of the single-stage unit, since the band width must still be maintained and, more important, the stability must not be allowed to suffer. Instability would result in loss of gain, and definition would suffer, while the background of the picture would become very dirty.

The two-stage pre-amplifier gives an overall gain of 24 to 26 db, its limitation in use being the unavoidable increase in background noise when at full gain.

From the circuit diagram, Fig. 3, it will be seen that the typical Televisor vision receiver circuit and layout have again been used, the aerial input being taken to a tap on the grid coil, damped by a suitable value of resistor. The coupling between the two stages is tuned anode, the grid resistor of the second valve damping the circuit effectively.

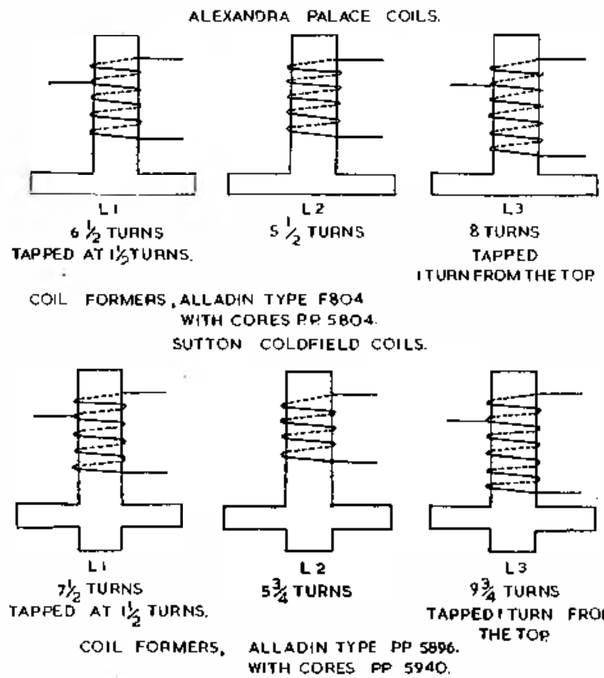


Fig. 4. Coil winding data

Fig. 3. The two-stage amplifier

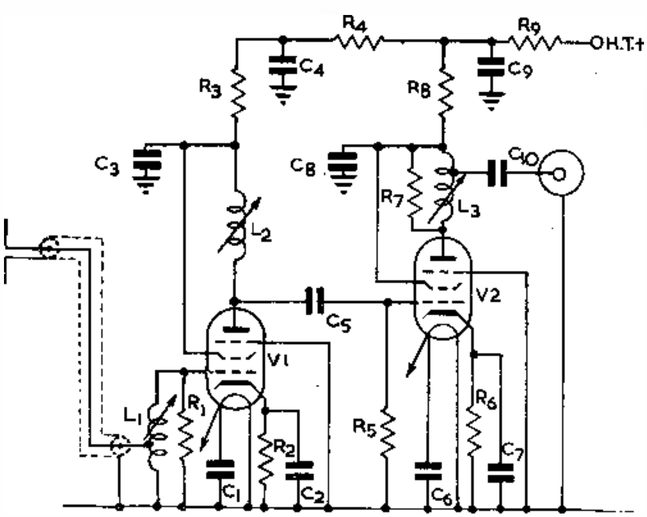
For A.P.—
$R1 = 2.2k\Omega$
$R5 = 2.2k\Omega$
$R7 = 1k\Omega$

For S.C.—
$R1 = 3.3k\Omega$
$R5 = 3.3k\Omega$
$R7 = 2.2k\Omega$

R2 220Ω
R3 6.8kΩ
R4 2.2kΩ
R6 220Ω
R8 10kΩ
R9 1kΩ

V1, V2 EF50

C1 500pF Micadisc T.C.C.
C2 0.001μF CM20. Mica T.C.C.
C3 0.001μF CM20. Mica T.C.C.
C4 500pF Micadisc T.C.C.
C5 0.001μF CM20. Mica T.C.C.
C6 500pF Micadisc T.C.C.
C7 0.001μF CM20. Mica T.C.C.
C8 0.001μF CM20. Mica T.C.C.
C9 500pF Micadisc T.C.C.
C10 0.001μF CM20. Mica T.C.C.



The amplified signal from stage two is taken from a tap at the low potential end of the anode coil through a .001 μF isolating capacitor to a Belling-Lee socket.

The anode coil of $V2$ is damped by a low resistor to ensure an adequate band width, while decoupling on both valves is effected by the use of fairly high value resistors. In service this unit may also be clamped to the gantry or mounted on the main Televisor chassis close to the input of $V1$, the aerial feeder then plugging in to the output socket.

For convenience the feeder coupling the pre-amplifier to $V1$ can be shortened until it is just long enough; this may result in some increase in signal strength.

Alignment of the two-stage unit can also be done on the signal. The aerial coil is adjusted for maximum response, after which the iron dust core is screwed in by approximately a quarter turn; the second tuned circuit then being tuned for maximum response and the iron dust core unscrewed by approximately the same amount. Finally, the output circuit may be tuned for maximum response, without further detuning, since it is already adequately damped. No difficulty should be experienced in operating either of these units, and if they are mounted at the input end of the vision receiver there should be no chance of instability.

It is confidently expected that the pre-amplifier will considerably extend the range of the Televisor, the probable limiting point being background noise. When this stage is reached nothing more can be done in the receiver; efforts should be made to improve the aerial system so as to improve the signal-to-noise ratio.

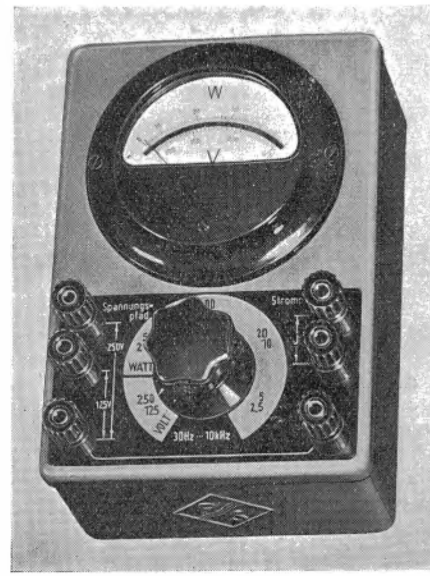
Dynamometer Type Wattmeter
for Audio Frequencies

By D. ALVEY, B.Sc. *

THE use of the dynamometer in the field of A.F. voltage, current and power measurement is a comparatively new development. Until recently its application had been limited to frequencies up to about 300 c/s., its development for use at higher frequencies being limited by the variation of meter consumption with frequency.

The extension for use over the full audio frequency range with an accuracy of 1 or 2 per cent has been achieved by adopting a light weight construction, and the incorporation of a suitable frequency correcting network to reduce power consumption. A range of instruments has been produced covering voltage determinations to 300 volts, currents to 3 amps. and powers up to 500 watts.

An analysis of the major sources of error and the design of these instruments will now be considered.



One of the new dynamometer type wattmeters

Basic Principles

Fig. 1(a) shows the basic arrangement of the dynamometer. A fixed coil $L1$ is wound in two sections and a moving coil $L2$ is suspended so that it rotates in the almost uniform field set up by the current in $L1$. Fig. 1(b) illustrates the circuit arrangement, $R1$ and $R2$ being the resistances of $L1$ and $L2$ respectively; M being the mutual inductance between $L1$ and $L2$. To avoid distortion of the field surrounding $L1$, metal parts have been eliminated in the region of the coil field and self-supporting air cored coils are employed.

The magnetic moment of the circulating current, $I2$, is given by the area of the coil, $A2$, multiplied by the current $I2$, ie., $A2I2$. The turning couple of the moving coil $L2$, assumed parallel with $L1$, is given by $A2I2H$, where H is the number of lines per unit area cutting $L2$ due to current in L1. By definition of mutual inductance M, the total number of lines cutting the moving coil is $MI1$, so that H is given by $MI1/A2$.

Therefore the turning moment D is $A2I2MI1/A2$. i.e., $D = I1I2M$. This assumes in-phase currents and a constant mutual inductance. To take into account the variation of the mutual inductance as $L2$ rotates and also a possible phase angle ϕ between the currents we must write M as $k\sqrt{L1L2}$ and hence

$$D = I1I2\sqrt{L1L2}\cos\phi\, F.(\theta) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

$F(\theta)$ is written to account for the variation of the coefficient of coupling k, which is a function of the angle of pointer deflexion θ.

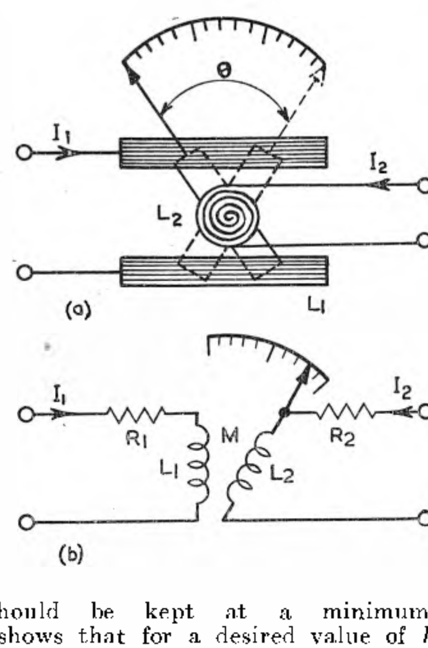


Fig. 1. Physical and circuit arrangements of the dynamometer

The deductions made from (1) depend upon the application of the meter.

(i) As a Current Meter.

When used for alternating current measurement, the two coils are switched in series and the inductance of the instrument is $(L1 + L2 + 2k\sqrt{L1L2})$. In order that the meter shall have a response which does not vary rapidly as frequency increases, the inductance should be kept at a minimum. Equation (1) shows that for a desired value of D the product $L1L2$ is fixed, and thus the inductance L of the meter is given by

$$L = L1 + L2 + \text{const.} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

k being fixed by the geometry of the instrument. For L to be a minimum, differentiation of (2) shows that $L1$ and $L2$ must be equal. This cannot be satisfied very easily in practice since, in the interests

* Dawe Instruments, Ltd.

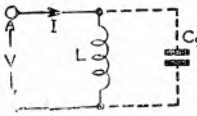


Fig. 2 (left). Effect of coil self-capacitance on current meter

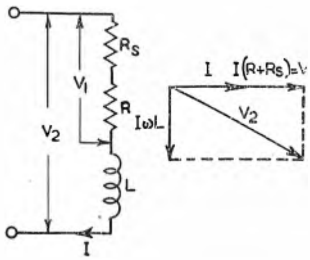


Fig. 3 (right). Circuit arrangement of voltmeter

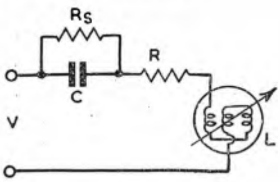


Fig. 4 (left). Circuit of voltmeter with equaliser added

of obtaining a sufficient turning moment with the smallest possible weight in the moving coil, $L1$ has to be made much larger than $L2$. As a result there is a large reactive power consumption Px, or poor power factor, at high frequencies which for the basic movement of the instrument is shown in the table:

Frequency	0.1	1	10	20	kc/s.
Px ...	8.5	85	850	1700	mW
C ...	42.3	0.423	0.00423	0.00106	μF

The capacitance C is that required in series with the windings to compensate for the inductive nature of the meter and, unfortunately, requires adjustment for each frequency. Should this be done, there is also the question of the stray self-capacitance Cs of the coils. This introduces a measurement error Δ which is derived as follows:

Referring to Fig. 2, the current drawn from the supply in the absence of Cs is $V/\omega L$. When Cs is included, the current drawn is $(V/\omega L - V\omega Cs) = V\ (1/\omega L - \omega Cs)$. Thus Cs makes a difference of $V\omega Cs$ in a value of $V\ (1/\omega L - \omega Cs)$, i.e., a fractional error Δ in measurement of

$$\frac{V\omega Cs}{V\ (1/\omega L - \omega Cs)} = \frac{\omega Cs}{(1/\omega L - \omega Cs)}$$

$$\text{i.e., } \Delta = \frac{\omega^2 LCs}{1 - \omega^2 LCs} = \omega^2 LCs \text{ approx.}$$

when Δ is a small fraction.

From the above we see that Δ increases as the square of the frequency, but, since $L = 60$ mH approx. and $Cs = 10$ pF the error even at 30 kc/s. is only about 2 per cent, so there is no point in reducing the number of turns to reduce Δ since this would also produce an undesirable decrease in current sensitivity.

(ii) As a Voltmeter.

When the two windings are in series, full scale deflexion is obtained with a current of 15 mA. For the basic movement this requires only 6.7 volts, so series resistances are added to give higher voltage ranges.

If the series resistance is denoted by Rs and the resistance of the windings by R, then we have the conditions shown in Fig. 3.

For a given current I, the presence of L changes $V1$ to $V2$. Thus there is an error during volts measurement of Δ v. This fractional error is given by

$$\Delta \text{v} = \frac{V2 - V1}{V1} = \frac{V2}{V1} - 1$$

From the vector diagram

$$V2 = I\sqrt{(R + Rs)^2 + (\omega L)^2}$$

i.e., Δ v =

$$\frac{I\sqrt{(R + Rs)^2 + (\omega L)^2}}{I\ (R + Rs)} - 1$$

$$= \sqrt{\frac{(R + Rs)^2 + (\omega L)^2}{(R + Rs)^2}} - 1$$

$$= \left\{1 + \left(\frac{\omega L}{R + Rs}\right)^2\right\}^{\frac{1}{2}} - 1$$

$$= \tfrac{1}{2}\left(\frac{\omega L}{(R + Rs)}\right)^2 \text{approx.}$$

Thus the error in voltage measurement due to the inductive nature of the meter increases as the square of the frequency and if we fix the permissible value of Δ v, the useful frequency range of the instrument is defined. The upper frequency limit follows from the above expression as:

$$\omega \text{ max.} = \frac{R + Rs}{L}\sqrt{2\ \Delta \text{ v}}$$

The value of Rs is fixed by the required voltage V, and if I is the full scale deflexion (f.s.d.) current, we have:

$$V/I = R + Rs, \text{ which gives}$$

$$\omega \text{ max.} = \frac{V}{IL}\sqrt{2\ \Delta \text{ v}} \qquad \text{.............. (3)}$$

This shows the frequency limit to vary with the range of voltage measurement required, and for Δ v = 2 per cent the following table results:

Voltage range	25	60	125	250	volts
Frequency limit ...	890	2130	4400	8900	c/s

Although Equation (3) indicates that decreasing L will give an increase in $\omega_{max.}$ it will also increase the f.s.d. current I of the meter, which in turn restricts $\omega_{max.}$ as in (3).

In order to raise $\omega_{max.}$ without increasing the f.s.d. current an equalisation network is used. The circuit of the instrument is then as in Fig. 4. By suitable choice of C the increase of inductive reactance with increase of frequency is compensated for by the increase of capacitive reactance so that the apparent resistance seen at the input terminals is almost constant. By making $C = 100$ pF, Δv is less than 1 per cent at 10 kc/s., so the useful frequency range is extended considerably.

(iii) As a Power Meter.

The arrangement of the instrument as voltmeter/wattmeter is shown in Fig. 5. Now consider power measurement.

The inductance of the fixed coil will normally introduce an error when power is being measured. Without the CR_s network the loss angle ϵ of the voltage circuit will cause a distortion of the phase difference ϕ between the two measured quantities; while it is required to measure $VI \cos \phi$, the actual measurement will be of $VI \cos (\phi + \epsilon)$ and this causes a fractional error in the product measurement of:

$$\Delta\epsilon = \frac{VI \cos \phi - VI \cos (\phi + \epsilon)}{VI \cos \phi}$$

$$\text{i.e., } \Delta\epsilon = 1 - \frac{\cos(\phi + \epsilon)}{\cos\phi}$$

$$= 1 - \frac{(\cos\phi\cos\epsilon - \sin\phi\sin\epsilon)}{\cos\phi}$$

$$= 1 - \cos\epsilon + \sin\epsilon\tan\phi$$

Since ϵ is a small angle, we can write

$$\Delta\epsilon = 1 - 1 + \epsilon\tan\phi \text{ approx.}$$

$$\text{or } \Delta\epsilon = \epsilon\tan\phi$$

This error is proportional to the loss angle ϵ of the fixed coil circuit and is therefore largely reduced by the use of the CR_s network. The remaining error can be calculated from the data supplied with the instrument and the appropriate correction can be made.

Applications

As seen from Fig. 5, the coils are switched in series for voltage readings, while for power measurement the moving coil is supplied separately from a current transformer. The moving coil has an inductance

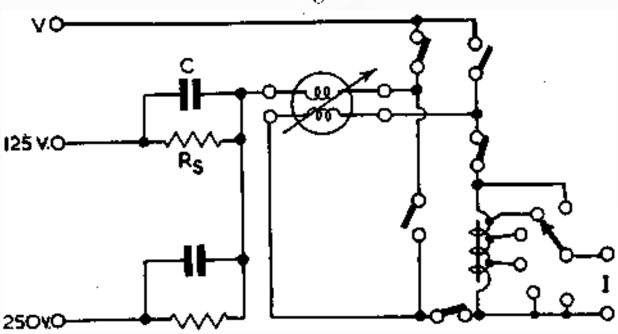


Fig. 5. Circuit of voltmeter/wattmeter (switches in wattmeter position)

of approximately 6 mH and a resistance of 150 ohms., while the fixed coil has an inductance of about 60 mH and a resistance of 300 ohms. The use of the range switches enables power measurements to be taken up to 500 watts, the lowest range permitting the reading of 0.05 watts. The actual ranges of the current and voltage circuits are:

Current Ranges		Voltage Ranges	
f.s.d.	Resistance	f.s.d.	Resistance
0.02 A	250 ohms	125 v.	8,000 ohms
0.08	15	250	16,000
0.4	0.6		
2.0	0.03		

The use of each of the current ranges in conjunction with the voltage ranges give the meter the following power ranges:

2.5 or 5 watts, 10 or 20 watts,
50 or 100 watts, 250 or 500 watts,

with an accuracy of ±2 per cent of full scale reading. The scale is calibrated in watts and volts, and may be used for a wide range of power measurements. The lower ranges can be used for measuring the output of small power valves and the consumption of the smallest transformers, while at the other extreme are the measurement of the output and performance of public address equipment and the consumption of mains and large audio-transformers.

Expression (1) shows the turning moment to depend upon the phase difference ϕ between the coil currents, thus a means is provided for measuring phase angles.

A problem which normally requires a great deal of time and several measurements is that of determining the optimum terminating impedance for a generator. The circuit of Fig. 6 shows how the optimum impedance Z' is easily obtained by varying Z for maximum power, Z' then being given by:

$$Z' = \frac{ZR1}{Z + R1}$$

This expression neglects $R2$, the resistance of the current circuit, since this is much smaller than Z in most practical applications. $R1$, however, which is the resistance of the voltage circuit and is in shunt with Z, must be taken into account.

The meter may also be arranged in the form of a voltmeter/ammeter, the instrument being described having the circuit of Fig. 7. The same basic move-

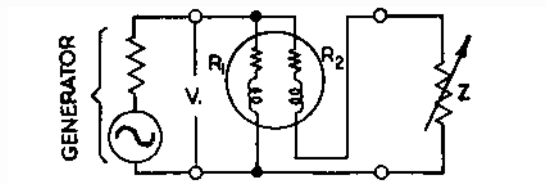


Fig. 6 Simplified arrangement for finding the optimum load for generator

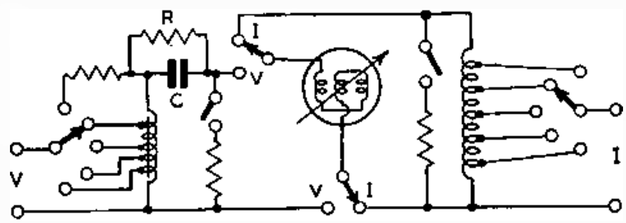


Fig. 7. Circuit of voltmeter-ammeter

ment is employed and an equalisation network RC in the voltage circuit. For range switching of both voltage and current, tapped transformers are employed. Calibration is in amps. and volts, the important data being as follows:

Current Ranges			Voltage Ranges	
f.s.d.	Resistance ohms	Inductance mH.	f.s.d. volts	Resistance ohms
30 mA	145	17	10	64
100 mA	15	1.6	30	629
300 mA	2	0.17	100	6,200
1 A	0.13	0.02	300	18,500
3 A	0.02	0.003		

The power consumption of approximately 0.2 watts on the current range is less than can be achieved with soft iron instruments, although the figures given for the voltage ranges show that its application as a voltmeter may be somewhat restricted by the consumption. A valve preamplifier can, of course, be used to reduce the loading of the instrument.

Some additional information concerning the performance of these instruments is contained in *Archiv für Technisches Messen*, J741-11, p. 124, December, 1948.

The Dynamometers described are manufactured by Messrs. Rohde & Schwarz, of Munich, Germany, and further information concerning their availability in this country is obtainable from Dawe Instruments, Ltd., of Hanwell, London, W.7.

A New Method of Synchronisation
For Long Range TV Reception

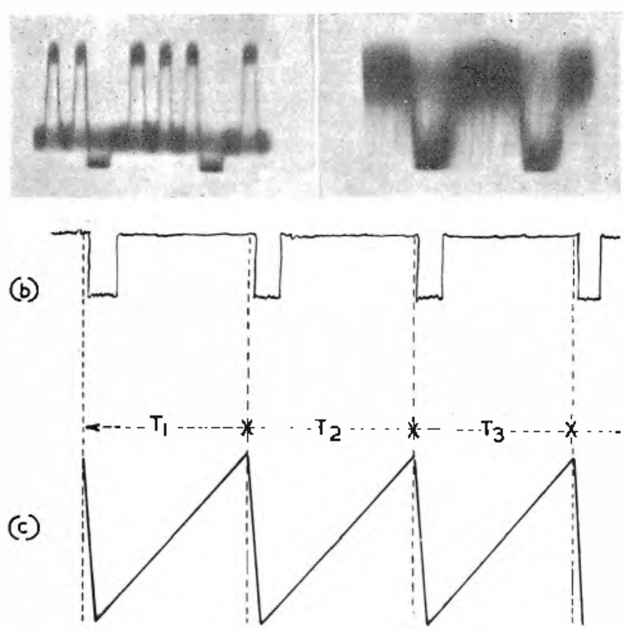
By T. B. TOMLINSON, B.Sc.(Hons.) Assoc.Brit.I.R.E.*

THE effect of noise on a television picture such as is obtained beyond the so-called fringe areas is two-fold. The first and obvious effect is that the noise components appear at the detector more or less superimposed on the modulation envelope in a substantially positive phase (Fig. 1(a)). This produces on the received picture a pattern of bright spots whose occurrence is completely random and shows up as a "speckled" appearance or, if viewed from a distance, a "mistiness" over the picture proper with a consequent loss of definition.

The second effect is illustrated by Fig. 1(b): even with quite elaborate circuits the separated synchronising pulses are likely to include irregularities due to the presence of noise. As in the more distant reception areas the synch. pulse is often completely filled with noise pulses, the possibility of obtaining a clean, noise free, separated pulse appears rather remote. The result is a tendency for the line time-base generator discharge to be initiated by a noise pulse appearing before the synch. pulse leading edge in the normal triggered type synchronising systems. This leads to a random variation in the exact moment at which the time base flyback commences, as shown in Fig. 1(c) where the periods $T1$, $T2$, $T3$, etc., are not equal as they would be in noise-free circumstances.

* Electronics Dept., University College, Southampton.

Fig. 1 (a) (Photograph) Noise on the modulation envelope. (b) and (c) Noise triggering the fly-back circuit at irregular intervals



The final effect is a lateral displacement of the scanning lines which is noticeable in the irregular or "ragged" vertical edges of the picture. The same effect occurs in the frame time base but, due to the much longer scanning period, the vertical displacement is unnoticeable except as a random fluctuation of interlace which is not nearly so disturbing to the eye.

The line displacement effect is generally far more serious than the brightness modulation effect. This can be readily demonstrated in an area of high signal strength by using a noise-free signal to operate the synch. system and by using a very noisy video signal from another receiver with a poor aerial system to modulate the cathode-ray tube. It will be seen that a video signal having noise pulses completely filling the synch. pulse will give a picture with entertainment value, but if the synchronisation is operated in a normal (trigger-type) manner from the same signal then the definition is so poor as to become intolerable.

Due to this and to ignition noise effects, several new type synchronising systems have been evolved and are generally known as automatic frequency-phase control or "flywheel synch." circuits.[1,2] Some of these have been tried out by the author with but moderate success, the only really satisfactory method appearing to be that known as the "sine-wave" method.[2] This makes use of a stable sine-wave oscillator (10,125 c/s.) with a reactance valve and phase discriminator connected in a similar manner to that used in automatic frequency control circuits (Fig. 2).

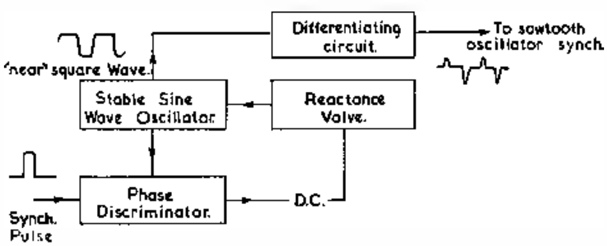


Fig. 2. Sine-wave "flywheel" synchronisation

An attempt is made to obtain a nearly square wave output from the oscillator which, after differentiation, provides a short pulse for synchronising a normal type sawtooth generator. In order to obtain the correct phase, i.e., for the actual synchronising pulse and the leading edge of the transmitted synch. pulse to be coincident it is necessary to adjust the tuned circuit of the discriminator to a frequency somewhat lower than 10,125 c/s. In practice, the synch. pulse obtained is a wide pulse of comparatively slow rise-time which results in a slowing up of the sawtooth generator flyback time. A better method is to pass the output of the sine-wave oscillator into a high

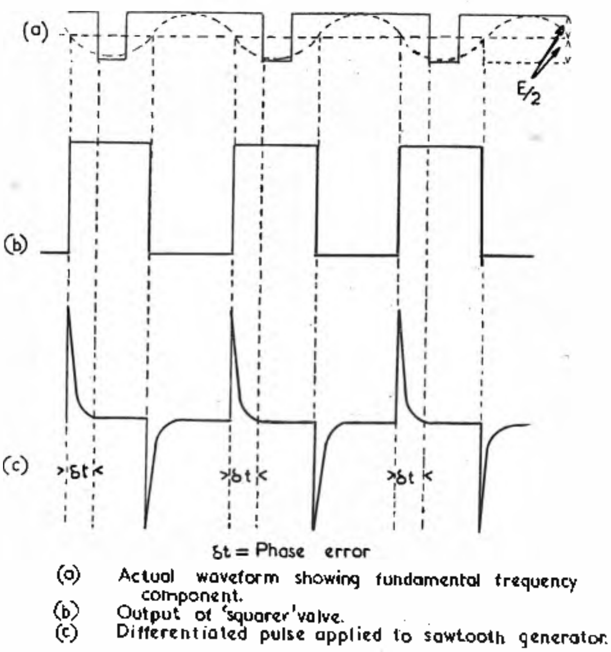


Fig. 3. The circuit arrangement of Fig. 2 gives the differentiated output shown at (c)

slope pentode connected as a " squarer " valve whose output after differentiation gives a short pulse with a sufficiently fast rise time. The circuit is then satisfactory and shows good immunity from noise, and although elaborate it has the advantage of being automatic in action.

A simpler circuit, and one believed to be original, has been employed by the writer for a considerable time with greatly improved reception at a distance of approximately 80 miles from the Alexandra Palace transmitter. The principle is to use a selective amplifier[3,4] to extract the fundamental frequency component from the separated synch. waveform (Fig. 3(a)). The amplifier is of the type using a Wien bridge circuit and has a bandwidth of the order of 40 c/s. centred on 10,125 c/s., thus discriminating against the noise components in the waveform as well as all the harmonic frequency components. The circuit details are given in Fig. 4.

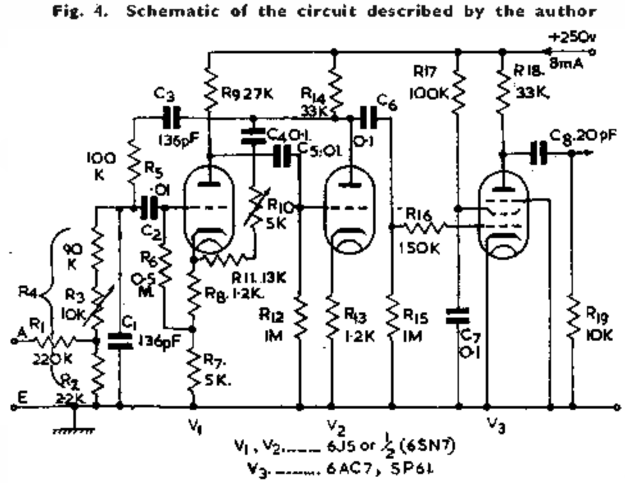
Briefly, there is frequently selective positive feedback via the bridge network $C3$, $R5$, $C1$, $R4$, the maximum feedback occurring at a frequency given by $fr = 1/2\pi\sqrt{R4R5C1C3}$ at which frequency the network phase shift is zero. At frequencies differing from fr the feedback voltage decreases and is shifted in phase closely similar to the action of a tuned circuit. When $R4 = R5 = 1/2\pi frC1 = 1/2\pi frC3$ it may readily be shown that the feedback voltage at fr is $1/3$, so that an amplifier gain of three times would just lead to oscillation. (See Appendix A). The gain is therefore reduced by negative feedback so as to be just less than this figure, the adjustment being provided by variable resistance $R10$, which acts as a reaction control.

The combined action of the two feedback networks is to give a high gain over a narrow band of frequencies and a low gain at all other frequencies. By applying only a small fraction of the separated synch. pulses to the amplifier by means of potential divided network, $R1$, $R2$, we can ensure that the output is an almost perfect 10,125 c/s. sine wave. The effective Q-factor obtainable with stability over a long period of time was found to be 250, with a corresponding amplification of 2,200 when the reaction control $R10$ was set just short of the oscillation point: a more critical adjustment gives a higher Q-factor, but more dependent on supply potentials. The sine wave output at the anode of $V2$ is of the order of 75 volts peak to peak and is fed into a " squarer " valve $V3$ whose output is differentiated by $C8$, $R19$ (Fig. 3(b) and 3(c)) and provides a short positive pulse which is used to synchronise the sawtooth generator. A blocking oscillator type was used, but any other standard circuit should be satisfactory. With negative synch. pulses applied to input at A the short pulse derived will be a positive pulse and vice versa, hence the type of sawtooth oscillator may be determined by the polarity of the separated synch. pulse. In this synch. system also, the centre tuning frequency must be slightly mistuned in order to get the correct phase relationship. (See Appendix B).

The circuit is substantially free from frequency drift, which is almost entirely due to temperature variation in the values of the bridge network components. For this reason high stability resistors and low tolerance protected silvered mica capacitors were used. However, it was soon realised that the main source of frequency drift was in the transmitted waveform due to the " spongey " lock with the 50 c/s. mains: there also occur quite sudden frequency changes in the fundamental, and considerable frequency fluctuation occurs on some outside broadcast transmissions. It is these effects which cause automatic phase control circuits to " hunt." As a result it was found that the phase of the square wave output varied sufficiently to cause considerable lateral displacement of the picture, part of which is then lost to view.

The best answer to this problem is to have the frequency adjustment variable resistor $R3$ on an extension cable and to operate it from the viewing position. This is not such a drawback as it sounds since it will only need occasional adjustment. In any case, for reception in and beyond the fringe areas it

Fig. 4. Schematic of the circuit described by the author



is an advantage to have the contrast control on an extension lead, so the writer has a small panel for the two controls which is placed on the arm of the chair at viewing times. The extension lead to $R3$ may be about 3 yds. of coaxial cable: this will have a capacitance of approximately 200 pF, whose reactance has negligible shunting effect on $R3$, even at its maximum value.

Since the phase changes more rapidly with frequency at higher Q values, it is not wise to adjust $R10$ too critically. The circuit will also operate when the reaction is adjusted to a continuously oscillating condition and the amplifier, now a sine wave oscillator, is synchronised in step with the input synch. pulse fundamental. Under these circumstances it will be found that the phase error changes rapidly with changes of fundamental frequency, resulting in a need for continual adjustment of the frequency control.

During the first part of the frame pulse, the line pulses are replaced by half line pulses and the 10,125 c/s. fundamental is temporarily absent. However, due to the high Q-factor of the circuit, the sine wave output dies away very slowly and does not modify the working of the circuit before the line pulses reappear, as its magnitude is always more than sufficient to drive the "squarer" valve.

For valve economy, $V1$ and $V2$ may be the two halves of a 6SN7, the operation being quite as satisfactory as for two 6J5's but, possibly due to cross coupling in the two sections, the position of $R10$ required to start oscillation differs from that at which oscillation ceases, i.e., an oscillation hysteresis effect occurs. This is not a disadvantage but means a little more care is required in the preliminary adjustments.

The setting up procedure is as follows:

1. Disconnect the synch. pulse input at A.
2. Adjust $R10$ until the circuit oscillates; oscillation is apparent when the scanning frequency is independent of the sawtooth generator synch. control over a considerable range of adjustment, i.e., the generator is synchronised to the oscillations.
3. Adjust $R3$ to bring the frequency of oscillation to exactly the fundamental synch. pulse frequency—a steady picture can now be seen on the screen.
4. Adjust $R10$ until oscillation just ceases—detected as in 2 above.
5. Connect up synch. pulse input at A.
6. Adjust frequency control $R3$ to obtain correct phase.

The last adjustment is the continually variable one, all the others are made in the preliminary setting up and should not have to be readjusted.

In conclusion, it may be stated that the above system has now been in operation at Southampton with a standard type receiver having an ordinary dipole and reflector as aerial system and has permitted consistent reception with entertainment value. On the occasions when the signal strength is sufficiently high, the definition is very good and the immunity from ignition noise very noticeable. It may be that the circuit could be used to advantage in a region of high ignition noise even though the signal strength were quite high.

The author would like to thank Mr. G. Q. McColl of the Electronics Department, University College, Southampton for his valuable assistance with the experimental work.

APPENDIX A

Let $R4 = R5 = R$
$C1 = C3 = C$

Now at fr, $(Xc)fr = \dfrac{1}{2\pi frC} = R$

At any other frequency f, $(Xc)f = \dfrac{R \cdot fr}{f}$

$$\text{Feedback ratio } \beta = \frac{\dfrac{R \cdot \dfrac{(-jfr \cdot R)}{f}}{R - \dfrac{jfr \cdot R}{f}}}{R - \dfrac{jfr \cdot R}{f} + \dfrac{R\,\dfrac{(-jfr \cdot R)}{f}}{R - \dfrac{jfr \cdot R}{f}}}$$

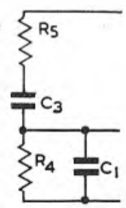
which reduces to $\beta = \dfrac{1}{3 + j(f/fr - fr/f)}$

i.e., $\beta = 1/(3 + jy)$ where $y = (f/fr - fr/f)$
When $f = fr$, $y = 0$, and $\beta = 1/3$.

Let Go = gain with no positive feedback ($Go < 3$)
and G = gain with positive feedback
then $G =$

$$\frac{Go}{1 - Go\beta} = \frac{Go}{1 - Go/(3 + jy)}$$

At "resonance" freq. fr, $Gr = \dfrac{3Go}{3 - Go}$



$$\therefore G \rightarrow Gr\,\frac{(1 + jy/3)}{1 + jy\,Gr/9}$$

At frequencies near fr, $y/3 << 1$

then $G \approx \dfrac{Gr}{1 + jy\,Gr/9}$

i.e., effective $Q \approx Gr/9$
and bandwidth (3db. points) $\approx 9fr/Gr$

Appendix B. Phase shift calculation

From Figs. 3(b) and 3(c) it can be seen that positive synch. pulse is developed approx. 20 μsec. in advance of the leading edge of the line synch. pulse (assuming no phase shift in the selective amplifier of the fundamental frequency component). This corresponds to a phase shift of 72°

$\therefore \tan^{-1} yGr/9 = 72°$
Now $y \approx 2\delta f/fr$ if $\delta f << fr$
whence $\delta f \approx 60$ c/s.

i.e., to introduce the correct phase lag, the tuning frequency must be 10,065 c/s. In practice it was found that some of the phase shift occurred in the amplifier itself due to stray capacitance effects and so not such a large degree of mistuning was necessary. Experiment showed that a mistuning of 60 c/s. leads to a phase shift of 90°.

References

[1] Automatic Frequency and Phase Control of Synchronisation in Television Receivers: K. R. Wendt and G. L. Fredendall. *Proc.I.R.E.* Jan. 1943.
[2] Automatic Frequency Phase Control of Television Sweep Circuits: E. L. Clark. *Proc. I.R.E.*, May 1949. Vol. 37, No. 5.
[3] Vacuum Tube Amplifiers: Chapter 10, by Harold Fleisher: Radiation Laboratory Series, McGraw-Hill.
[4] Variable Audio Frequency Selective Amplifiers, by S. W. Punnet: report to be published in *Journal of Brit. I.R.E.*

The Influence of High-Order Products in Non-Linear Distortion

By D. E. L. SHORTER, B.Sc. (Eng.), A.M.I.E.E.*

THE commonly accepted figures for the maximum allowable non-linear distortion in broadcasting and reproducing systems are based on work carried out many years ago. Since then, new kinds of apparatus producing forms of distortion not covered by the early experiments have come into use, with the result that subjective assessments of non-linear distortion do not always agree with assessments based on measurement. This article describes some recent experiments showing the extent of the error which can result from the common practice of taking the R.M.S. total harmonic distortion of a system as a criterion of linearity.

Comparative tests were carried out on six different systems, which are designated in the Tables by the letters A to F. The distortion introduced by each system varied only slightly with signal amplitude up to a fairly well defined overload point, and the investigation was confined to signals not exceeding this amplitude. In four of the systems (A, B, C, F) the principal distortions were substantially constant over most of the frequency range, and in the other two (D and E) the application of negative feed-back had reduced the distortion over the lower part of the frequency range, without appreciably affecting it for frequencies of 1,000 c/s. and above. With these last two systems, therefore, the standard of quality was set by the distortion of signals above 1,000 c/s., and since most programme material has from time to time a large proportion of its energy around and above this frequency, valid comparisons of *maximum* distortion could still be made between the various systems.

Subjective assessments of non-linear distortion were made on the six systems by listening to a programme on a wide-range coaxial-horn loudspeaker. Much of the test programme consisted of solo pianoforte, the reproduction of which has been found to be seriously impaired by quite a small amount of non-linear distortion. An immediate comparison was made by listening to the programme reproduced through a transmission circuit having very low inherent distortion, the system under investigation being periodically introduced into the chain.

Objective assessments of the non-linearity were made in the first instance by measuring the R.M.S. sum of distortion products produced by the application of either a single tone or of two tones together. It soon became evident, however, that a more detailed analysis was called for, and eventually a single test tone was used with a wave analyser capable of indicating distortion products as small as 0.01 per cent of the fundamental. In addition, the input and output of the system under test were applied to the X and Y plates of a cathode-ray oscilloscope, the phase of the input being adjusted to produce a line. The oscillogram was then examined for sharp bends or kinks, which would indicate the presence of high-order harmonics. The input frequency used for the harmonic distortion measurements was 250 c/s. for systems A, B, and C, and 1,000 c/s. for systems D, E, and F.

* Engineering Research Department, B.B.C.

TABLE I

System	Subjective Classification of Distortion	R.M.S. Sum of Harmonics
D	Bad	3.7%
B	Perceptible	3.3%
C	Just perceptible	2.6%
A	Bad	2.3%
E	Just perceptible	0.62%
F	Not perceptible	0.41%

Table I shows the rough subjective assessment of the distortion of the six systems, alongside the R.M.S. total harmonic measured at a level 8 db. below the overload point. The systems are tabulated in decreasing order of harmonic distortion, and it will be seen that there is poor correlation between the subjective assessment of distortion and the total harmonic measurement.

The oscillograms for systems A and D showed pronounced kinks, suggesting the presence of high-order harmonics, and similar though less prominent discontinuities were displayed by system E. That high-order harmonics are more offensive than low has long been recognised, and it was thought that if allowance for this factor were made, the discrepancies of Table I might be accounted for. The amplitude of each measured harmonic was therefore weighted by a factor proportional to its order relative to the second, a procedure laid down as long ago as 1937 in an R.M.A. specification† for testing broadcast receivers, but, as far as it is known, not widely adopted. According to this procedure, the amplitude of the nth harmonic is multiplied by $n/2$, so that the figure for the 2nd harmonic is unchanged. The result of this weighting is given in Table II, in which the systems

TABLE II

System	Subjective Classification of Distortion	R.M.S. Sum of Harmonics with nth Harmonic Amplitude multiplied by n/2
D	Bad	6.7%
A	Bad	5.1%
B	Perceptible	5.1%
C	Just perceptible	2.8%
E	Just perceptible	1.3%
F	Not perceptible	0.8%

are again presented in decreasing order of the distortion figure. It will be seen that the balance is now partially redressed, but that, contrary to the subjective assessment, systems A and B appear to be equally bad.

A more drastic weighting, in which the amplitude

† "Specification for Testing and Expressing Overall Performance of Radio Broadcast Receivers—Part 2—Acoustic Tests." Published by the Radio Manufacturers' Association. Revised December, 1937. See page 5 under "Distortion Factor."

of the nth harmonic is multiplied by $n^2/4$, was therefore tried. The result with this weighting, which again leaves the 2nd harmonic unaffected, is given in Table III. This time, the order of merit obtained

TABLE III

System	Subjective Classification of Distortion	R.M.S. Sum of Harmonics with nth Harmonic Amplitude multiplied by $n^2/4$
A	Bad	19.4%
D	Bad	16.5%
B	Perceptible	8.6%
E	Just perceptible	4.5%
C	Just perceptible	3.3%
F	Not perceptible	2.2%

from the distortion figures confirms that obtained by subjective tests, and the difference between systems A and B is clearly brought out.

The series of figures resulting from this heavy weighting of harmonic amplitudes converges very slowly if at all, so that the total distortion figure depends to some extent on the number of terms taken. It is therefore desirable to place some limit on the number of weighted harmonics that are included in the total. One method is to include only harmonics up to a certain order. This method was tentatively applied by excluding all harmonics above the 15th, but was ultimately rejected because the results discriminated in favour of the worst system, such as A, which had 0.03 per cent of 25th harmonic. Since heavy weighting brings into prominence some components which were initially too small to be accurately measured, it may be preferable rather to exclude all harmonics having less than a certain amplitude. This criterion was applied to the present data, all terms based on harmonic figures of less than 0.03 per cent being excluded as insufficiently accurate.

The importance of high-order harmonics, even when individually small in energy content, suggests that distortion should be regarded in terms not so much of the separate harmonics as of the complete series or of the composite wave-form that this series represents. The properties of the wave-form which are significant for this purpose remain to be investigated. It may be noted, however, that the product obtained by weighting the distortion wave-form proportionally to the square of the frequency, i.e., by adding 12 db. per octave rise, or differentiating twice, gives a measure of the reciprocal of the radius of curvature of the wave-form, and is therefore related to the sharpness of any corners on it.

The type of input-output characteristic which produces many high-order harmonics from a single-frequency tone will also produce a large number of inter-modulation products when several tones are simultaneously applied. For example, where a single tone of frequency p would produce a single harmonic $2p$, two tones of frequencies p and q would produce two sum and difference frequencies, $p \pm q$. Similarly, a harmonic $3p$ originating from a single tone corresponds to four inter-modulation products ($p \pm 2q$, $2p \pm q$) from two tones, a harmonic $4p$ to six inter-modulation products and so on. Consequently, when the amplitude characteristic of a transmission system has a sharp discontinuity, the number of alien frequencies generated during the transmission of a programme must be very large. It is therefore important that this form of distortion should be recognised whenever it occurs, and also that objective tests should be so framed that they give appropriate numerical expression to its effects. Much speculation is possible as to the best way of achieving this end, but there is as yet insufficient evidence to justify the use of any one formula. The results of the experiments described above show the danger of designing equipment to the usual figure of total harmonic distortion without reference to the type of non-linearity concerned. The time appears ripe therefore for the early experiments on permissible distortion limits to be repeated, taking into account all forms of non-linearity known to occur in modern systems of sound transmission and reproduction.

Sciaky Collector Head

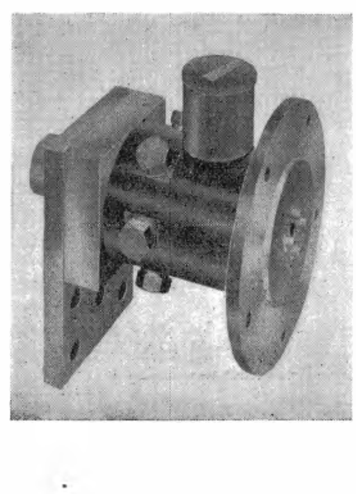
Advantages claimed over the more usual carbon brushes are: lower voltage drop and longer life.

THE function of the Sciaky collector head is to transmit heavy currents from a stationary to a rotating member. This device, a development from a thorough study of the action of transmitting secondary welding currents on electric resistance seam-welding machines, replaces copper carbon brushes such as are used in transmitting current in the electro plate tinning line in the steel industry.

The copper carbon brush is comparatively inefficient because of the relatively high voltage drop under operating conditions, causing considerable heat at contact points and consequently producing rapid wear and high maintenance. The collector head ensures long life with negligible brush drop because of the inherent design and quality of highly conductive materials. The overall efficiency of the head is extremely high since its resistance, as measured from the stationary connexion to the rotating face plate, is less than 10 microhms.

This particular head is rated at 10,000 amperes continuous D.C. current, 12 volts at 400 r.p.m.; however, heads with different ratings can be supplied. The approximate overall dimensions are 10 in. by $11\frac{1}{2}$ in. by $13\frac{1}{2}$ in. with a face plate 10 in. in diameter.

Manufacturers: Sciaky Bros., 4915 West 67th Street, Chicago, Illinois.



The collector head has pure silver contacts, which are fused to copper segments, bearing by spring pressure on the rotating shaft. Up to 2 gallons of water a minute are required for cooling

Ceramic Dielectrics for Electronic Signal Devices

THE relation of systematic variation in composition of ceramic dielectrics to their properties is being investigated in the Porcelain and Pottery Section at the National Bureau of Standards. Such investigations are necessitated by rapid developments in the field of electronics, which have introduced operating conditions too severe for electrical insulating materials commonly used as dielectrics. Paper, for example, is a satisfactory dielectric for some types of capacitors, but at elevated temperatures it chars and thus fails. In contrast, ceramic dielectrics not only withstand temperatures of 500° C. and over, but, more important, ceramic dielectrics having superior properties make possible a reduction in the size of capacitors compared with those made of paper and mica.

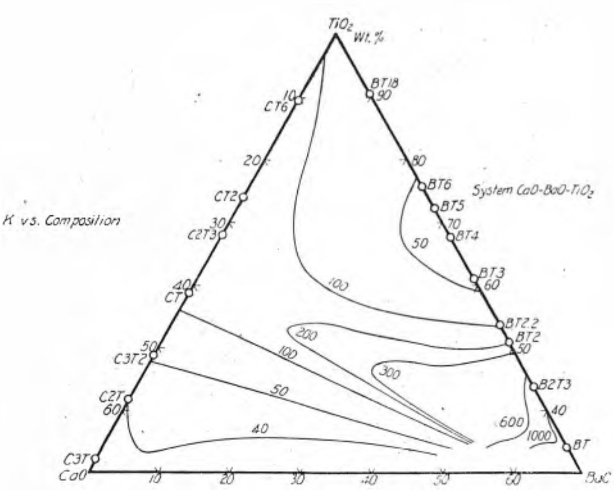
One step in the development of miniature capacitors is the fabrication of ceramics in the form of thin plates comparable in thickness to that of paper and mica. A technique for dry-pressing and firing has been evolved in the production of small plates 0.003 to 0.006 in. thick. The principal change in the conventional dry-pressing procedure involves special treatment of the mixtures of calcines and bonding agent. With reduction in thickness of plates, it is essential that lumps in this mixture be broken up by passing it through a No. 50 sieve. The resulting powder may be distributed in small quantities in the cavity of the hardened steel mould to form a layer of uniform thickness. A pressure of 20,000 lb. per square inch on this layer of powder converts it into a plate a few thousandths of an inch thick. Despite their thinness, these plates are sufficiently strong to be ejected from the mould without cracking and also to be transferred without breakage to a sheet of glass for drying. When dry, the green plates are stacked and stored until needed for firing.

LINEAR THERMAL EXPANSION, SELECTED SPECIMENS OF CERAMIC DIELECTRICS

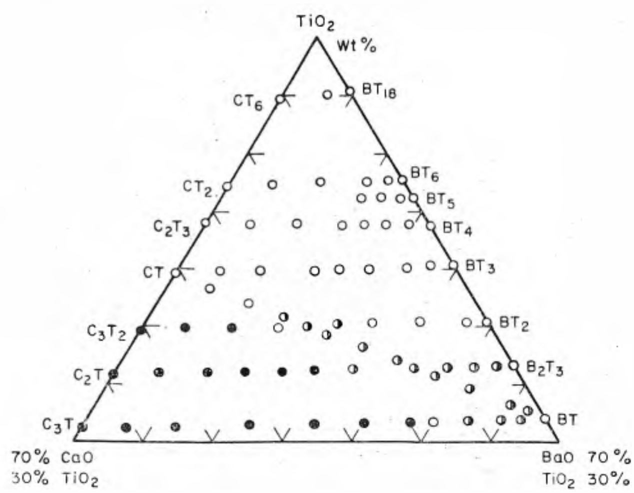
Specimen Designation	Temperature range from 25° C to—						
	100° C	200° C	300° C	400° C	500° C	600° C	700° C
	Per cent	Per cent	Per cent	Per cent	Per cent	Per cent	Per cent
C2T ...	0.07	0.18	0.30	0.42	0.555	0.68	0.82
CT ...	.07	.185	.31	.43	.56	.69	.82
C2T3 ...	.07	.175	.29	.405	.51	.63	.75
CT2 ...	.065	.17	.28	.39	.50	.62	.735
CT6 ...	.06	.15	.25	.35	.45	.555	.655
CB5 ...	.055	.16	.285	.41	.545	.68	.81
BC26 ...	.06	.155	.265	.365	.475	.585	.705
6BC5 ...	.055	.15	.255	.355	.46	.57	.685
BC6 ...	.065	.175	.285	.40	.52	.64	.765
B2C25 ...	.07	.19	.325	.45	.59	.725	.87
BC35 ...	.07	.18	.34	.47	.605	.71	.85

In order to preserve the flatness of the plates during the firing procedure, some changes were necessary in the usual arrangement of the plates on the refractory supports. Excessive warpage invariably occurs if the plates are distributed in a single layer on the plane surface of the refractory. Flat plates are obtained, however, when they are stacked on such a refractory, and the stacks are weighted with a refractory disk. To prevent adherence at high temperatures, the stacked plates are separated from each other by thin layers of air-floated zirconium dioxide. Firing for 1 hour at 1,445° C. results in matured plates, as indicated by less than 0.1 per cent of water absorption.

It is desirable that the capacitance of capacitors fabricated from ceramic dielectrics should not be affected by variations in temperature over a wide range, such as −60° to 200° C. Experimental



Systematic variation of composition produced calcium-barium titanate dielectrics having widely different electrical properties. The range of values for dielectric constant, K, is shown on the ternary diagram for the system CaO-BaO-TiO_2. B=BaO, C=CaO, and T=TiO_2



The various calcium-barium titanate dielectrics have been plotted on the ternary diagram to show their electrical properties (K, dielectric constant, and Q, reciprocal of the power factor). Compositions shown as black full circles decrease in K and Q with time; half-circles decrease in K and increase in Q with time; circles are stable capacitiors assembled so far at this Bureau contain plates, 0.75 by 0.75 by 0.005 in., which are composed of a series of barium-strontium titanates with peak values of 12,000 to 18,000 dielectric constant (K) at 5-degree C. intervals from $-60°$ to $120°$ C. Such assemblies have a capacitance of 5 to 15 microfarads per cubic inch with a variation in capacitance of 11 per cent for changes in temperature from $-60°$ to $100°$ C. For comparison, the dielectric constant of paper and mica is under 10, and the volume they require in capacitors for equivalent duty is correspondingly larger. At 200° C., however, the capacitance for the ceramic capacitors decreases to one-third the maximum value and the direct-current leakage becomes excessive.

In some instances, designers of electronic equipment have called for the fabrication of a capacitor of given capacitance, temperature coefficient of capacitance, shape, and size. This in turn required specific properties on the part of the ceramic dielectric. Several types of single plate and tubular capacitors, fabricated at this Bureau, have met the designers' specifications. The selection of some of these dielectrics was based upon data recently obtained from the Bureau's investigation of the properties of calcium-barium titanate dielectrics.*

Ceramic dielectrics are applicable to electronic signal devices not only for the armed services, but also for the commercial production of radio, radar and television sets, and hearing aids. Other investigators are examining the uses of ceramic dielectrics in research problems that involve very high voltage, X-rays, and instantaneous photography.

—*National Bureau of Standards Bulletin*, December, 1949.

* For further technical details, see "Properties of calcium-barium titanate dielectrics," by E. N. Bunting, G. R. Shelton, and A. S. Creamer, *J. Research*, NBS 43, 237 (1949) RP2025.

A Television Sound Rejector

By C. H. BANTHORPE*

ONE of the problems which has to be overcome in a vision receiver is that of sound break through, and this is particularly difficult in a T.R.F. receiver, due to the small percentage difference between sound and vision frequencies. The subject has been covered fairly completely by at least two writers[1,2] but there appears to be no reference to the circuit known as the "Bridged-Tee" filter, although it has certain advantages for this purpose.

The circuit has been used for many years in a number of ways, e.g., for the measurement of "Q," incremental inductance, frequency, impedance, and for distortion measurements[3] and has the property of providing infinite attenuation at one frequency. In America it is finding favour as a filter in television, but is little used here.

One circuit is shown in Fig. 1, *L*, *C* and *C* forming a delta. If this is transferred into the equivalent tee, we get Fig. 2, *Z3* being negative. If a resistor of this value is put in series with *Z3* the shunt arm will become a short circuit at one frequency and no energy can flow through the filter. At other frequencies the value of *Z3* changes and as *R3* remains fixed, the sum of *R3* and *Z3* is no longer zero, and the filter passes energy.

If the dynamic impedance ($Q\omega L$) of *L.R CC.* is Zo at the reject frequency, then the value of *R3* must be Zo/4 for best rejection.

Using standard components, an attenuation of >300:1 has been maintained in production and by carefully selecting *R3*, very much higher ratios can readily be obtained.

Two practical circuits used in production by the writer are shown in Figs. 3 and 4; the first is useful in the aerial circuit of a vision only receiver, and the second is used in a circuit where the first two valves are used for sound and vision amplification.

It has been found best to use silver mica capacitors for the traps and the resistors are of the high stability cracked carbon type made by Dubilier, otherwise trouble may be experienced due to thermal drift.

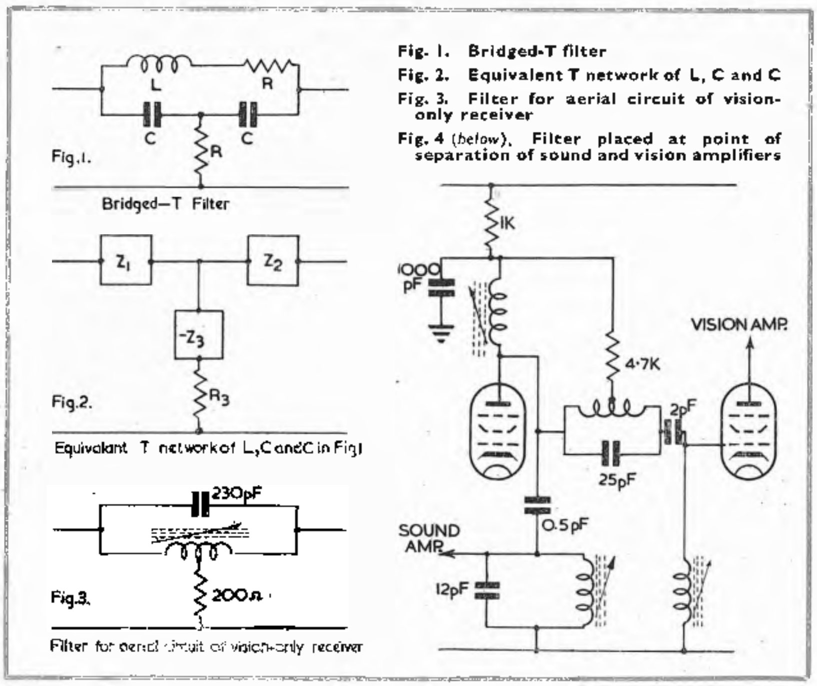


Fig. 1. Bridged-T filter
Fig. 2. Equivalent T network of L, C and C
Fig. 3. Filter for aerial circuit of vision-only receiver
Fig. 4 (*below*). Filter placed at point of separation of sound and vision amplifiers

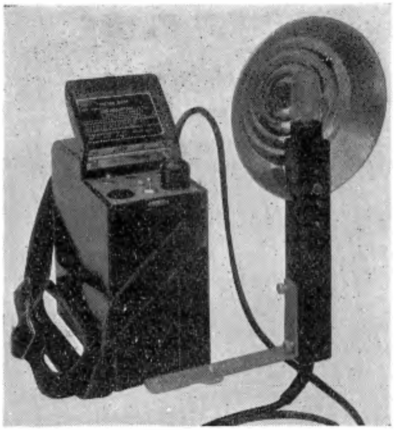
References

[1] Dr. K. R. Sturley. *Journal of the Television Society*, March 1945, p. 100.
[2] W. T. Cocking. *Wireless World*, Dec. 1946, p. 417.
[3] F. E. Terman. *Radio Engineer's Handbook*, pp. 910, 918, 919, 944, 953.

* Central Equipment Ltd.

Electronic Equipment

A monthly record of British electronic apparatus, components and accessories, compiled from information supplied by the manufacturers.



Portable Electronic Photo-Flash

(Illustrated above)

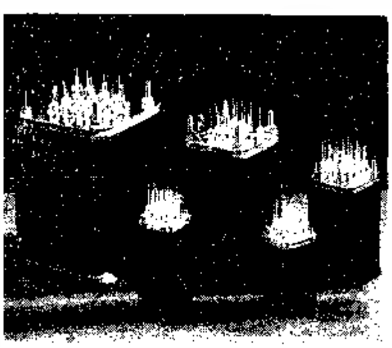
ELECTRONIC photo-flash equipment operated by charging a capacitor to a high voltage and then discharging the stored energy through a Xenon-filled tube. The resultant visible flash is of high intensity, short duration, and is particularly suitable for photographic purposes.

The capacitor discharge is effected from a pulse transformer, the low voltage primary of which is controlled by a set of contacts fitted to the camera shutter. Precise synchronisation of the flash to the camera shutter is thereby obtained.

In the "Courtenay Master" the tube operates at a loading of 100 joules, a guaranteed minimum of 10,000 flashes being obtained at this loading. The light spectrum emitted is similar to noon daylight in quality and is quite suitable for colour film. Exposures on HP3 or Super XX at distances of 10 feet with apertures of *f*16 are normal with usual developing techniques.

The "Courtenay Master" is of small size, low weight, robust build, contains a built-in accumulator charger, operates from both the internal accumulator or A.C. mains and, with its high light output, is in all ways suitable for portable or studio use.

Clive Courtenay & Co., Ltd.,
5, Horsham Road,
Dorking,
Surrey.



Miniaturised Hermetically Sealed Transformers and Chokes

(Illustrated bottom left)

THE Ferranti range of miniature hermetically sealed mains transformers, audio transformers and chokes is specially suited for use in equipments in which full tropicalisation, miniaturisation and reliability are important.

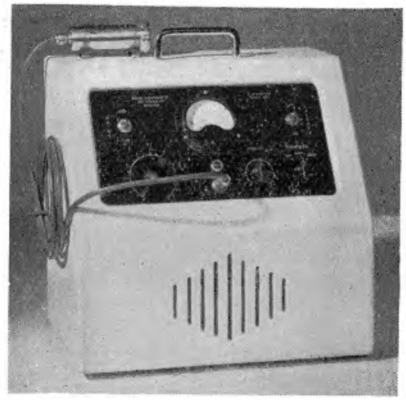
Monel metal is used for housing the transformers in cans because of its anti-corrosive properties, which make it eminently suitable for extreme tropical conditions. The cans have ceramic bush-type terminals and are spray-painted black to give maximum heat dissipation. The transformers are designed to customers' specific requirements to operate at frequencies from 400 to 2,800 cycles.

Designed to conform to the requirements of Inter-Service Specification R.C.S. 1000, the transformers are robustly constructed, and can withstand a shock test of 60g.

The coils are paper interleaved and wound on moulded bobbins. The whole construction is light and compact, with coil and core unit rigidly clamped inside the can. All transformers are subjected to the sealing test specified in Inter-Service Specification R.C.S. 214 before they are filled with petroleum jelly or oil, and are sealed above their operating temperature to allow for the safe expansion of the impregnant under operating conditions.

An outstanding feature of the range is the large number of terminals that can be provided. Ceramic bush-type terminals are used, fitted into dimpled holes in the lid of the can and soldered internally. A tapping panel can be mounted over the ceramic bushes to provide for the selection of individual tappings. Another feature is the assembly of a can with terminals emerging at both ends. This not only makes provision for an increased number of terminals, but also makes it possible for a particular terminal to be located at the end most convenient for external connexion. In certain circumstances it is also possible to house two transformers in one can.

Ferranti Ltd.,
Hollinwood, Lancs.



The Panax Rate-of-Count G-M Monitor

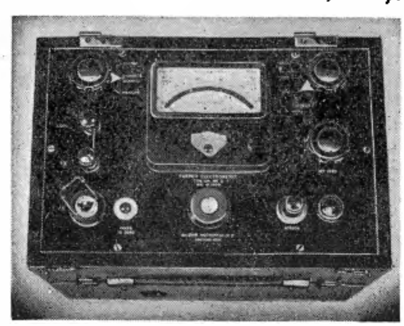
(Illustrated above)

THE Model 40A is a portable instrument designed for checking contamination and locating sources of activity in laboratories using radioactive substances. It can also be used effectively for general assay work.

A speaker gives audible indication of activity and a visual rate-meter gives ranges of 0-1,000 ; 0-10,000 ; and 0-50,000 counts per minute. An aluminium window tube of 6 mgm. per cm^2. is supplied in a carrier as standard, but mica window tubes can be fitted if required.

The instrument operates on A.C. mains voltages and has a consumption of 50 watts. It is finished in a heat-resisting plastic and weighs approx. 24 lb.

Panax Equipment, Ltd.,
347, London Road,
Mitcham, Surrey.



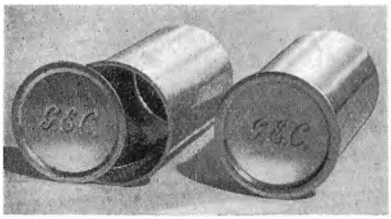
Cold Pressure Welding

(Illustrated below)

THE cold pressure welding process, developed by the Research Laboratories of The General Electric Co., Ltd., is now being used for the sealing of aluminium containers, in which the G.E.C. are sending electronic tubes overseas. It is possible that this application of the cold welding technique may be the forerunner of radical changes in packing methods for the products of many industries.

The flanged container shown is formed from an aluminium slug by impact extrusion. Then, after the welding faces of the flange and the disk lid have been cleaned, they are welded in one quick operation by the well known method. The weld is airtight, and the container will protect the contents from changes in atmospheric pressure and humidity.

The General Electric Co. Ltd.,
Magnet House,
Kingsway,
London, W.C.2.



Farmer Electrometer Voltmeter

(Illustrated below left)

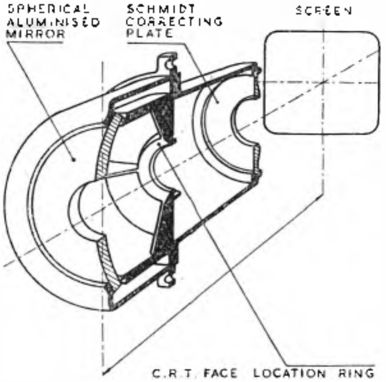
THE Baldwin-Farmer Electrometer Voltmeter has a high degree of robustness and stability of calibration—+5 per cent or —10 per cent deviation from the nominal mains voltage will affect the zero and calibration by less than 0.2 per cent of full scale : in addition to its use as a very sensitive electrometer it is also calibrated for direct reading in volts, and has BS 89 first-grade accuracy. The indicating instrument, being a pivoted moving-coil microammeter, is rapid in response. The circuit is such that the capacitance is very low—less than 1pF.

An electrometer-valve is used and the resulting insulation is approximately 10^{16} ohms, the internal high-insulation points being sealed and a desiccated compartment ensuring freedom from leakage and variability of insulation due to atmospheric changes. Zero drift is less than 0.5 per cent of full scale deflexion per hour, after five minutes warming up on 50V range, or less on higher ranges.

The instrument has a range of 0-50-100-500 volts, with insulation to withstand up to 1,000 volts between X terminals and the earthed panel or mains earth.

As an amplifier the electrometer voltmeter provides sufficient output to operate recorders, controllers or remote indicators, such as in a lecture theatre.

Baldwin Instrument Co. Ltd.,
Dartford, Kent.



Television Projection System

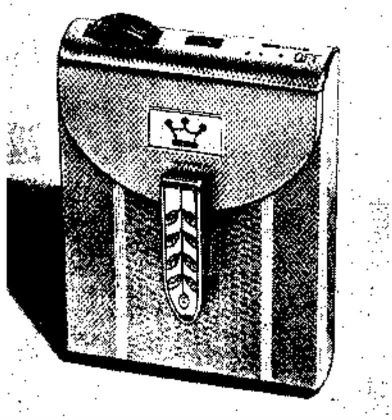
A new method of producing the components for an all-glass Schmidt optical projection system has cut manufacturing costs considerably. A total projection distance of 36 inches is required to give a 15 in. by 12 in. picture ; if necessary this can be attained by the use of a surface aluminised mirror. When used to throw a picture on to an I.C.I. processed perspex plate, ample picture brilliance is produced by a $2\frac{1}{2}$ inch diameter C.R.T. by virtue of the directional properties of the screen. The viewing area is limited by this special screen to within a 15 degree arc in the vertical plane, and a 90 degree arc in the horizontal plane. An important advantage of this screen is that it is not affected by room lighting to the same extent as a direct-viewing C.R.T.

Imperial Chemical Industries Ltd.,
Black Fan Road,
Welwyn Garden City, Herts.

Optical Works, Ltd.,
32, The Mall,
Ealing, W.5.

Micropak Hearing Aid

THE Bonochord Micropak is a new one-piece hearing aid housed in an engraved case of jewellers' gold to give the appearance of a miniature cigarette or vanity case. Size is $3\frac{1}{8}$ in. by $2\frac{1}{8}$ in. by $\frac{5}{8}$ in., and total weight, with batteries, is 5 ounces. The whole instrument fits into the palm of an average size hand.

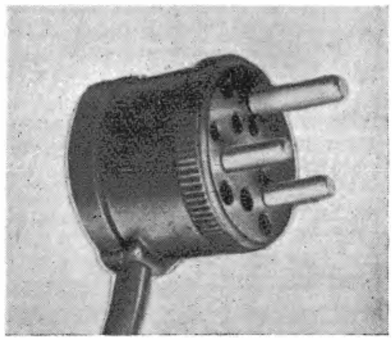


The microphone is completely concealed behind a decorative coronet motif, integral with the external case. There are three stages of amplification, with pentode output, crystal wide-range microphone, automatic volume limiting, and an ultra low drain circuit reducing the H.T. drain to 0.4 mA, and giving a serviceable life of approximately 350 hours from the Ever-Ready B121 15V sub-miniature H.T. Low tension supply is by standard Ever Ready D14 "pen" cell giving 8 hours life, or by the Vidor Kalium U7 "pen" cell giving 30-35 hours. Running costs average under one halfpenny per hour of use.

Miniature insert receivers for air conduction, either crystal or magnetic, are provided, and a bone conduction receiver can be used. Maximum air to air gain is 50 db.

Effective frequency range is from 250 to 5,000 cycles per second, and the response curve is reasonably flat throughout. All valves are plug in type, and the chassis can be changed in one minute by the removal of only two screws.

Bonochord Limited,
48, Wigmore Street,
London, W.1.



Six-way Plug

MESSRS. Crater Products, Ltd., have produced a 6-way plug which is covered by world patent. Although very little larger than the standard 15-amp. 3-pin plug, it contains, in the body, six complete different sets of pins to fit sockets ranging from 2-pin 2-amp. up to 3-pin 15-amp. The pins are threaded on their back ends and by manipulation of a six-position shutter different combinations drop out as required ; it is then only necessary to give each pin a twist until it is locked.

To ensure that the different combinations can be utilised to the fullest extent, the smaller 2-pin types, i.e. 2-pin 2-amp. and 2-pin 5-amp. are arranged round the edge of the plug, so making it possible to insert it into sockets which are near to the floor. A cable grip is employed in the top cap and the whole plug is made in high shock resisting material.

Crater Products, Ltd.,
The Lye, St. Johns,
Woking, Surrey.

BOOK

Laboratory and Workshop Notes

Compiled and edited by Ruth Lang. 1949. Edward Arnold & Co. 272 pp. 195 figs. Price 21s.

READERS of the *Journal of Scientific Instruments*, and others concerned in the design or construction of laboratory apparatus, will find these notes selected from the *Journal* interesting, ingenious and stimulating. Dr. Lang has chosen 181 short articles which have appeared in the *Journal* between 1925 and 1948 (under the title heading), and has classified them in eight sections; workshop tools and devices; clamps, supports and agitators; soldering, brazing and welding; glass manipulation and silvering; vacuum and pressure technique; electrical and magnetic devices; optical techniques, and devices for liquids and gases.

The nearest approach to electronics is made by an author who points out that a capacitor in series with the A.C. mains is the cheapest and simplest method of providing magnetic relays. However, this book should certainly contribute to the improvement of technical methods in all laboratories which have not committed themselves exclusively to electronic techniques as a solution to every problem.

There is little to criticise in the selection. The articles are nearly all of the high standard expected of the *Journal of Scientific Instruments*: and it is only rarely that the editor has allowed duplication to occur. There is a good index; and the reminiscences of Sir Charles Boys make amusing, as well as interesting, reading. The book perhaps suffers a little from its miscellaneous nature, but it suffers more from its price. Since the whole of the subject matter has been freely available to the publishers, the price seems rather high; and it is to be hoped that the further selection which is to be made from the *Journal's* Notes will be published at a price more within the range of the amateur engineer, to whom this type of book has a considerable appeal.

C. J. DICKINSON.

Wireless Simply Explained

By R. W. Hallows. 255 pp. 111 figs. 1st Edition, 1949. Chapman & Hall. Price 10s. 6d.

THE aim of this book is to bridge "the gulf between the book of the 'wireless without tears' kind and the large textbook." According to the author, little has been done to provide such a bridge; however, little appears here which cannot be found in many other admirable books. "Wireless Simply Explained" can only be regarded as another elementary textbook, designed to appeal to the complete beginner.

Starting from the assumption that the reader knows nothing of "Wireless," the author leads him from Ohm's Law applied first to D.C. and then to A.C. circuits, through discussions on series and parallel resonance, the valve as a detector, amplifier, and oscillator, to the complete superheterodyne. A welcome chapter headed "Things to Come" briefly outlines some types of modulation other than amplitude modulation. Finally, four appendices deal with "Power Factor," "A Mathematical Shorthand" (an explanation of the laws of indices), "Power Factor and Tuned Circuits" and "The New Letter Symbols for Valves."

Inevitably the weakest chapters of this book are the introductory ones, where the author is concerned with the very rudiments of electrical knowledge. Once the difficulty of explaining what conductors or insulators or other fundamental concepts are, is passed, the book proceeds in a brisk though rather "chatty" way. Some of the methods are unorthodox (e.g., discussing the complete superheterodyne receiver before discussing various amplifier circuits), and the diagram representing frequency changing as amplitude modulation at the intermediate frequency of the oscillator by the signal will undoubtedly cause many unnecessary difficulties.

The binding and the printing are, as with all the publications from this house, excellent, though two minor mistakes (pages 73 and 184) seem to have escaped the proof reader. As a textbook for an introductory class where the lecturer can fill in the gaps, this book should be valuable, but for serious individual study its wide scope has rendered the general treatment superficial.

K. G. LOCKYER.

Microwaves and Radar Electronics

By E. C. Pollard and J. M. Sturtevant. 426 pp. Chapman & Hall, 1948. Price 30s.

A PHYSICAL research laboratory has, in the years since the war, changed profoundly in the nature

REVIEWS

and complexity of the apparatus which it makes and uses. Much of the equipment and many of the techniques are those developed for radar purposes. To give an introduction to developments in the radar field for those whose interests lie in their applications to physics, chemistry and other sciences is the primary aim of this book.

The range covered is wide, as a list of the chapters will show. These are: Electromagnetic Fields and Microwaves, Coaxial Lines Waveguides and Cavities, Production of Microwaves, Microwave Technique, Pulse Circuits, Cathode Ray Tube Indicators, Tuned Amplifiers, Amplification of Very Weak Signals, Servomechanisms and Computors, Miscellaneous Circuits, Radar and its Accessories, Microwave Communications, Microwaves in Physical Research.

In a field so wide, the choice of material is difficult, but the authors have succeeded admirably in selecting representative topics and in describing principles and salient features without extraneous detail. The reader is assumed to have a reasonable knowledge of valves and circuits, and sufficient mathematics to understand the elements of electromagnetic theory.

The book, although it has a title suggesting a content similar to a number of recent books, differs considerably in its treatment. It does not attempt to be a logical exposition of radar or microwave electronics for the engineering student, but rather to be a guide to those who have to use microwave apparatus and require sufficient background to give them an intelligent understanding of it. The inclusion of a paragraph describing the pedestrian task of "getting a klystron to oscillate at the right frequency" is a good example of the authors' appreciation of their task, although this must not be taken to indicate the general level of the treatment.

There are sections, however, which seem superfluous, for example, listed details of a number of production radar equipments. On the other hand one wishes that more space could have been given to matters important in physical applications, such as the accuracy of power measurement, measurement of receiver sensitivity, and the effect of dielectrics in waveguides and cavities. The scope also is narrowed by being largely confined to the techniques used at the Radiation Laboratory, M.I.T. Thus it is natural but unfortunate that about half the references are to Radiation Laboratory reports, which will be inaccessible to most readers in this country. The last chapter is a good introduction to the physical uses of microwaves, and includes the elements of microwave spectroscopy, radiation measurements and cavity accelerators (strangely enough without any mention of loaded waveguide types).

The errors noted were trivial, except perhaps on p. 91 where the constant 2.405 in equation (4.1) is that appropriate to the TM_{01} mode, not to the TE_{11} mode as stated.

The book is attractively produced, with wide margins and good illustrations. Altogether it is a useful book, easy to read, and is sufficiently complete also to provide a good introduction for the engineering student.

H. R. L. LAMONT.

Aerials for Metre and Decimetre Wavelengths

By R. A. Smith. 218 pp. 125 figs. 1949. Cambridge University Press. Price 18s. net.

THE appearance of this volume in the *Modern Radio Technique* series will be welcomed by all those who have been waiting for a connected account of the war-time developments in short-wave aerial technique. The present volume deals only with aerials of a more or less conventional type, leaving the developments of the centimetre wave-band, and in particular the use of wave-guides and horns, to another volume of the series. It does, however, include a treatment of slot aerials (as alternatives to dipoles), and refers to one case where a paraboloid is used to produce a narrow beam of radiation. The bulk of the book refers to aerials for wave-lengths of more than one metre; in spite of the inclusion of the word "decimetre" in the title, only eight pages refer specifically to aerials for wavelengths between 10 and 100 cm., although of course much of the other material is applicable within this frequency range.

While the book is primarily a practical treatise, over one-third of its pages is devoted to the theory of dipole radiators and assemblies thereof, including the use of reflectors and "directors." However, the theory is not unduly difficult, and can be read with profit by most practical workers in the field. The remainder of the book describes applications in some detail. One might make the criticism that these are very largely drawn from the work of one war-time establishment; nevertheless, as a selection of illustrations of the various types of aerial they are admirable. Generally a knowledge of pre-war short-wave techniques is assumed, and the problems of the transmission lines associated with aerials are much less fully treated than is the theory of the radiators themselves. Had the emphasis been rather more on shorter wavelengths, this subject might perhaps have deserved a fuller treatment, since the departures from prewar practice naturally increase when the wavelength becomes shorter. The problems of switching on the feeders of multi-element directive arrays is also treated rather briefly. We note also that the author describes aerial impedance in terms of separate graphs of the resistive and reactive components against frequency, whereas at shorter wavelengths it is becoming increasingly common to plot a curve showing the frequency variation on the complex impedance plane, or on a circle diagram chart.

In spite of these criticisms, the book is most valuable, and should be widely read. It appears singularly free from major errors, although a few minor blemishes have been noticed, such as the failure to state whether front-to-back ratios are quoted as field-strength or power ratios, and the substitution of "vertical" for "horizontal" at the foot of p. 199. The fact that the curve of Fig. 98 crosses the axis also excites curiosity, but is not explained. The publishers, too, have done their work well and produced an attractive volume; one could only wish that it could have appeared sooner, for a delay of 2½ years from the date of the preface to that of publication appears somewhat protracted.

C. H. WESTCOTT.

BOOK REVIEWS (Continued)

Basic Electrical Principles

By M. G. Suffern. 430 pp. 275 figs. 1949. McGraw-Hill, Ltd. Price 19s. 6d.

THIS American textbook "is designed to give electrical trades students training in the fundamentals on a vocational education level."

The introductory chapter deals with the electron theory and electrostatics. This is followed by a chapter on batteries, then one on the electric circuit and the applications of Ohm's Law. The next four chapters are on electromagnetism, alternating currents, inductance, and capacitance. Then, the effects of inductance and capacitance, the electric generator, the electric motor, rectification of alternating current, and the electric meter are considered.

The simplification in presentation leads to some overstatements, as "All A.C. meters will operate on direct current as well as alternating current" (p. 375). The omission of vector representation throughout the book places a limitation on the treatment of any given topic. In spite of this, however, the author deals with a variety of electrical plant. Thus the chapter on "the electric generator" mentions single and polyphase alternators in addition to all the common types of D.C. generator. The chapter on the electric motor deals with D.C. motors, the rotary "convertor," the dynamotor, as well as induction, synchronous, and repulsion types.

The book on the whole is well written though some errors are apparent. Thus on p. 168 ". . . since the self inductance is initially very high" gives the impression that self inductance is a variable quantity. The variation in effective inductance due to an iron core may be intended but this is not clear. The caption to Fig. 138b implies an exception to Newton's First Law of Motion. (This particular exception used to be applied to Irish pigs.) Fig. 138b shows a boy running downhill with a snowball and the caption reads "As the inertia is overcome by some force, the object tends to move in a direction opposite to the direction of the applied force." If anyone is concerned about the number of electrons in a coulomb, p. 193 gives the figure of six hundred and twenty-eight million million in place of the usual 6.28 by 10^{18}.

The book contains a large number of diagrams and illustrations, a useful feature of which is that the corresponding symbols and circuit diagrams are also shown.

As I am sure the author would agree, the book is suitable only for elementary students who need to have simple ideas about a wide range of electrical equipment.

WILLIAM FRASER.

Magnetic Recording

By S. J. Begun. 242 pp. 146 figs. 1949. Murray Hill Books Inc., New York, U.S.A. Price 5 dollars.

DR. S. J. BEGUN, vice president and chief engineer of the Brush Development Company, is well-known for his early work on magnetic recording, particularly with the wire medium, and this new volume is the first comprehensive survey of the field to be published in book form.

The first chapter deals with the historical aspects of magnetic recording, concluding with a useful list of references and Patent specifications. Chapters 3 and 4 are devoted to fundamentals of magnetism and the theory of magnetic recording, and the author is to be congratulated on his lucid explanations of the complexities of magnetic phenomena. The difficult topic of alternating-current biasing is admirably treated. The various magnetic recording media from the original carbon steel wire to the latest powder-coated tapes and other components of the magnetic recording system are discussed in Chapter 5.

Twenty-six types of commercial magnetic tape, wire and disk recorders are described and illustrated in Chapter 6, accompanied by a number of circuit diagrams and excellent line-drawings showing the operation of the various drive mechanisms for recording, playback and rewind. Chapter 7 touches on a number of uses of magnetic recording, ranging from time delay and artificial reverberation to applications as "memories" in electrical computing devices and in the motion picture industry.

The penultimate chapter is a valuable study of instruments and equipment, e.g., automatic *B-H* curve tracers, coercive-force meters, loop testers, necessary to evaluate the performance of magnetic record materials and systems. The book concludes with some prophetic views on the possible supercession of the disk gramophone by magnetic methods. Each chapter ends with a helpful bibliography. A glossary and index are included.

Bearing in mind that this book was written in 1948, it is perhaps churlish to criticise the omission of references or descriptions of British magnetic recording equipment, such as the E.M.I. and Gaumont-Kalee, and only minor printing errors have been detected, e.g., p.133 Pugsley is misspelt and on pp. 89 and 91 the running chapter heads are incorrect. This readable, substantially non-mathematical text can be thoroughly recommended as a basic reference work.

DONALD W. ALDOUS.

The Copenhagen Plan

In accordance with the Copenhagen Wavelength Plan the B.B.C. changed the wavelengths of nearly all its programmes on March 15.

The re-arrangement of broadcasting frequencies within the "European Zone" should help to establish better reception in Europe generally after the confusion due to deviations from the Lucerne Plan, which was implemented in 1934. There were 26 countries which signed the Plan, but to a certain extent the success of the Plan will depend on the eight non-signatories. The Plan does not immediately benefit this country, but European broadcasting would have drifted into a state of chaos without it.

The alterations, which make up one of the largest engineering projects that the B.B.C. has tackled since 1939, will cost between £200,000 and £300,000. More than half of this sum, however, is the cost of building a new high power transmitter at Daventry for the Third Programme. Most of the B.B.C. medium wave stations are substantially increasing their power.

Station	Programme	Copenhagen Plan: Frequency kilocycles per second	Copenhagen Plan: Approximate wave-length metres
Home Service			
Moorside Edge	North of England Home Service	692	434
Burghead, Redmoss, Westerglen	Scottish Home Service	809	371
Penmon, Washford, Wrexham	Welsh Home Service	881	341
Brookmans Park	London Home Service	908	330
Start Point ...	West of England Home Service	1,052	285
Droitwich, Norwich	Midland Home Service	1,086	276
Lisnagarvey, Londonderry, Stagshaw	Northern Ireland and North of England Home Service	1,151	261
Bartley, Clevedon	West of England Home Service	1,457	206
Light Programme			
Droitwich ...	Light Programme	200	1,500
Brookmans Park, Burghead, Lisnagarvey, Londonderry	Light Programme	1,214	274
Light Programme—*continued*			
Moorside Edge, Plymouth, Redmoss, Redruth, Newcastle, Westerglen	Light Programme	1,214	247
Third Programme			
Daventry, Edinburgh, Glasgow, Newcastle, Redmoss	Third Programme	647	464
Belfast, Bournemouth, Brighton, Cardiff, Dundee, Exeter, Fareham, Hull, Leeds, Liverpool, Manchester, Middlesbrough (Stockton), Plymouth, Preston, Redruth, Sheffield	Third Programme	1,546	194

ORANGE SORTING BY X-RAYS

(Abstracted from the February, 1950 issue of *Electronics*)

WHEN an orange is frost-bitten, certain physical and chemical changes occur inside the orange which decrease its value to the consumer. A fully automatic X-ray inspection system has recently been developed which checks the internal condition of each orange and routes it into one of six classifications, depending on internal condition, at a rate of 10 a second. The new equipment has three times the capacity of previous equipments of this nature.

Juice cells in frost-bitten oranges break down, allowing the juice to collect in the fruit's centre. The juice is then reabsorbed by the tree, leaving a light, juiceless fruit.

The equipment has two sets of X-ray-sensitive cells. One views individual oranges as they pass down two rows of endless belts. The amount of X-radiation each orange passes is recorded electronically in a matter of 1/100 second. Another radiation current is simultaneously recorded as an X-ray tube transmits through a predetermined standard, which is equivalent to the X-ray capacity of an average-size normal orange. These two signals are compared, and their difference is used to actuate the classification switches. The classifications are stored in a memory system which initiates a tripping mechanism that catapults the orange into the proper chute. The equipment, which was developed by H. D. Roop of the Automatic X-ray Corp. of Los Angeles, is designed to compensate for outsize oranges.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Television Planning

DEAR SIR,—On seeing a map of the proposed B.B.C. Television Chain, I could not resist comparing it with a forecast which I was concerned in preparing in 1943. (Published in ELECTRONIC ENGINEERING, Vol. 16. No. 191, January, 1944.) The comparison in tabular form is given below.

In 1943 I under-estimated the possibilities of high masts (e.g., Sutton Coldfield aerials 1,300 ft. above sea level) and the Sutton Coldfield Station is a better proposition for the Midlands proper than my two stations at (a) and (b).

I hope the same will be true of the replacement of my separate Yorkshire and Lancashire stations (c) and (d) by Holme Moss. I have not checked the profiles, but with 1943 ideas of mast height I did not think a station on the Pennine plateau could serve both Sheffield and Manchester, which are fairly well "under" the high ground at the southern end of the Pennines (ground which is mostly higher than Holme Moss).

We are fairly well agreed on South Coast (Southampton), Newcastle, Belfast and Edinburgh-Glasgow. But why a station at Aberdeen? Is this government policy to tempt workers out of overcrowded Glasgow into the under-populated Highlands, or a tribute to the strength of Scottish Nationalism? On the other hand, must East Anglia always be the Cinderella of British radio services? For years we have heard of East Anglians being compelled to rely on the Continent for most of their medium-wave broadcasting: will they now rely on Paris and perhaps a future Brussels and Amsterdam for television?

In contrast with Scotland, the Welsh Nationalists fare badly in the B.B.C. plan. I gave South Wales its own transmitter on the technical grounds that I doubted whether the population in the valleys behind Swansea (e.g., Merthyr Tydfil) could be given a service from a transmitter on the far side of the Bristol Channel, but it also provided a viewing-point for Welsh culture. Even if the field-strength is sufficient, what will the Welsh say to being given all-English programmes from London studio via Somerset transmitter?

I hope, therefore, that the Television Advisory Committee will consider the potential consumers' probable views before a purely technical plan is put into effect. (Were low-power television stations put at Aberdeen and Plymouth but not East Anglia because those are the places where the B.B.C. now has medium-wave transmitters?) If we cannot afford permanent studios at all transmitters cannot the B.B.C. have a mobile studio—a sort of out-size O.B. unit with a certain amount of studio space—to be sent on tour so that we could have in turn a "Scottish month," "Midlands month," etc.? Television must not be a "London steam-roller," one more avalanche to bury the remains of regional culture and activities.

D. A. BELL,
Birmingham 17.

1943 Forecast	1949 B.B.C. Plan
Existing London Station, Alexandra Palace	
4 'Midlands' stations:	2 'Midlands' stations:
(a) Birmingham (for S.W. Midlands)	(a) Sutton Coldfield (for all Midlands S. of Pennines)
(b) Melton Mowbray (for Leicester, Nottingham)	(b) Holme Moss (for both Yorkshire and Lancashire)
(c) Huddersfield (for Yorkshire)	
(d) Macclesfield (for Lancashire and Cheshire)	
N.E. Station near Durham	N.E. Station near Newcastle
1 Scottish Station:	2 Scottish Stations:
10 m. W. of Falkirk	(a) 10 m. S. of Falkirk
	(b) Aberdeen
Belfast	Belfast
Isle of Wight	Southampton
2 S.W. Stations:	2 S.W. Stations:
(a) Bristol	(a) Bridgwater (?) (for Somerset and Bristol Channel)
(b) Swansea	(b) Plymouth
East Anglia	

The Design of R-C Oscillator Phase Shifting Networks

DEAR SIR,—Mr. W. R. Hinton in his article "The Design of R-C Oscillator Phase Shifting Networks" states that "there is little to choose between R-C and C-R sections from the point of view of harmonic distortion" and that "the use of more than three sections is seldom justified."

In my experience, the shunt C network inherently gives far less distortion than the shunt R equivalent.

A network consisting of three or four shunt C cells in tandem is a form of low pass filter, and conversely, the shunt R counterpart is a form of high pass filter. If the same input, fundamental plus harmonic, is applied to each, then quite obviously the output from the shunt C network will be much purer than that from the shunt R network. The shunt C network output will have less percentage harmonic than the input signal, whereas the output from the shunt R network will actually have greater percentage harmonic content than is present in the original input.

In a practical R-C oscillator circuit, if output is taken from the output of the network, and it is arranged that the non-linear element limiting oscillatory amplitude results in distortion in the signal applied to the input of the network, then it is obvious from the above that the shunt C version will be far superior to the shunt R counterpart. In a shunt C oscillator developed by the writer several years ago, the harmonic content was -55 db on the fundamental, although the amplifier gain was sufficiently in excess of the network attenuation to permit of level constancy within ± 0.2 db over a wide frequency range. This harmonic performance, with the same degree of over-driving, could not have been approached by the shunt R version.

Although the superiority of the shunt C network is dependent upon taking output from its output terminals, and arranging that the non-linear parameter limiting oscillatory amplitude does not debase the output waveform, nevertheless these requisite conditions are so obvious that it is surprising that the low harmonic potential of the shunt C network has not been more widely appreciated.

With regard to the number of cells it is not difficult to show that four cells offer a distinct advantage in performance over three cells.

The transmission loss through the network is less for the frequency shifted in phase by 180°. This has two beneficial effects, first more negative feedback and/or reduction of amplitude phase shift at high frequencies can be achieved, and secondly, for given distortion, a larger voltage amplitude is available from the network output.

There is also a further consideration. The frequency stability of an R-C oscillator of this type is directly related to the rate of change of phase and amplitude of the network output frequencies immediately adjacent to the frequency undergoing 180° phase shift. In the case of a single cell, the greatest $d\phi/df$ occurs at the frequency for which the phase shift is 45° and $R = X$. This condition also corresponds very nearly to the greatest de_o/df. Therefore greatest "effective Q", if this term can be misemployed, would result from a network of four cells each shifting the phase by 45°, as apart from the slightly greater rate of change of phase per cell, the extra fourth cell gives additional phase selectivity. In practice, the shunting of all but the last cell prevents the realisation of this ideal condition in a network of equally proportioned cells, but the four cell arrangement is a nearer approximation than the three cell, and can be expected to have better frequency stability, other factors being equal.

Another factor peculiar to the shunt C network is that each R value for a given frequency is less with the four cell than with the three cell. As these values are already too high for good stability at low frequencies, this reduction is all to the good.

The circuits shown in Figs. 3 and 4, which I presume are for variable frequency application, do not appear to provide for constant network attenuation with change of frequency, and if this interpretation is correct, would appear to have limited practical application.

Finally, I remain to be convinced that the mathematical means offered by Mr. Hinton are so badly needed as would appear from the preface to his article. Personally, I have found that to change three or four resistors initially fitted and producing a frequency differing from that required by a measured amount for the then easily evaluated correct values has much to recommend it from the point of view of time, accuracy and mental effort. If, however, one is in the mood for mental stimulus and cares to assign phase shift requirements per cell, ordinary A.C. methods of successive conversions from admittance to impedance applied from the end cell back to the input can yield a similar result the hard way.

Yours faithfully,

T. S. Dutton.

Quebec, Canada.

The author replies:

Mr. Dutton is probably right to draw attention to statements made without adequate explanation, but it is a pity he did not quote me in full as I did offer a clue to my mental attitude by saying, "There is little to choose between R-C and C-R sections from the point of view of harmonic distortion *as the necessary purity of waveform can be obtained with either kind of oscillator by careful design.*" Likewise, in saying that, ". . . the use of more than three sections is seldom justified," I intended to convey the idea that the advantages of four sections are quite often almost balanced by disadvantages, and to obtain an outstanding improvement over a well designed three-section oscillator it is advisable to consider an alternative technique, such as one described by F. G. Clifford in Electronic Engineering, June 1945, p. 560, which seems to offer distinct advantages in ease of amplitude control.

Before going on to deal with specific points in Mr. Dutton's letter, it may be as well to clarify what I mean by good design practice. First I should say the Engineer should design to fulfil a a definite requirement with due regard to cost, weight, size and simplicity. Secondly, techniques and circuits which are capable of precise design should be chosen so as to reduce experimental work to an absolute minimum. (Here we are talking about design and not development.)

For example, if purity of waveform and stability were important it would be good design practice to employ some thermal lamp-device to keep the amplitude well within the linear characteristic of a class "A" driving amplifier. In this case it would not matter much whether the phase shifting networks were C-R or R-C except that the former eliminates the need for a blocking capacitor and grid leak and this economy might justify the choice of a C-R system. On the other hand, it might be easier to use an R-C system where variable capacitors are used to vary the frequency, but the real point is that the designer is free to choose from considerations other than harmonic distortion.

(A disadvantage of R-C networks is that the finite terminating impedance introduced by the blocking capacitor and grid leak invalidates the design equations, and a choice has to be made between Matrix Algebra and a soldering iron, but a C-R network oscillator can be designed very precisely).

In contrast to the above approach the use of an overdriving technique does not appeal to me if purity of waveform is important, as the designer is then limited in his choice by having to eliminate harmonics he need not have introduced.

Now to deal with the question of three versus four sections. First it is necessary to correct an unfortunate mistake in Mr. Dutton's letter and point out that the greatest rate of change of phase angle with frequency, in a single C-R or R-C section occurs at zero frequency; that is where the capacitive reactance is large compared with the resistance and not when "R = X." This alters the argument as it means that the optimum number of sections is not approximately four. Fig. 1 shows the phase-frequency curves of three, four, and five section networks near the frequency for 180° phase shift; it will be seen that the slope of a four section network is only about 27 per cent steeper than that of a three section network, and this theoretical improvement must be considered in relation to the increase in impedance of the shunt elements with which it is generally accompanied. This is important because my personal experience has been that one cause of frequency drift is the change in grid/cathode capacitance of the valve terminating the network.

My view is that if any uneasiness about stability exists, one is usually concerned to improve it by a factor of at least five times, and analysis confirms my experience that the improvement due to four sections is not outstanding and is seldom justified.

A Zero-Beat Indicator

Dear Sir,—The article describing "A Zero Beat Indicator" by Mr. K. Goldsmith which you published in your December issue warrants some further elucidation.

In the first instance I am very surprised to learn that cathode-ray tubes are widely used in industry as detectors of frequency balance in frequency measuring apparatus, for they are insensitive, cumbersome, require copious ancillary equipment, and it is usually quite difficult to interpret their indications. It is also interesting to hear of a mixer, the output of which is simply the difference of the two input frequencies.

If an amplifier is to be introduced between the mixer and the frequency meter, and it is not to affect materially the operation of the frequency meter, then it must be directly coupled, but this instrument is described as an R-C coupled amplifier (having a time constant of 0.01 sec. and a transformer coupled output) and I fail to see how any amount of positive feedback can achieve the required result.

The claim that the device behaves like a super-regenerative circuit would appear to be questionable, as it is stated that the output carries the input signal as modulation; this does not agree with the usual theory of super-regenerative circuts. It is claimed, too, that the circuit will detect the direction of the frequency deviation from zero, but I cannot see anything in the circuit by which this might be achieved.

If a relay were substituted for the meter it would be quite useless for remote indication as there are frequencies above and below the zero beat where it would operate as well as at the zero beat.

Yours faithfully,

J. E. James.

Crosskeys, Mon.

The author replies:

I do not think that cathode-ray tubes are insensitive for the comparison of frequencies, and that their associated equipment is so copious as to render them useless in industry.

"Beat frequency" is normally defined as the difference between two frequencies, (e.g., Terman, Radio Engineers Handbook, p. 507). When discussing "beat frequency," therefore, I do not see any point in enlarging on any other frequencies which are obtained from a mixer. The word "simply", implying "only", does not occur in the paper.

The amplification of R-C coupled amplifiers is substantially linear between about 200 c/s. and 10,000 c/s. The response of an R-C type frequency meter is usually negligible below 200 c/s. No useful purpose is served, therefore, by improving the amplification at frequencies below, say, 200 c/s. by direct coupling.

At very low output frequencies of the mixer, the self-oscillation of the amplifier is periodically quenched. We are thus dealing with an "on-off" modulation, which, if its frequency is sufficiently low, can be detected on the output meter. I certainly do not agree that the output "carries" the input signal modulation, and I cannot find in the paper the word "carry" suggesting amplitude modulation of the output frequency.

The frequency meter permits only detection of whether the mixer output approaches or recedes from zero frequency. It cannot act as a discriminator, i.e., it does not show whether f_1-f_2 is positive or negative.

It is clear that the device is of use primarily where it is essential to find the actual "zero beat." A simple relay used for remote indication would, of course, operate if the mixer output frequency coincides with the frequency of self-oscillation of the amplifier, and also at the "zero beat" of the oscillator harmonics. These difficulties do not normally arise if the device is used, say, for "null" indication in dielectric measurements. Special precautions may, however, have to be taken in some special cases.

The Design and Limitation of D.C Amplifiers

Dear Sir,—In the recent article on D.C. amplifiers by E. J. Harris and P. O. Bishop (Electronic Engineering, October 1949), an underestimation of the effect of H.T. variation seems to have been made.

The authors neglected the out-of-phase voltage which appears at the anodes of the first stage when H.T. is changed. However, vigorous analysis shows that this source, so far from being negligible, makes much the longest contribution to the total output change.

The approximate expression for the equivalent signal input e, which produces the same out-of-phase output as an H.T. change E, is

$$e = (1/\mu - 1/\mu_2)\,E$$

Substituting the following values: the difference between μ_1 and μ_2 = 20 per cent, their mean value = 20, and E = 1 volt, we obtain:

$$e = 10 \text{ mV.}$$

This is about 700 times larger than the value calculated by Messrs. Harris and Bishop, and as a matter of fact is of the right order compared with values measured in practice.

Yours faithfully,

A. T. Fuller.

Hitchin, Herts.

NOTES FROM THE INDUSTRY

National Radio Exhibition

The National Radio Exhibition, which is to be held at Castle Bromwich, Birmingham, from September 6 to 16 next, will be at Earls Court, London, in September 1951, 1952 and 1953.

The Radio Industry Council announced that an earlier decision to hold the exhibition at Olympia, London, in June instead of in the autumn had been changed, but this change came too late to secure convenient autumn dates and the necessary space at Olympia.

In view of the facilities offered at Earls Court, it was decided to go there rather than hold the exhibition at Olympia at a less convenient time.

The National Radio Exhibition before going to Olympia in 1926 was held at the Royal Albert Hall in 1924 and 1925, the White City in 1923 and the Horticultural Hall in 1922. Sixteen National Radio Exhibitions have been held in all. The one coinciding with the Festival of Britain will be the 18th.

Televising the Boat Race

The University Boat Race on April 1 will be televised on an even more ambitious scale than last year, and B.B.C. engineers and outside broadcast staff are now working on arrangements employing practically the full mobile resources of the television service as well as many additional technical facilities provided by specialist departments.

This year there will be four mobile units in action along the river bank from Putney to Mortlake using eleven cameras in addition to a waterborne camera following the crews in the launch *Consuta*. The launch will carry its own sound transmitter to provide a special commentary for viewers.

By spacing the cameras along the course it is believed that the Varsity crews could be kept in sight almost continuously from the shore-based cameras as well as from the camera in the launch.

R.F. Heating

We regret that on page 104 of the March issue of ELECTRONIC ENGINEERING the equation

$$\frac{\text{Total Power Input}}{\text{R.F. Power Output}}$$

became inverted. It should have read

$$\frac{\text{R.F. Power Output}}{\text{Total Power Input}}$$

Most Northerly Broadcasting Station in Norway

In 1933 the most northerly broadcaster in the world was installed in Vadso in the province of Finmark in Norway. This station, which was of 10 kW, was within the Arctic circle and furnished programmes to the northern half of Norway and Lapland. During the occupation of Norway the station was destroyed by the Germans.

A new station has now been built and a 20 kW broadcaster was put into service recently to cover the same area, where about 45,000 people are scattered over 19,000 square miles.

Both the original transmitter and the replacing transmitter were manufactured by Standard Telephones and Cables, Ltd., in England, and supplied through their associate, Standard Telefon og Kabelfabrik A/S, Oslo.

United Nations Post

Mr. H. B. Rantzen, head of Engineering Designs Department, B.B.C., has accepted an appointment with the United Nations as Director of Telecommunications Services. Mr. Rantzen, after 20 years with the B.B.C., will leave at the end of March to take up his new duties in New York.

Mr. Rantzen's new post will give him great responsibility in many fields where electronics lend their important facilities to much of the work at the United Nations. The conference halls and committee rooms employ public address systems, ultra short-wave transmitters and midget portable receivers for simultaneous interpretation in five languages.

Changes of Address

The head office of RCA Photophone, Ltd., has lately been transferred to 36 Woodstock Grove, Shepherds Bush, W.2.

Information on RCA radio equipment and valves is obtainable from the Engineering Products Department at the same address. (Telephone: Shepherds Bush 1200.)

Messrs. Fielden (Electronics), Ltd., on the completion of their new factory, have moved to Paston Road, Wythenshawe, Manchester.

Messrs. F. C. Robinson & Partners, Ltd., have moved from their old address to 287 Deansgate, Manchester 3; telephone numbers Deansgate 6607 and 6602.

Publications Received

A Review of G.E.C. Progress in 1949.

This is an account of the company's progress and activities during the past year. Further details of any aspect of this report may be obtained from the *General Electric Co., Ltd., Magnet House, Kingsway, London, W.C.2.*

Synopsis of Cintel Instruments.

This leaflet lists the full range of electronic instruments made by *Cinema-Television, Ltd., Worsley Bridge Road, Lower Sydenham, London, S.E.26.*

Export Marketing.

This booklet reviews the export position during the last 20 years, and gives suggestions for improving individual firms' production. It can be obtained from *The Osborne-Peacock Co., Ltd., 47 Cannon Street, E.C.4.*

G.E.C. Sound Equipment.

The range of equipment described in the publication is built to a system of rack assembly, whereby any number and combination of sound inputs, amplifiers, remote control and monitoring devices, switching gear, etc., can be made up to form a complete equipment. *The General Electric Co., Ltd., Magnet House, Kingsway, W.C.2.*

Permanent Magnets.

The purpose of this pamphlet is to provide a quick reference to the "Eclipse" standard ranges of permanent magnets. *James Neill and Co. (Sheffield), Ltd., Composite Works, Sheffield* 11.

Review of Activities during 1949.

This is an outline of the achievements and interests of the *English Electric Co., Ltd., Stafford,* during 1949.

The BEAMA.

A short account of the BEAMA's growth and activities. Copies obtainable from the *British Electrical and Allied Manufacturers' Association, 36 and 38 Kingsway, London, W.C.2.*

DC/AC Convertors.

Some recent literature on DC/AC convertors published by *Valradio, Ltd., 57 Fortress Road, Kentish Town, N.W.5.*

Marconi Broadcasting.

This booklet describes recent advances in broadcasting transmitters and general equipment manufactured by *Marconi Wireless Telegraph Co., Ltd., Marconi House, Chelmsford,* and some of their uses.

MEETINGS THIS MONTH

INSTITUTION OF ELECTRICAL ENGINEERS

Unless otherwise stated, all meetings are held at 5.30 p.m. at the Institution of Electrical Engineers, Savoy Place, London, W.C.2.

Measurements Section

Date: April 4.
Lecture: Dielectric Admittances in Transformers.
By: A. H. M. Arnold, Ph.D., D.Eng.

Date: April 18.
Discussion: Temperature Measurement.
Opened by: J. A. Hall, B.Sc., and C. R. Barber, B.Sc.

Radio Section

Date: April 12.
Lecture: A Review of some Television Pick-up Tubes.
By: J. D. McGee, M.Sc., Ph.D.
Lecture: The Design of a Television Camera Channel for Use with the C.P.S. Emitron.
By: E. L. C. White, M.A., Ph.D. and M. G. Harker, B.Sc (Eng.).

Date: April 24.
Discussion: The Relation between Production, Operation and Maintenance of Service Radio Equipment.
Opened by: D. H. Hughes.

Cambridge Radio Group

Date: April 18. Time: 8.15 p.m.
Held at: The Cavendish Laboratory, Cambridge.
Lecture: The Structure, Electrical Properties and Applications of the Barium-Titanate Class of Ceramic Materials.
By: Professor Willis Jackson, D.Sc., D.Phil.

North-Eastern Centre

Date: April 20. Time: 7 p.m.
Held at: The Literary and Philosophical Society, Westgate Road, Newcastle-on-Tyne.
Lecture: The Faraday Lecture on "Radar".
By: R. A. Smith, M.A., Ph.D.

North-Eastern Radio and Measurements Group

Date: April 3. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: Radar Automatic Tracking.
By: F. J. V. Ritson, B.Sc.

Scottish Centre

Date: April 18. Time: 7 p.m.
Held at: Royal Technical College, Glasgow.
Lecture: Faraday Lecture on "Radar."
By: R. A. Smith, M.A., Ph.D.

South Midland Radio Group

Date: April 27. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Energy Conversion Devices for Electrical and Electronic Measurement of Non-Electrical Quantities.
By: J. C. Finlay.

Southern Centre

Date: April 19. Time: 6.30 p.m.
Held at: The Technical College, Brighton.
Lecture: The Continental Development of Single-Anode Mercury-Arc-Rectifier Valves for High Power.
By: H. von Bertele.

BRITISH SOUND RECORDING ASSOCIATION

Date: April 21. Time: 7 p.m.
Held at: Royal Society of Arts, John Adam Street, W.C.2.
Lecture: Practical Microgroove Recording and Reproduction.
By: A. R. Sugden and R. W. Lowden.

SECRETARIES OF ASSOCIATIONS

INSTITUTION OF ELECTRICAL ENGINEERS

The Secretary, Institution of Electrical Engineers, Savoy Place, W.C.2.

Cambridge Radio Group

G. E. Middleton, M.A., University Engineering Laboratory, Cambridge.

North-Eastern Radio and Measurements Group

G. A. Kysh (Asst. Sec.), Carliol House, Newcastle-on-Tyne, 1.

North-Western Radio Group

A. L. Green (Asst. Sec.), 244, Brantingham Road, Chorlton-cum-Hardy, Manchester, 21.

South Midland Radio Group

W. H. Brent, B.Sc., Regional Director's Office, Midlands Region (G.P.O.), Civic House, Great Charles Street, Birmingham, 3.

BRITISH INSTITUTION OF RADIO ENGINEERS

The General Secretary, 9, Bedford Square, London, W.C.1.

Scottish Section

A. A. M. Turnbull, A.M.Brit.I.R.E., 68, Launderdale Gardens, Glasgow, W.2.

West Midlands Section

R. A. Lampitt, A.M.Brit.I.R.E., 20, Northfield Grove, Merry Hill, Wolverhampton.

INC. RADIO SOCIETY OF GT. BRITAIN

General Secretary, New Ruskin House, Little Russel Street, W.C.1.

BRITISH SOUND RECORDING ASSOCIATION

Richard W. Lowden, "Wayford," Napoleon Avenue, Farnborough, Hants.

TELEVISION SOCIETY

Constructors Group

G. T. Clack, 10, Tantallon Road, Balham, London, S.W.12.

Midland Centre

W. Summer, 31, Beach Road, Bournville, Birmingham, 30.

Leicester Centre

A. E. Moore, 110, Doncaster Road, Leicester.

RADAR ASSOCIATION

Technical Committee: J. C. Warren, 57, Gallows Hill, Kings Langley, Herts.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

W. H. Fox, A.M.I.E.E., Engineer-in-Chief's Office (T. P. Branch), Alder House, E.C.1.

INSTITUTION OF ELECTRONICS

Lecture Sec.: W. Summer, 31, Beech Road, Bournville, Birmingham, 30.

North West Branch

L. F. Berry, 105, Birch Avenue, Chadderton, Lancs.

SOCIETY OF RELAY ENGINEERS

T. H. Hall, M.Brit.I.R.E., 23, Dalkeith Place Kettering, Northants.

INC. RADIO SOCIETY OF GREAT BRITAIN

Date: April 28. Time: 6.30 p.m.
Held at: Institution of Electrical Engineers, Savoy Place, W.C.2.
Lecture: Mobile V.H.F. Equipment.
By: J. R. Brinkley, A.M.Brit.I.R.E.

INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: April 19. Time: 5 p.m.
Held at: The Conference Room, 4th Floor, Waterloo Bridge House, S.E.1.
Lecture: Applied Mathematics in the Telephone Area.
By: E. Hopkinson, B.Sc., A.M.I.E.E.

SOCIETY OF RELAY ENGINEERS

Date: April 25. Time: 2.30 p.m.
Held at: Lecture Hall at the Royal Society of Arts, John Adam Street, W.C.2.
Lecture: Wired Television.
By: A Technician from Messrs. Link Sound and Vision Services, Ltd.

INSTITUTION OF ELECTRONICS

North West Branch

Date: April 18. Time: 6.30 p.m.
Held at: The Reynolds Hall, College of Technology, Manchester.
Lecture: Cathode Ray Tubes for Television.
By: J. A. Darbyshire, M.Sc., Ph.D., F.Inst.P.

TELEVISION SOCIETY

Main Society and Constructors' Group

Date: April 14. Time: 7 p.m.
Held at: The Cinema Exhibitor's Association, 164, Shaftesbury Avenue, W.C.2.
Lecture: The Correction of Telephone Lines for Television Outside Broadcasts.
By: T. Kilvington.

Midlands Centre

Date: April 3. Time: 7 p.m.
Held at: Lecture Hall (1st Floor), The Crown Restaurant, Corporation Street, Birmingham.
Lecture: Producing for Television.
By: George More O'Ferrall.

Leicester Centre

Date: April 5. Time: 7 p.m.
Held at: Room 104, The College of Art and Technology, Leicester.
Lecture and Demonstration: Large Screen Television Projection Unit.
By: Mullard Electronics Ltd.

BRITISH INSTITUTION OF RADIO ENGINEERS

Date: April 20. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Lecture: U.H.F. Propagation and Characteristics.
By: D. W. Heightman.

West Midlands Section

Date: April 26. Time: 7 p.m.
Held at: Wolverhampton and Staffs. Technical College, Wulfruna Street, Wolverhampton.
Lecture: Intermodulation Analysis.
By: C. R. Amey.

Scottish Section

Date: April 6. Time: 6.45 p.m.
Held at: Institution of Engineers and Shipbuilders, Glasgow.
Lecture: Electrical Measurements.
By: F. M. Bruce, M.Sc., Ph.D.
Joint Meeting with the Institute of Physics.