WIRELESS ENGINEER

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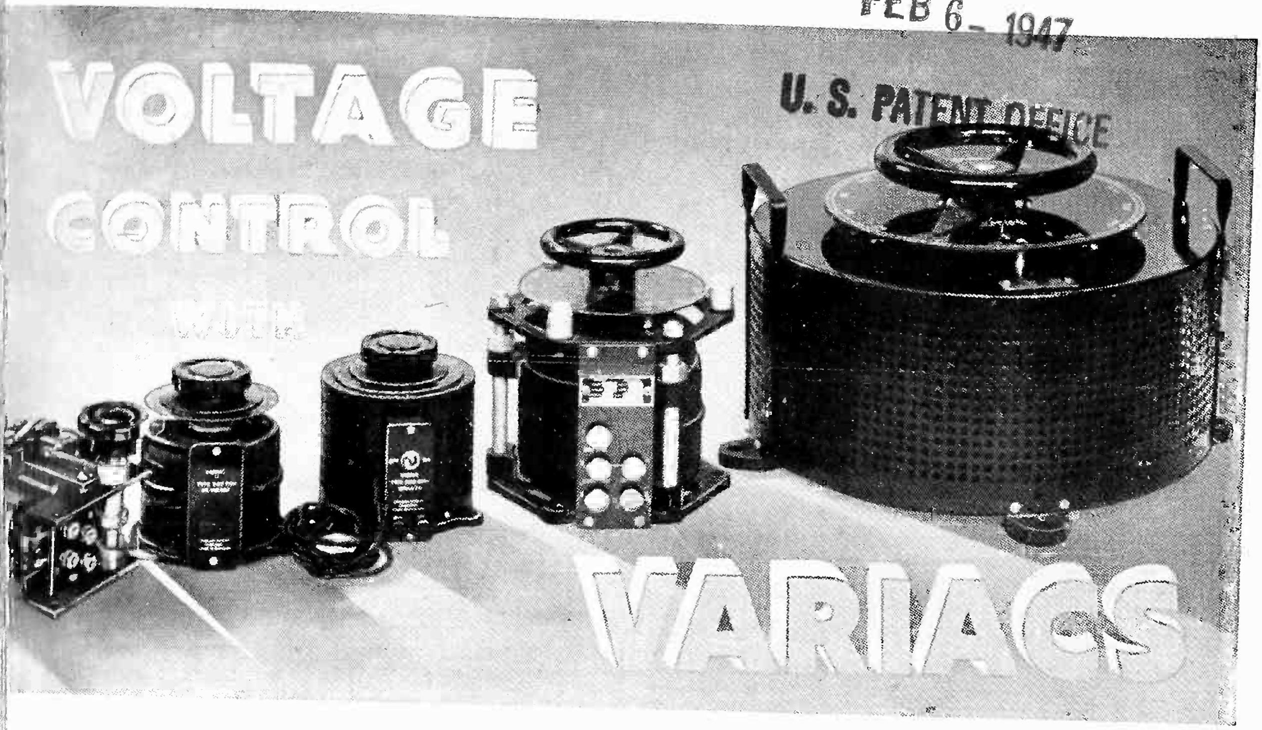

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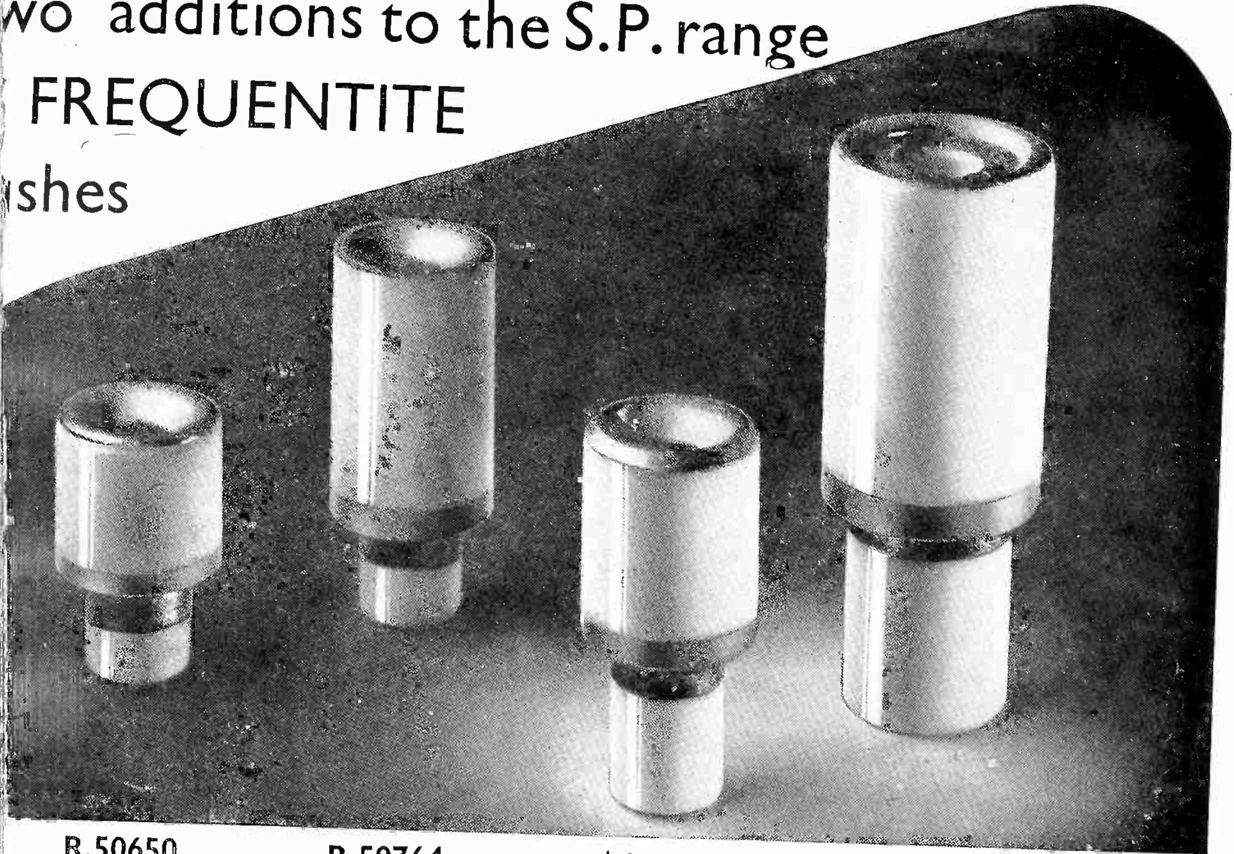

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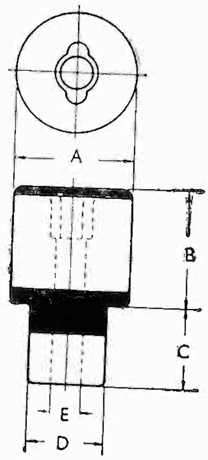

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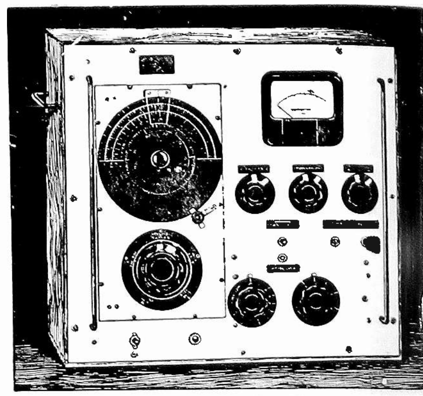

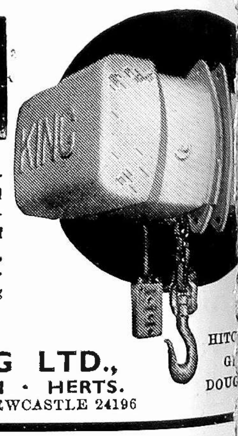

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WIRELESS ENGINEER

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EDITORIAL

The Use of Analogies

SOME years ago a student complained that whenever he asked a question about an electric circuit he was told to imagine a pipe full of water, etc. He said that he knew all about the pipe full of water but wanted to know about the electric circuit and he disapproved of being fobbed off with hydraulic analogies. We were reminded of this by some statements made by Professor Livens in a recent paper* on "The energy and mechanical relations of the magnetic field" in which he criticizes the use of magnetic and electric analogies as well as magnetic and mechanical analogies. He says, "The persistent adherence to a supposed analogy with the electrostatic case, which does not exist, and to frequently fanciful analogies of the stress-strain type, which have introduced more confusion than they have saved, seem to have blinded to the real solution those few authors who were able to see clearly that there are anomalies in the usual discussion, and the usual dimensional treatment which starts by assuming—again on the basis of some supposed stress-strain analogy—that the nature of the vectors of force and induction in the field must be different, has perhaps not helped in any way to clear the air for a more satisfactory treatment."

This raises some interesting and important points, especially for those engaged in teaching. Electrical engineering differs from mechanical engineering in that the student of the latter is dealing with concepts, such as force, mass, deflection, temperature, etc., with which he has been familiar since childhood, whereas the electrical student is introduced to a system of new and intangible concepts, such as potential difference, electromotive force, current, magnetic induction, etc. It is difficult to see how these can be explained without the use of analogies; the obvious way of explaining new and intangible concepts is by means of familiar and tangible concepts. It is, of course, essential to emphasize the danger of pushing the analogies too far; they should perhaps be regarded not as analogies but rather as parallel lines of thought. It is surely better for a student to have even a crude mental picture of what is happening in electric and magnetic phenomena than to have no mental picture at all, but to regard it all as a collection of words and symbols. The nomenclature of electromagnetism shows that it was largely based on mechanical analogies; the words current, pressure, tension, electromotive force, resistance, all suggest mechanical lines of thought, and the natural explanation of the unknown in terms of the known. We disagree with Professor Livens' contention that analogies of the stress-strain type have introduced more confusion than they have saved; on the contrary we think that the confusion that existed a few years ago with regard to *B* and *H* has been eliminated largely owing to the application of the stress-strain line of

*Phil. Mag. January 1945, Vol. XXXVI, p. 1.

thought, or, as some might prefer to call it, the cause and effect line of thought. When current passes round a coil it produces a magnetic field; when a pull is applied to a rod it produces an extension. The strain produced at any point in the material of the rod depends on the stress (i.e., the pull per unit area at the point) and on a property of the material. If one can assume that the stress is uniform over the cross-section it can be calculated by dividing the total pull by the cross-sectional area. Similarly the magnetic induction B produced at any point, say, in a mass of iron, must depend on some cause which we call the magnetizing force H at the point and some property of the material which we call the permeability. Instead of the total prime cause being distributed over a cross-sectional area as the pull is in the mechanical specimen undergoing a tension test, it is distributed over the length of a closed path. The prime cause in the magnetic case is the current or ampere-turns and the localized cause H is the current per unit length—neglecting such details as 4π.

Is Magnetizing Force a Force?

Just as in the special case of the stress being uniform over the cross-section, the localized cause is obtained by dividing the total cause by the area, so, in the special case of the uniformly wound toroid, the localized cause H is obtained by dividing the total cause, viz. the total current or ampere-turns, by the length. In both cases one has the total cause, the localized cause, and the localized effect depending on some property of the material, and we fail to see how such parallel lines of thought—or fanciful analogies as Professor Livens calls them—can introduce confusion or blind anyone to the real solution. We gather from Professor Livens' paper that there is some confusion due to a misunderstanding of the term "magnetic force" or "magnetizing force." When one refers to H as the magnetizing force one does not use the word "force" in its mechanical sense; the unit of H is not the dyne, although junior students sometimes make the mistake and give a field strength as so many dynes. It is true that H multiplied by a pole strength m gives a force in dynes, but this only shows that H itself cannot be a force in the mechanical sense. When we speak of a magnetic force or magnetizing force we mean the intensity of the magnetizing cause acting at a point. When one speaks of the force of circumstances or the force of an argument, one is using the word "force" to designate something that cannot be measured in dynes and, although somewhat far-fetched, these examples may serve as a warning against regarding H, by whatever name it may be called, as a mechanical force.

Cause (H) and Effect (B)

Now the mechanical force on a conductor carrying a current or on a moving electron depends on the magnetic induction or flux density B in which it is situated and not on the causative force H which produced the magnetic induction. This fact has led some people to suggest that B and not H should be called the magnetic or magnetizing force. In the paper to which we have referred, for example, Professor Livens says: "This last step in our thesis involves a complete inversion of the roles of B and H," and again "If this thesis is accepted, then we must prepare for a complete inversion of the whole of our traditional notions on this subject, that is, of course, if we admit that the claims of theoretical consistency must ultimately take precedence over those of traditional and almost universal and long standing practice." Apparently the idea is to refer to the magnetic induction B as the magnetizing force and to the magnetizing force H as the induction. We cannot imagine for a moment that Professor Livens expects his suggestion to be adopted.

If instead of an iron toroid linked by a current we have a dielectric toroid linked by a changing magnetic flux, the magnetic force or cause H is replaced by the electric force or cause $\mathcal{E}$ which depends only on the changing flux and the length of path and is independent of the material of the toroid. It is the causative force acting at any point and its effect will depend on the material of the toroid. I has been replaced by $d\phi/dt$ and μ by κ. Surely nothing but good can come from drawing the student's attention to these parallel lines of thought; it is one's duty to do so. In each case one can divide the effect into two parts, viz. that which would exist in the absence of the magnetic material or dielectric and that due to its presence, thus getting the two formulae

$$B = \mu_0 H + 4\pi J$$

and $4\pi D = \kappa_0 \mathcal{E} + 4\pi P$
in which J is the intensity of magnetization, D the displacement and P the polarization of the dielectric. Also

$$\mu = \frac{B}{H} = \mu_0 + 4\pi J/H$$

and $$\kappa = \frac{4\pi D}{\mathcal{E}} = \kappa_0 + 4\pi P/\mathcal{E}$$

For the energy per cm³ we have in one case $\mu H^2/8\pi$ and in the other $\kappa \mathcal{E}^2 8\pi$.

In view of these formulae it is difficult to understand how anyone can complain of the "persistent adherence to a supposed analogy which does not exist." The parallelism is so striking that it should be emphasized when teaching the subject, but always with a warning that parallel lines of thought and symbolism do not indicate any physical likeness in the phenomena involved. G.W.O.H.

DIPOLE WITH UNBALANCED FEEDER*

By D. A. Bell, M.A., B.Sc.

IN the course of their paper on "Aerial Impedance Measurements" L. Essen and M. H. Oliver[1] mention that a half-wave dipole can be connected directly to a concentric feeder cable without any great impedance mis-match, but that they have not investigated the symmetry of the field.

The complementary observations have been made by the writer during the examination of directional aerials of the type commonly used for television reception, namely a vertical dipole associated with a single half-wave parasitic element. The aerial under examination was connected to a receiver, a low-power oscillator was set up about 10 wavelengths away, and a polar diagram of the aerial was measured by rotating the aerial supporting mast about its (vertical) axis and plotting received signal strength against the azimuthal angle of the aerial assembly. Two alternative methods of connecting the aerial to the receiver were available:

(i) Through screened and balanced twin cable and a balanced coupling-coil to the first tuned circuit of the receiver.

(ii) Through coaxial cable and a coupling coil having one side earthed and joined to the outer of the coaxial feeder. The coaxial cable was connected directly to the aerial, the central conductor to one limb of the dipole and the outer conductor to the other limb.

Preliminary tests with a simple dipole failed to reveal any substantial difference in signal strengths between the two feeder systems; but with the addition of a director adjusted to give maximum suppression in the backward direction, the balanced feeder gave a symmetrical polar diagram while the coaxial feeder gave an asymmetrical diagram. These are shown by the full line and dotted curves respectively in Fig. 1. This asymmetrical characteristic can produce a sharper minimum than would otherwise be available; but the exact shape of the characteristic is likely to vary with aerial site and feeder length, since it depends upon the "vertical-aerial pick-up" effect of the whole elevated system and upon the impedance of the outer conductor of the feeder.

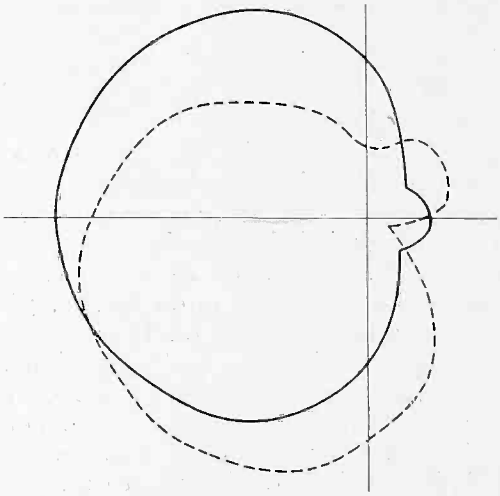

Fig. 1. *Polar diagrams of vertical dipole and director with balanced (solid line) and unbalanced (dotted line) feeders.*

A balanced aerial should preferably be connected to a coaxial feeder through a balancing device, of which the simplest form

*MS. accepted by the Editor, July 1946.

is the quarter-wave "sleeve" as shown by Essen and Oliver in their Fig. 8, and the equivalent circuit shown in Fig. 2 gives an idea of the behaviour of a centre-fed half-wave dipole connected in this way. The symmetrical dipole is represented by a source of e.m.f. E_1 at the centre of a resistance of 76 ohms total value, and the terminals A, B of the dipole are connected to a feeder which is assumed to be of 76 ohms characteristic impedance and non-reflectively terminated: so far it makes no difference whether the feeder consists of balanced-twin or coaxial cable, and no "earth" has been introduced. Practical dipoles, however, are not mounted in free space but usually near an "earth," whether this be the surface of the ground, the body of an aircraft, or the general structure of a tower on which the aerial is mounted.

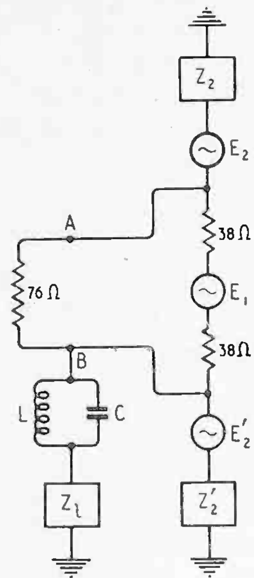

Fig. 2. *Equivalent circuit of dipole aerial.*

Now it is well-known that an elevated conductor connected to earth by a wire constitutes an aerial system: an alternating current can be established in the wire by a radio wave, and this current can be detected by a suitable instrument as shown in Fig. 3(a). Moreover, it is known that this phenomenon can be represented by an equivalent circuit of the type illustrated in Fig. 3(b), where the aerial is replaced by a voltage generator E in series with an "aerial impeddance" Z_{ae}. Returning now to the dipole, when associated with a feeder cable going down to a receiver situated on the "earth" each limb of this is equivalent to an aerial of the type of Fig. 3(a), and according to the equivalent circuit of Fig. 3(b) this fact is indicated in Fig. 2 by adding the generators E_2 and E'_2 each in series with an appropriate impedance Z_2, Z'_2. Since the two limbs are similarly situated above the earth, the two generators will be in the same phase, and so long as the load connected between A and B is balanced to earth, they will introduce no net difference of potential between these terminals. But if, for example, terminal B were directly earthed, the generator E_2 would be connected in series with Z_2 and the load, and would cause a voltage to appear across the load; i.e., across the receiver input circuit.

In most cases the outer of the concentric feeder is earthed at the receiving end of the feeder, not at the aerial end, and the impedance to earth presented at B is then a function of feeder length: for odd multiples of $\lambda/4$ it is very high and for even multiples very low if the transmission line made up of cable outer and earth is of low loss. In practice, there is likely to be considerable loss due to the waterproof covering over the outer of the cable and to miscellaneous objects between cable and earth; but it is perhaps worth noting also that Schelkunoff[2] has given figures for the input impedance of an infinitely long conductor in free space, and for television feeder and frequencies as used by the writer it is of the order of 500 ohms. Since losses will damp out any resonance effects, it seems probable that the impedance to earth presented by the length of outer conductor of the coaxial feeder would be of the order of 500 ohms.

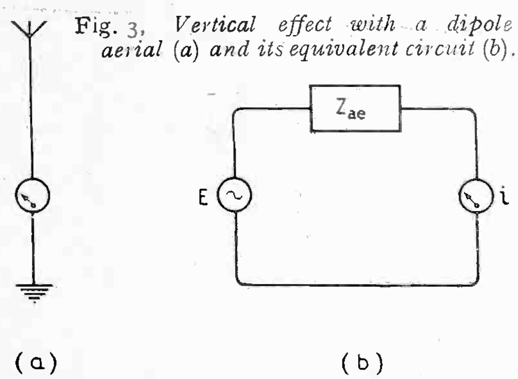

Fig. 3, *Vertical effect with a dipole aerial (a) and its equivalent circuit (b).*

This is the impedance represented by Z_l in Fig. 2, and two deductions can be made from the magnitude assigned to it: (i) So long as the load impedance between A and B is small compared with 500 ohms, the generator E_2 is likely to produce only a secondary effect, and (ii) any impedance (such as that represented by L, C in Fig. 2) which may be added in order to reduce this effect must be appreciably greater than 500 ohms. The resonant-line analogue of the parallel-resonant circuit L, C of Fig. 2 is a short-circuited quarter-wave line, or in other words the quarter-wave "sleeve" on the outer of the concentric cable to which reference has already been made, and which is illustrated in more detail in Fig. 4. As an alternative to the general explanation that a short-circuited quarter-wave line presents a high impedance at the open end, the mode of operation of the sleeve may be interpreted as follows. Assuming the sleeve to be a perfect conductor, it will act as a screen so that any current in the cable produces no external field; and this implies that for any current i_1 down the conductor 1 (the outer of the coaxial cable), there must be an equal and opposite current $-i_1$ along the inside of the sleeve, and these currents will balance at the closed end of the quarter-wave so as to leave no net current to flow down the remainder of the feeder. The resonance of the quarter-wave is necessary to minimize the unbalancing of the dipole due to attaching this impedance to one limb of it but not to the other. (There are more elaborate balance-to-unbalance transformers which maintain a higher degree of symmetry but also function as "stubs" in parallel with the balanced circuit.)

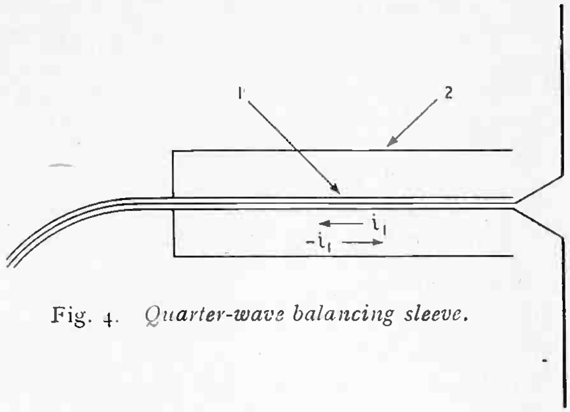

Fig. 4. *Quarter-wave balancing sleeve.*

This note is based on work carried out in the Research Laboratories of A. C. Cossor Ltd., and acknowledgments are due to Mr. L. H. Bedford, Director of Research, and to other members of the Research Staff with whom these questions have been discussed.

REFERENCES

1 *Wireless Engineer*, December 1946, Vol. XXII, p. 587.

2 "Theory of Antennas of Arbitrary Size and Shape," by S. A. Schelkunoff, *Proc. Inst. Rad. Engrs.*, September 1941, Vol. 29, p. 493.

THEORY OF THE EQUIVALENT DIODE*

By G. B. Walker

(The Mullard Radio Valve Co., Ltd.)

SUMMARY.—The theory of the equivalent diode is discussed and a new method, based on electrostatic considerations, is suggested whereby the equivalent diode can be uniquely determined.

Introduction

THE "equivalent" diode is a concept which has always lacked a precise definition, and like some other concepts in electronic engineering has taken on a different colouring to suit the taste of the user. The importance of the concept lies in the fact that it makes a rough calculation of the current passed by a triode possible, but there can be no doubt that if the physics of the triode were more simple the concept would never have arisen. One might be led to believe, therefore, that the equivalent diode is no more than a useful device, essentially arbitrary and having no real existence, but we feel that the logic of the question still merits consideration.

The supposition that a diode can be constructed which will have the same current-voltage relationship as a given triode has a practical basis in the discovery by Van der Bijl[1] in 1913 that for a limited range of applied voltages the current passed by a triode obeys the empirical law—

$$I = k(V_a + \mu V_g + \epsilon)^{\beta} \qquad \qquad (1)$$

k, μ, ϵ and β being constants.

The form of this law, in which the applied voltages are combined in a linear term, together with the experimental fact that β has a value in the neighbourhood of 3/2 indicates that over the limited voltage range a triode behaves as a diode. This is the only justification for an attempt to form the conception of an equivalent diode but we must appeal to theory for aid in advancing

* MS accepted by the Editor, June 1946.

the matter. It is at this point that a divergence of opinion occurs.

The Electrostatic Approach

The favourite approach to the question, and in the writer's opinion the most consistent, as will be shown later, is to postulate that a diode is equivalent to a triode if, when the cathode is cold and no current is flowing, the off-cathode electrostatic field strength is the same in both valves. Let us examine this requirement for the case of a triode having planar electrodes.

The average value of the off-cathode field strength in the absence of space current can be written—

$$F = \frac{V_a + \mu V_g}{a + \mu g} \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

where g is the spacing between cathode and grid and a is the spacing between the cathode and anode. μ is the amplification factor of the valve calculated on electrostatic grounds and depending only on the geometry of the valve.

For the equivalent diode we can write

$$F = \frac{V_d}{d} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

where d is the spacing between cathode and anode.

Thus

$$\frac{V_d}{d} = \frac{V_a + \mu V_g}{a + \mu g}$$

This equation is insufficient to determine both V_d and d, so let us write

$$V_d = \alpha(V_a + \mu V_g)$$

and

$$d = \alpha(a + \mu g)$$

where α is some, so far undetermined, dimensionless constant.

Now the current passed by the diode is

$$I = 2.34 \times 10^{-6} \frac{V_d^{\frac{3}{2}}}{d^2}$$

(omitting the term ϵ in (1) which takes contact-potential differences into account), and so for the triode we have

$$I = 2.34 \times 10^{-6} \frac{(V_a + \mu V_g)^{\frac{3}{2}}}{\sqrt{\alpha}\,[a + \mu g]^2}$$

The difficulty now is to evaluate α or, which is the same thing, find the proper spacing in the equivalent diode. Many plausible though divergent arguments have been given which need not be discussed here.

Most writers give it the value $\frac{1}{\mu}$, (c.f. Dow[2] and Benham[3]), the latter coming to this decision after considering no less than six alternatives. There can be no doubt that in most cases the value should be in the neighbourhood of $\frac{1}{\mu}$ as current measurements on a triode will confirm, but let us examine what happens when μ tends to zero. In the limit this is equivalent to the removal of the grid from the valve and we can deduce, therefore, that the current passed should be

$$I = 2.34 \times 10^{-6} \frac{V_a^{\frac{3}{2}}}{a^2}$$

whereas according to (7) if $\alpha = \frac{1}{\mu}$ and μ tends to zero I vanishes!

The Current Approach

To obviate the difficulty in determining α and at the same time to pay greater attention to the behaviour of space charge in the valve some writers have rejected the electrostatic approach. We shall now examine an ingenious method given by Fremlin[4] which illustrates the essential features of the current approach.

Fremlin accepts the expression

$$I = k\,(V_a + \mu V_g)^{\frac{3}{2}} \quad \ldots \quad \ldots \quad (9)$$

and evaluates k in the following way.

If the grid of the triode is removed the anode current is,

$$I = 2.34 \times 10^{-6} \frac{V_a^{\frac{3}{2}}}{a^2}$$

or

$$V_a = \frac{I^{\frac{2}{3}}}{(2.34 \times 10^{-6})^{\frac{2}{3}}}\, a^{\frac{4}{3}} \quad \ldots \quad (10)$$

The potential at the plane where the grid was, is

$$V_g = \frac{I^{\frac{2}{3}}}{(2.34 \times 10^{-6})^{\frac{2}{3}}}\, g^{\frac{4}{3}} \quad \ldots \quad \ldots \quad (11)$$

and so if the grid is replaced and maintained at that potential no change should occur in the current passed by the valve. Thus, by substituting (10) and (11) in (9) he obtains

$$I = k\left[\frac{I^{\frac{2}{3}}\, a^{\frac{4}{3}}}{(2.34 \times 10^{-6})^{\frac{2}{3}}} + \mu \frac{I^{\frac{2}{3}}\, g^{\frac{4}{3}}}{(2.34 \times 10^{-6})^{\frac{2}{3}}}\right]^{\frac{3}{2}}$$

$$\text{or } k = \frac{2.34 \times 10^{-6}}{[a^{\frac{4}{3}} + \mu g^{\frac{4}{3}}]^{\frac{3}{2}}}$$

$$\text{and so } I = 2.34 \times 10^{-6} \frac{(V_a + \mu V_g)^{\frac{3}{2}}}{[\alpha^{\frac{4}{3}} + \mu g^{\frac{4}{3}}]^{\frac{3}{2}}} \quad \ldots \quad (12)$$

Apparently the equivalent diode has disappeared from the argument, but if we compare expressions (12) and (7) we see that in Fremlin's treatment α is given the value $\frac{[a^{\frac{4}{3}} + \mu g^{\frac{4}{3}}]^3}{[a + \mu g]^4}$. In this way it is possible to define an equivalent diode.

Critical Discussion

Let us examine the implications in the use of the $\frac{3}{2}$-power voltage law.

It has been shown by Langmuir and Compton[5] that the current passed by a diode, irrespective of the shape of the electrodes, is proportional to the $\frac{3}{2}$ power of the anode voltage on the one condition that electrons leave the cathode with no initial velocity. In the proof the fact emerges that the potential at any point in the field is proportional to the anode voltage. It cannot be disputed, therefore, that Fremlin's formula for the current passed by an ideal triode (ideal in that there is no emission velocity) is correct, but only for values of grid and anode voltage such that $\dfrac{V_g}{V_a} = \left[\dfrac{g}{a}\right]^{\frac{4}{3}}$.

For the law (1) to have any practical value it is essential that V_g and V_a can be varied independently, or to put the matter otherwise, it is essential that μ is constant. Now in a valve, μ is approximately constant only under two conditions :—

(1) that space charge is negligible in the region between the grid and the anode, and

(2) that the disturbance of the electric field by the wires of the grid does not extend so far as to cause field irregularities at the cathode.

We can assume that condition (2) is satisfied in Fremlin's case but condition (1) is certainly not. In fact it is necessary for him to suppose that space charge is "saturated" between the grid and anode. This is undoubtedly a weakness in his position. From a practical point of view the part of the triode characteristic which most nearly conforms to Van der Bijl's empirical law corresponds to a range of grid voltage in the neighbourhood of zero. In this range μ is hardly affected by space charge and here only is it legitimate to combine the grid and anode voltages in a linear term.

A New Suggestion

To summarize the position at which we have arrived, we may say that law (1) applies only when it is impossible to tell whether the valve is a diode or a triode by observing the field at the cathode and when space charge has a negligible influence on the field beyond the grid. Now in the electrostatic approach we postulated that in the absence of space charge the electric field strength at the cathode must be identical in the triode and equivalent diode. Let us postulate further that the variation in that field strength with a change in cathode voltage is also the same for both. This is not a new fact but is a fuller statement of identity.

In the general case, the field a the triode cathode can be written as a linear function of the applied voltages as follows :—

$$F = A[(V_a - V_k) + \mu(V_g - V_k)]$$

and for the diode

$$F' = A'(V_d - V_k)$$

where A, A' and μ depend on the valve dimensions only. In addition to the condition

$$F = F'$$

we have also $\dfrac{\partial F}{\partial V_k} = \dfrac{\partial F'}{\partial V_k}$

$$\text{thus } A[(V_a - V_k) + \mu(V_g - V_k)] = A'(V_d - V_k)$$

$$\text{and } A(1 + \mu) = A'$$

Writing $V_k = 0$ we obtain

$$V_d = [V_a + \mu V_g]/(1 + \mu)$$

$$\text{or } \alpha = 1/(1 + \mu)$$

and the corresponding expression for the current with planar electrodes becomes

$$I = 2.34 \times 10^{-6} \frac{\sqrt{1 + \mu}\,[V_a + \mu V_g]^3}{[\alpha + \mu g]^2}$$

Like Fremlin's result this expression is correct when μ tends to zero or infinity. In practice μ is rarely less than 3 and so the departure of $\sqrt{\dfrac{\mu}{1 + \mu}}$ from unity is small.

One point in favour is that the 'electrostatic' μ is used and the formula is restricted to the range in which the true μ and the electrostatic μ coincide.

The principal difficulty in obtaining experimental verification of these formulae is caused by the neglect of emission velocity. It is to be noted, however, that the equivalent diode as deduced by the last-mentioned method is unique whatever the emission velocity may be.

Acknowledgment

The author wishes to express his indebtedness to the Management of the Mullard Radio Valve Company Limited for permission to publish this work.

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DESIGN OF CONSTANT IMPEDANCE EQUALIZERS*

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SUMMARY.—Some useful properties of inverse networks (as used in line equalization) are deduced and applied to the development of simple practical design procedures involving no calculations when suitable test equipment is available.

1. Introduction

THE method suggested in this article applies to all bridged-T, ⌉ and L networks fulfilling the condition that $a \times b = Z_0^2$ where a and b are the impedances designated as such in Figs. 1(a), (b), and (c), and Z_0 is the iterative impedance of the respective network. So far as the bridged-T network is concerned a modified form is sometimes employed where the branch impedances shown as Z_0 in Fig. 1(a) appear modified by the factor $1/c$. A network of this kind [shown in Fig. 1(d)] cannot be treated *directly* by the method to be developed. It will, however, be shown in the appendix that this network is obtained from the one of Fig. 1(a), for which $c = 1$, by a simple star-mesh transformation.

Lattice networks can be treated by a modified version of the suggested design method, which will also be discussed.

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2. Theoretical Basis of Design

The theoretical basis of the design can be represented in the form of three propositions, for which proofs will be given under 2.2.

2.1. *The three principles of the design*

I(a). A bridged-T network with $c = 1$ can be converted into an ⌉ network of the same iterative transfer coefficient and the same iterative impedance by short circuiting the Z_0 arm next to the output terminals of the network.

I(b). A bridged-T network with $c = 1$ can be converted into an L network of the same iterative transfer coefficient and the same iterative impedance by disconnecting the Z_0 arm next to the output terminals of the network.

These two rules obviate the necessity of treating the three networks separately. The ⌉ and the L network are simply viewed as derivatives of the bridged-T network.

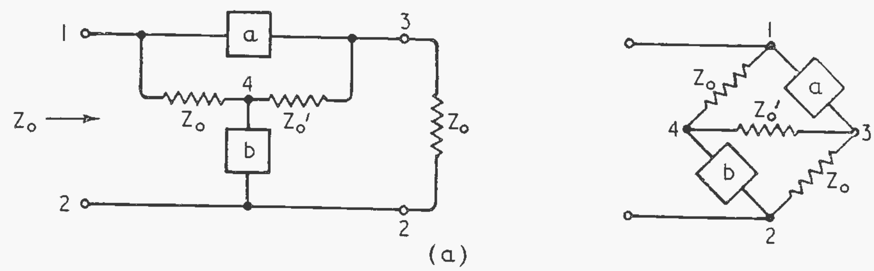
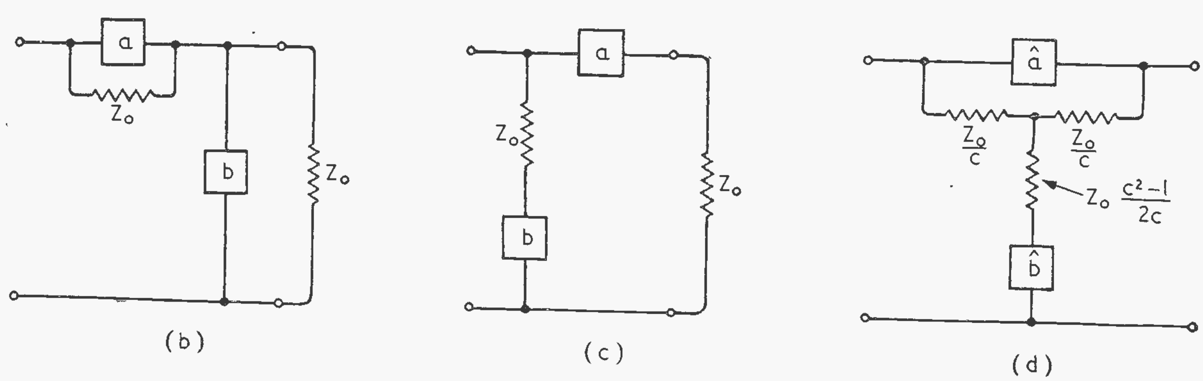


Fig. 1. *Bridged-T and the equivalent bridge networks are shown at (a), with ⌉ and L networks at (b) and (c). A modified form of (a) appears at (d).*



II(a). If the shunt arm of a bridged-T network with $c = 1$ is connected in series with an impedance equal to the iterative impedance of the network, the natural logarithm of the vector ratio of the voltage across both, to the voltage across the shunt arm is equal to the iterative transfer coefficient of the whole network.

The following is a dualistic restatement of the above and is only given for the sake of completeness.

If the shunt arm of a bridged-T network with $c = 1$ is connected in parallel with an impedance equal to the iterative impedance of the network, the natural logarithm of the vector ratio of the total current to the current in the iterative impedance is equal to the iterative transfer coefficient of the whole network.

II(b). If the significant series arm [i.e., "a" in Fig. 1(a)] of a bridged-T network with $c = 1$ is connected in series with an impedance, equal to the iterative impedance of the network, the natural logarithm of the vector ratio of the voltage across both to the voltage across the iterative impedance is equal to the iterative transfer coefficient of the whole network.

Restating the above:

If the significant series arm of a bridged-T network with $c = 1$ is connected in parallel with an impedance equal to the iterative impedance of the network the natural logarithm of the vector ratio of the total current to the current in the series arm is equal to the iterative transfer coefficient of the whole network.

The first forms of statements II(a) and II(b) are the basis of the new method.

III. A bridged-T network with $c = 1$ can be converted into a lattice of the same iterative impedance and iterative transfer coefficient if the impedance of its shunt arm is doubled, placed in series with an impedance equal to the iterative impedance and used as one of the four branches of the lattice; the branch adjacent to it is then found by halving the impedance of the significant series arm of the bridged-T network and placing it in parallel with an impedance equal to the iterative impedance.

This rule enables the method to be extended to lattice networks.

2.2. *Proofs for the three principles employed.* *I*(*a*) and *I*(*b*)

The bridged-T network of Fig. 1(a) can be redrawn as a bridge in the manner shown. The condition for balance of this bridge is $a \times b = Z_0^2$; if this condition is fulfilled the current and voltage distribution in the remainder of the network will be unaffected by the value of the "galvo-arm." If the latter is short-circuited Fig. 1(b) results, if open circuited Fig. 1(*c*). The only purpose of the impedance Z_0' [terminals 4 and 3 of Fig. 1(a)] is to make the network symmetrical.

It remains to prove that the condition $a \times b = Z_0^2$ furnishes iterative networks which can be used as equalizers.

Consider Fig. 1(a) and assume terminals 4 and 3 to be open-circuited. This is permissible in the light of the above. The input impedance between terminals 1 and 2 will be

$$Z_{IN} = \frac{(a + Z_0)(b + Z_0)}{a + b + 2Z_0} = \frac{ab + Z_0 b + Z_0 a + Z_0^2}{a + b + 2Z_0}$$

but $ab = Z_0^2$;

$$Z_{IN} = \frac{Z_0^2 + Z_0 b + Z_0 a + Z_0^2}{2Z_0 + a + b} = \frac{Z_0(2Z_0 + a + b)}{2Z_0 + a + b} = Z_0$$

This proves that the input impedance of all three networks is the same as the load impedance if $a \times b = Z_0^2$; they are therefore suitable as equalizers. Normally Z_0 will be independent of frequency and purely resistive. The reasoning employed is, however, still valid if Z_0 is a function of frequency and has an angle, so long as the product $a \times b$ is made equal to Z_0^2 at all frequencies.

II(*a*) *and II*(*b*)

In view of what has been said before, it is clear that the iterative transfer coefficient is the same for the three networks. It can easily be found in terms of Z_0 and a or Z_0 and b if, for instance, terminals 4 and 3 Fig. 1(a) are considered open-circuited. The well-known expressions for a and b follow readily and are given here for reference:

$$b = \frac{Z_0}{\epsilon^\theta - 1}; \quad a = Z_0(\epsilon^\theta - 1)$$

where θ is the iterative transfer coefficient of the networks.

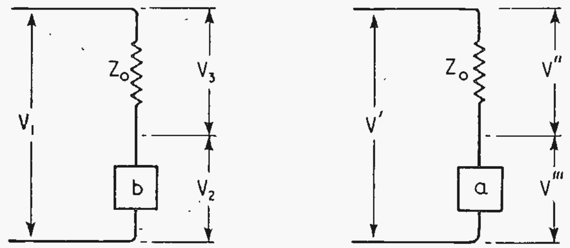
If now branch b is placed in series with Z_0, as shown in Fig. 2(a), the ratio of the applied voltage to the voltage across b is given by :—

$$\frac{V_1}{V_2} = \frac{Z_0 + b}{b} = \left(Z_0 + \frac{Z_0}{\epsilon^\theta - 1}\right)\left(\frac{\epsilon^\theta - 1}{Z_0}\right)$$
$$= \epsilon^\theta = \epsilon^\alpha \angle\beta$$

Similarly for branch a, Fig. 2(b). The ratio of the applied voltage, to the voltage across Z_0 is given by

$$\frac{V'}{V''} = \frac{Z_0 + a}{Z_0} = \frac{Z_0 + Z_0(\epsilon^\theta - 1)}{Z_0}$$
$$= \epsilon^\theta = \epsilon^\alpha \angle\beta$$

where α is the attenuation of the complete network in nepers and β its phase-shift in radians.



(a) (b)

Fig. 2. *Basic circuits used in determining the networks " a " and " b."*

III

The well-known expressions for the two arms of a lattice network are

$Z_1 = Z_0 \tanh \frac{\theta}{2}$ and $Z_2 = Z_0 \coth \frac{\theta}{2}$ where Z_1 and Z_2 are the impedances so designated in Fig. 3.

This can be rewritten as

$$Z_1 = Z_0 \frac{\epsilon^\theta - 1}{\epsilon^\theta + 1} \text{ and } Z_2 = Z_0 \frac{\epsilon^\theta + 1}{\epsilon^\theta - 1}$$

It follows from these expressions that Z_1 can be considered as consisting of $Z_0(\epsilon^\theta - 1)$ in parallel with a similar impedance and both in parallel with Z_0.

The proof is simple :

$Z_0(\epsilon^\theta - 1)$ in parallel with $Z_0(\epsilon^\theta - 1)$

$$= \frac{Z_0(\epsilon^\theta - 1)}{2}$$

$\frac{Z_0(\epsilon^\theta - 1)}{2}$ in parallel with Z_0

$$= \frac{Z_0^2(\epsilon^\theta - 1)}{Z_0(\epsilon^\theta - 1) + 2Z_0} = Z_0 \frac{\epsilon^\theta - 1}{\epsilon^\theta + 1};$$

Similarly the branch Z_2 can be considered as consisting of $Z_0 \frac{1}{\epsilon^\theta - 1}$ in series with a similar impedance and both in series with Z_0.

The significant elements of this type of network are therefore equal to $Z_0(\epsilon^\theta - 1)$ and $Z_0 \frac{1}{\epsilon^\theta - 1}$ respectively and can be found by the method discussed (see section 3.3).

3. Practical Design Procedure

3.1. *To meet given attenuation requirements*

Considering Figs. 2(a) and (b) and applying Principles II(a) and (b) we get

$$\frac{V_1}{V_2} = \epsilon^\theta \text{ and } \frac{V'}{V''} = \epsilon^\theta;$$

if only the moduli of the two voltages are of interest we can write :

$$\frac{|V_1|}{|V_2|} = \epsilon^\alpha \text{ and } \frac{|V'|}{|V''|} = \epsilon^\alpha.$$

If actual measurements are carried out the voltmeter used will, of course, ignore the angles. If instead of a voltmeter a transmission measuring set if employed the readings corresponding to V_1 and V_2 will be $M = \log_e \frac{|V_1|}{|V_R|}$ and $m = \log_e \frac{|V_2|}{|V_R|}$ respectively, if V_R is the reference voltage and the set is calibrated in nepers.

The difference between the two readings is then :

$$M - m = \log_e \frac{|V_1|}{|V_R|} - \log \frac{|V_2|}{|V_R|}$$
$$= \log_e \frac{|V_1|}{|V_2|} = \log_e \epsilon^\alpha = \alpha$$

In other words the difference between the two readings gives the attenuation of the bridged-T or of the L network of which b forms a part, in the units in which the set is calibrated. Similar considerations hold for branch a and the voltages V' and V''.

This enables us to find the component values of, say, branch b experimentally. All

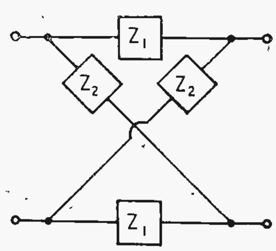


Fig. 3. *Basic lattice network.*

that is necessary is to adjust the components until $M - m$ is equal to the desired attenuation of the complete network at all frequencies of importance. The branch a is then found by carrying out a similar adjustment, or, for preference, simply by making it inverse to b with respect to R_0, following the well-known rules for networks which are inverse with respect to a constant resistance.

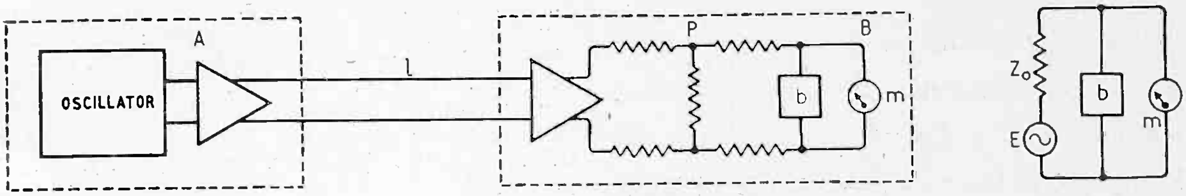
As an example, the detailed procedure for the adjustment of an equalizer suitable for a lump-loaded line is discussed in Appendix II.

3.2. *To equalize a given line*

If the line for which the equalizer is intended is accessible, a more elegant way of carrying out these adjustments is possible and was suggested to the author by Mr. H. J. Orchard, of the Engineer-in-Chief's School, P.O. Eng. Dept.

In practice it is more convenient to employ the circuit of Fig. 5 which enables us to find the arms of the lattice directly. It has been shown that the arm Z_2 of the lattice of Fig. 3 $\left(\text{equal to } Z_0 \coth \frac{\theta}{2}\right)$ can be considered as consisting of an impedance equal to Z_0 in series with an impedance $q = 2 \times Z_0 \frac{1}{\epsilon^{\theta} - 1}$.

Applying the basic idea of Proposition II to this circuit it follows that $\left|\frac{V^{x}}{V^{xx}}\right| = \epsilon^{\alpha}$,



(a) (b)

Fig. 4. *Circuit in (a) shows method of determining "b" arm of equalizer for line "l" which is available for test. Equivalent electrical circuit is shown in (b).*

The circuit of Fig. 4(a) is set up, where l is the line for which the equalizer is to be adjusted and P is an attenuator of say, 10 db, to mask any deviations of the output impedance of the amplifier from its design impedance.

Provided the design impedances of the attenuator and associated amplifier are equal to the iterative impedance of the equalizer, the equivalent circuit of this arrangement, shown in Fig. 4(b), is identical with the circuit previously used.

Inspection of Figs. 4(a) and (b) shows that E [the equivalent e.m.f. in Fig. 4(b)] when expressed in db will vary with frequency by exactly the same amount as that by which A (the line attenuation) varies with frequency, if a constant input is applied to the line and the amplifier gain is assumed to be flat with frequency.

Since E in db corresponds to M in section 3.1 and in Appendix II, it follows from the results obtained in this section that the components in b should be adjusted so that m [the reading of a db meter, Fig. 4(a)] is the same at all frequencies and equal to the value obtained at the highest frequency.

3.3. *Lattice networks*

Principle III points the way to an experimental design for lattice networks. One could, of course, simply find the arms a and b of a bridged-T network having the same iterative parameters as the desired lattice and convert to the lattice afterwards.

where α is the attenuation of the complete lattice, V^{x} is the voltage applied to this experimental circuit and V^{xx}, the voltage across q, is the "significant" part of the lattice arm. Therefore q can be found experimentally in the usual manner and is then placed in series with an impedance equal to Z_0 to furnish the complete arm of the desired lattice. The other arm of the lattice can be found in an analogous manner or better by reciprocation.

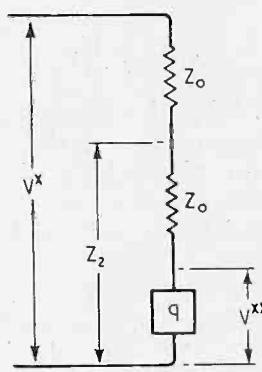


Fig. 5. *Circuit used in finding the arms of a lattice network.*

3.4. *Inverse networks not of constant-resistance.*

Networks of the bridged-T and L type whose impedance is not constant with frequency can be designed by adapting the above method.

It will be necessary to design a two-terminal network first, whose impedance-frequency response follows the desired law. This network is then used in place of R_0 either in actual experiment or for theoretical investigations. Let the impedance of this network $= Z_0(\omega)$. The branches a and b will now be inverse only in the wider meaning of the word; i.e., the product $a \times b$ will still be equal to Z_0^2, but as Z_0^2 is not constant with frequency and may have an angle, reciprocation cannot be carried out by following the simple rules which apply to constant-resistance networks. Two ways out of the difficulty offer themselves.

One can either find both networks a and b separately by means of a method similar to the one discussed, or else find one network first and determine the other by means of the bridge circuit of Fig. 6.

In the first case one network, say b, is obtained by finding a suitable configuration and its component values such that $\frac{|V_1|}{|V_2|}$ [Fig. 2(a)] follows the desired law, a record also being kept of the ratio $\frac{|V_1|}{|V_3|}$ at the different frequencies.

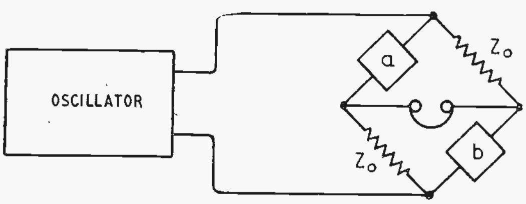


Fig. 6. *Bridge circuit used in the determination of networks which are not of constant-resistance type.*

But now $\frac{V_1}{V_2}$ must be equal to $\frac{V'}{V''}$ [Fig. 2(b)] both with respect to modulus and angle. The angles of the voltage ratios represent the phase-shift of the ultimate network. We are no longer dealing with constant-resistance networks, where the angle of a need not be considered, as it automatically becomes equal and opposite to the angle of b by the process of reciprocation. Having kept a record of $\frac{|V_1|}{|V_3|}$ one could now find a configuration for a (which will not simply be inverse to b in the narrower sense of the word) and adjust its components until $\frac{|V'|}{|V''|} = \frac{|V_1|}{|V_2|}$ and $\frac{|V'|}{|V'''|} = \frac{|V_1|}{|V_3|}$ at all frequencies of consequence, this will make the angles of the two networks equal, but they may still have the wrong sign, which would have to be decided from inspection.

This method is cumbersome and the use of the bridge circuit of Fig. 6 recommends itself. Once one network, say b, has been found by paying attention to the modulus of $\frac{V_1}{V_2}$ only, the network is inserted in the bridge. The configuration and component values of a are then found by adjusting until the bridge is balanced over the prescribed range; $a \times b$ will then be equal to Z_0^2 at those frequencies.

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APPENDIX I

1. *"Derived" bridged-T networks*

The completed branch b of a bridged-T network ($c = 1$) will normally contain a series resistance R which, in conjunction with an inverse shunt resistance S of branch a determines the basic loss of the network. The *remainders* of the branches a and b are shown as $\hat{a}$ and $\hat{b}$ [see Fig. 7(a).]

In order to save one resistor a transformation from mesh to star can be carried out according to Fig. 7(b), where k, k' and p are the branches of a star replacing the mesh consisting of S, R_0 and R_0'. Resistor p is of course combined with R in the practical execution of the network as in Fig. 7(c). The values for k and p follow from the star-mesh transformation theorem and are given by :—

$$k = k' = \frac{R_0 S}{2R_0 + S}\,;\; p = \frac{R_0^2}{2R_0 + S} = \frac{S}{R_0}\,k\,;$$

Resistance k is identical with $\frac{Z_0}{c}$ of the bridged-T network shown in Fig. 1(d) and mentioned in the introduction to this article, while $p + R$ constitute the resistance $Z_0 \frac{c^2 - 1}{2c}$. The proof of this will be given later.

A bridged-T network with $c > 1$ is therefore obtained by first designing a bridged-T network with $c = 1$, subsequently using the above star-mesh transformation.

It is equally permissible to carry out a transformation from star to mesh, replacing the star formed by R_0, R_0' and R of Fig. 7(a) by a mesh. Combining the horizontal branch of this mesh with S furnishes another derived network with a saving of one resistor. There seems, however, to be no advantage in this method over the transformation from mesh to star.

Perhaps it should be mentioned that, although the derived bridged-T networks have the same iterative parameters as their prototypes, it is not permissible to short-circuit or open-circuit k' in order to obtain an ⌉ or L network in the manner in which Z' was treated in the case of the basic bridged-T network.

2. *Connection between conventions used under* 1 *and notation employing the factor c.*

By definition let $k = \dfrac{R_0}{c}$; then $c = \dfrac{R_0}{k}$

but $k = \dfrac{R_0 S}{2R_0 + S}$; $\therefore c = \dfrac{2R_0 + S}{S}$;

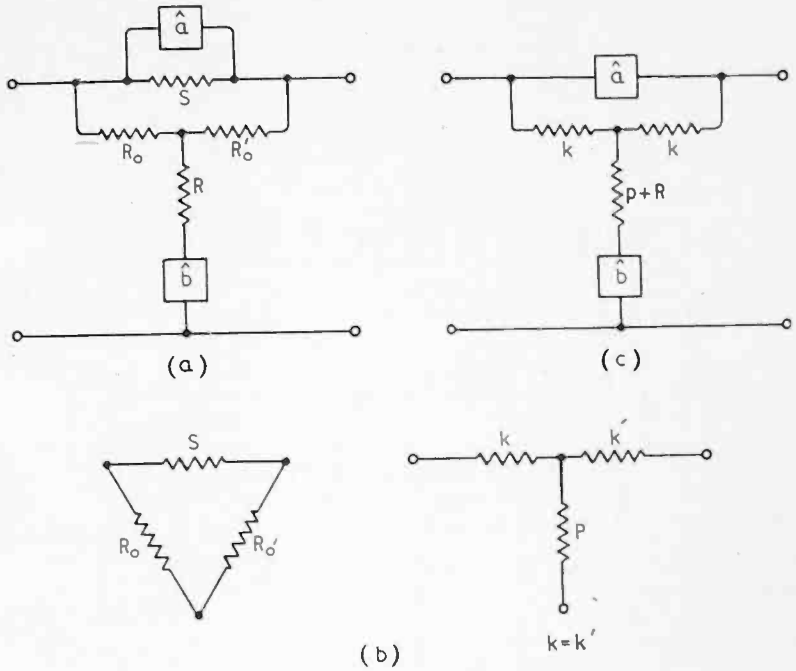


Fig. 7. *In the derived bridged-T network a star-mesh transformation is used to save a component.*

It will be noted that this explains why passive networks with $c < 1$ are physically impossible.

Now :— $p + R = \dfrac{R_0^2}{2R_0 + S} + R$;

but $RS = R_0^2$ and $R = \dfrac{R_0^2}{S}$

$$\therefore p + R = \frac{R_0^2}{2R_0 + S} + \frac{R_0^2}{S}$$

$$= R_0 \frac{2R_0S + 2R_0^2}{2R_0S + S^2} = R_0 \frac{4R_0^2 + 4R_0S}{4R_0S + 2S^2}$$

$$= R_0 \frac{(2R_0 + S)^2 - S^2}{2(2R_0 + S)S} = R_0 \frac{\dfrac{(2R_0+S)^2}{S^2} - 1}{2\dfrac{2R_0 + S}{S}}$$

$$\text{but } c = \frac{2R_0 + S}{S}$$

$$\therefore p + R = R_0 \cdot \frac{c^2 - 1}{2c}$$

This establishes the identities mentioned under (1) of this Appendix.

APPENDIX II

Detailed Design Procedure of Bridged-T Equalizer suitable for a Lump-loaded Line

The design is commenced by concentrating on branch b [Figs. 1(a), (b) and (c)].

For the adjustment of b the circuit of Fig. 8 is set up, consisting of a variable frequency oscillator S, a level measuring set, in "through level" position U* a resistance R_0 equal to the design impedance of the equalizer and variable components L, C, and R forming the branch b of the equalizer under consideration.

The dashes in the subsequent expressions indicate conditions at different frequencies, one dash for the values at the highest frequency of the band considered, two dashes for the lowest frequency.

Let A be the attenuation, at any frequency, of the line for which the equalizer is to be designed.

Let M be the voltage, expressed in db, applied to the experimental circuit, at any frequency, as indicated by the dashes.

Let m be the voltage, expressed in db, measured across the branch b of the network under construction, at any frequency, as indicated by the dashes.

Let n be the desired attenuation of the network at any frequency, as indicated by the dashes.

Normally A will be greatest at the highest frequency considered. The equalizer network will, however, have a "residual attenuation" of n' at this frequency. The equalizer must be so designed that the total attenuation of the line plus equalizer is $A' + n'$ at all relevant frequencies.

First the frequency of the oscillator is set to the highest frequency considered and the tuned circuit adjusted until $M' - m'$ is a minimum. This furnishes $n' = M' - m'$, the value of R being of no practical consequence at this stage.

The oscillator is now set to the lowest frequency

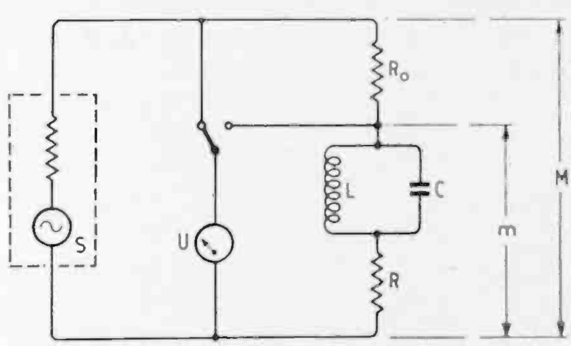


Fig. 8. *Circuit used in determining the constants of an equalizer for a lump-loaded line.*

considered; the attenuation n'' required at this frequency is

$$(A' + n') - A'' = M'' - m''$$

* A level measuring set is said to be in "through level" position if the conditions set up in the instrument are such that the impedance between its terminals can be considered as infinite.

R is therefore adjusted until $M'' - m'' = (A' + n') - A''$.

The correct attenuation at intermediate frequencies is obtained by altering the ratio $\frac{L}{C}$ without altering $L \times C$ in such a way that $M - m = (A' + n') - A$ within the required tolerance at all relevant frequencies.

As M is adjustable at will the work can be made easier by keeping M to zero db at all frequencies. In this case the above expression becomes :—

$$- m = (A' + n') - A$$

In other words the components of b are then adjusted in such a way that m is numerically equal to the desired attenuation at all frequencies.

Perhaps it is still speedier to make M negative and numerically equal to A at all frequencies, then

$$- A - m \doteq (A' + n') - A$$

$$\therefore m = - (A' + n') = \text{a constant.}$$

This means that all that is necessary in this case is to adjust the tuned circuit at the highest frequency until m' becomes a (negative) minimum, and then carry out all subsequent adjustments in such a way that m remains constant within the tolerances allowed.

CLASS B AUDIO-FREQUENCY AMPLIFIERS*

By F. Butler, B.Sc., A.M.I.E.E.

SUMMARY.—Distortion caused by the variable input impedance of an amplifier in which grid current flows during a part of each cycle of the driving voltage is reduced by the use of an output stage having an input impedance which is very low at all values of signal amplitude. In this case, the additional parallel loading represented by grid-current flow is comparatively small.

An earthed-grid, cathode-coupled amplifier has the desired characteristics and has the further advantage that although it requires considerable excitation power, a large proportion of this appears as useful output. The division of power between the final and penultimate stages is of special importance in the design of high-power modulators, in which it may be difficult to secure the desired output from a single stage.

1. Introduction

At audio frequencies, the input impedance of a conventional triode amplifier is almost infinite, except when the amplitude of the excitation voltage exceeds the standing negative grid bias. The intermittent flow of grid current then constitutes a variable loading on the driver valve, and unless the voltage regulation of this stage is exceptionally good, amplitude distortion becomes serious, due to the clipping of the crests of the positive half-cycles of the driving voltage. Abrupt changes in input impedance also tend to cause undesirable transient oscillations, or "ringing", often difficult to suppress.

Negative feedback can be used to lower the output impedance of the driver valve, at the expense of a corresponding reduction in stage gain. Alternatively a step-down driver transformer may be employed, its turns ratio being subject to the restriction that adequate excitation for the output valves must be available without excessive distortion in the penultimate stage. For the best results, the output power from the driver must be greatly in excess of the minimum excitation power requirements of the final amplifier, the balance being dissipated in a dummy load.

An examination of the properties of the earthed-grid, or cathode-coupled, amplifier shows that it is almost ideally suited to operation under Class B conditions. In the first place, its mean input impedance is very low at all values of signal level, and, in comparison, the maximum variations due to intermittent grid current flow are negligible. Moreover, the high excitation power required is not entirely dissipated as in the conventional system. By proper design, a large fraction of the driver output can be transferred to the final stage. In the case of high-power amplifiers or modulators, the contribution from the driver reduces the requirements from the final stage and thus simplifies design.

2. Cathode-input Amplifier

Fig. 1 shows a triode driver V_1 coupled to an earthed-grid amplifier V_2. Replacing the actual valves by their equivalent generator

*MS. accepted by the Editor, May 1946.

circuits and restricting the analysis to linear conditions, equations may be derived for the overall stage-gain, the final power output and the requisite output from the driver stage. The case of impedance coupling will first be discussed and the necessary modifications for ideal transformer coupling will then be given.

Replacing the actual circuit of Fig. 1 by its electrical equivalent, shown in Fig. 2, the network equations may be derived.

Let E = input voltage to V_1.
E_1 = equivalent generator e.m.f. of V_1, $= \mu_0 E$.
i_1, i_2 = mesh currents in the network.
R_0 = anode slope resistance of V_1.
μ_0 = amplification factor of V_1.
z = effective anode load impedance of V_1, excluding the input impedance of V_2.
= z_1 and z_2 in parallel.
R_L = load resistance of V_2.
R_a = slope resistance of V_2.
μ = amplification factor of V_2.
$R_2 = R_L + R_a$.
E_c = grid-cathode potential of V_2.
= voltage drop across the impedance z.
$E_2 = \mu E_c$.
E_0 = final output voltage.

Applying Kirchhoff's laws, it can be shown that the ratio of the output voltage E_0 to the equivalent generator voltage E_1 of the driver valve is given by :

$$\frac{E_0}{E_1} = \frac{z}{R_0 + z} \cdot \frac{(\mu + 1) R_L}{R_L + R_a + (\mu + 1)\dfrac{R_0 z}{R_0 + z}} \qquad (1)$$

The overall stage gain E_0/E is obtained by substituting, in equation (1), $E_1 = \mu_0 E$.

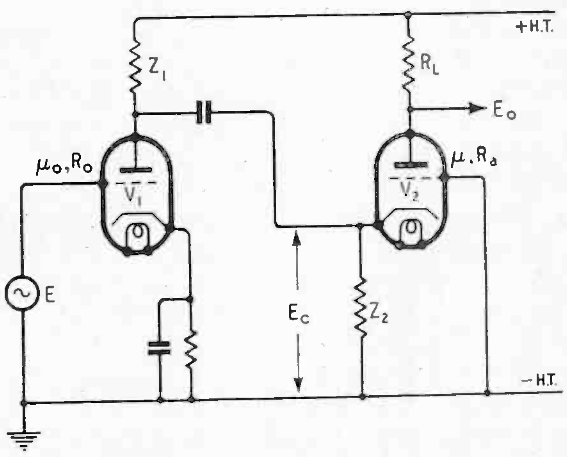


Fig. 1. *Basic circuit of a cathode-input stage and its driver.*

The terms of this equation have been grouped in such a manner as to stress the physical significance of each. First, there is a reduction of gain below the theoretical maximum, due to the finite value of the load impedance z, the extent of the loss being settled by the relative magnitudes of load and valve impedance in the first amplifier stage. The term $(\mu + 1)\dfrac{R_0 z}{R_0 + z}$ represents an increase of effective valve resistance caused by impedance in the cathode circuit of the final stage. The increase is due to a form of negative feedback similar to that which

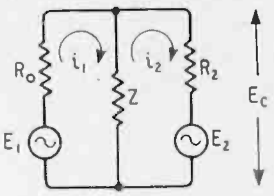


Fig. 2. *Equivalent circuit of the amplifier of Fig. 1.*

is observed in the case of an amplifier with an unshunted cathode-bias resistance. The impedance $\dfrac{R_0 z}{R_0 + z}$ is the parallel combination of the driver-valve resistance and its load impedance, and is the effective output impedance viewed from the cathode of the driven stage. With a fixed value assigned to the load resistance R_L, the maximum gain is obtained when z, the parallel combination of z_1 and z_2, is infinite. For all practical purposes this assumption is permissible if z_1 and z_2 are iron-cored chokes of sufficiently high inductance.

Reference to Fig. 2 shows that, in this special case, the two valves are effectively in series and the currents i_1 and i_2 become identical.

Setting $i_1 = i_2 = i$, with $z = \infty$, the output voltage $E_0 = iR_L$ and equation (1) may be re-written :

$$E_1/i = Z_i = R_0 + \frac{R_a + R_L}{\mu + 1} \quad .. \quad (2)$$

E_1 is the equivalent generator voltage and R_0 the internal output impedance of V_1. The second term of equation (2) is clearly the input impedance of V_2, which forms the load impedance of V_1. The practical design of an amplifier of this kind then reduces to the problem of securing a proper match between the generator impedance R_0 and the load impedance $\dfrac{R_a + R_L}{\mu + 1}$ and to the proper choice of the final load impedance R_L. The latter is selected according to well-established rules and, for standard valves, may be derived from manufacturers' data sheets. There remains the question of matching R_0

to $\dfrac{R_a + R_L}{\mu + 1}$. In general, direct coupling will be found impracticable and a matching transformer must be used. The insertion of an ideal transformer of turns ratio n between the valves converts the load impedance to the value $Z = n^2 . \dfrac{R_a + R_L}{\mu + 1}$.

The exact theory of transformer coupling is very complex but subject to the condition that the primary inductance is sufficiently large to make its reactance, at the lowest significant frequency, of the order of ten times the first valve slope resistance, there is little loss of accuracy in considering only the ideal case.

3. Cathode-Follower Driver Valve

It has already been stated that the relative values of the resistances R_0 and $\dfrac{R_a + R_L}{\mu + 1}$ are in practice such as to preclude the possibility of direct coupling.

It is of interest to consider the use of a cathode-follower as a driver valve. The circuit is shown in a schematic form in Fig. 3.

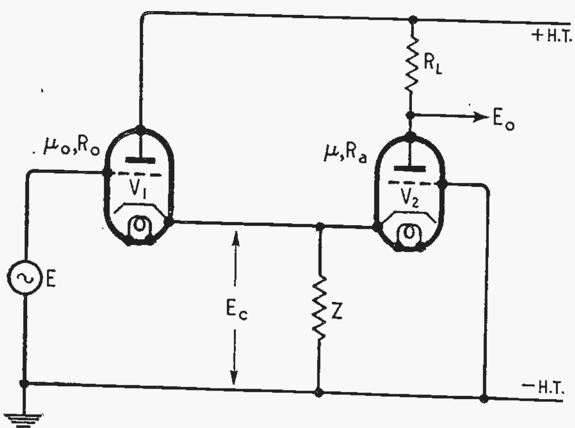


Fig. 3. *Cathode-input amplifier driven by a cathode-follower.*

Normally Z will be an iron-core inductance, and instead of R_L, an output transformer will be used. The modifications for push-pull operation simply involve duplication of the valves and the use of centre-tapped chokes and transformers.

The theory already given is sufficiently general to cover the present case. The cathode-follower valve can be replaced by an electrical equivalent. If μ_0, R_0 are respectively the amplification factor and slope resistance of the valve and if E is the input voltage, the equivalent generator has an e.m.f. $\mu_0/(\mu_0 + 1) . E$ and an internal resistance $R_0/(\mu_0 + 1)$.

Substituting these values in equation (1), the stage gain is given by :—

$$\frac{E_0}{E} = \frac{\dfrac{\mu_0}{\mu_0 + 1} . Z}{\dfrac{R_0}{\mu_0 + 1} + Z} . \frac{(\mu + 1) R_L}{R_L + R_a + (\mu + 1) \dfrac{\dfrac{R_0 Z}{\mu_0 + 1}}{\dfrac{R_0}{\mu_0 + 1} + Z}} \quad \ldots \quad \ldots \quad (3)$$

Assuming μ_0 and $\mu \gg 1$ and setting $Z \to \infty$, equation (3) can be simplified and becomes: $E_0/E = \dfrac{\mu R_L}{R_L + R_a + \mu/g_m}$, where g_m is the mutual conductance of V_1.

The use of a cathode-follower avoids the necessity for a coupling transformer and a centre-tapped choke may be substituted. The greatest disadvantage is the low gain, which calls for the use of additional pre-amplifier stages. It must also be remembered that the driver and the driven stages are effectively in series and therefore the alternating component of anode current is the same in each. This consideration limits the choice of suitable driver valves. The foregoing remarks are restricted to the use of valves under linear, Class A conditions. Modifications must be introduced to cover the case of Class B operation. These will now be discussed in empirical terms.

4. Design of Class B Cathode—Coupled Amplifiers

The successive steps to be performed in the design of a push-pull Class B cathode-coupled amplifier are as follows. The remarks apply to the use of triodes, using interstage transformer coupling. This case is of more general interest than that of the cathode-follower.

(i) Select a pair of valves to give the desired output from the final stage under normal Class B conditions.

(ii) Consult valve tables and determine the anode-to-anode load resistance, the h.t. voltage and negative grid bias.

(iii) Calculate the input impedance of the output valves, $\frac{R_a + R_L}{\mu + 1}$, modifying the result to allow for push-pull Class B operation.

(iv) Determine the r.m.s. excitation voltage required to load the output valves. This information can also be obtained from valve data sheets.

(v) Calculate the maximum-signal driving power required from the penultimate stage to operate the output valves, using the known r.m.s. value of the driving voltage and the known input impedance.

(vi) Choose a pair of driver valves capable of supplying this power with an adequate reserve.

(vii) From valve tables, find the proper load impedance for the driver valves.

(viii) Select the turns ratio and power-handling capacity of the coupling transformer required to convert the input impedance of the output valves to the load value required by the penultimate stage.

In view of the fact that published valve characteristics are the results of static observations, it is preferable to make measurements of the input impedance under working conditions instead of performing the calculation shown in para. (iii) above. This is because the separate valves of a Class B stage are alternately active and quiescent, which complicates the analysis.

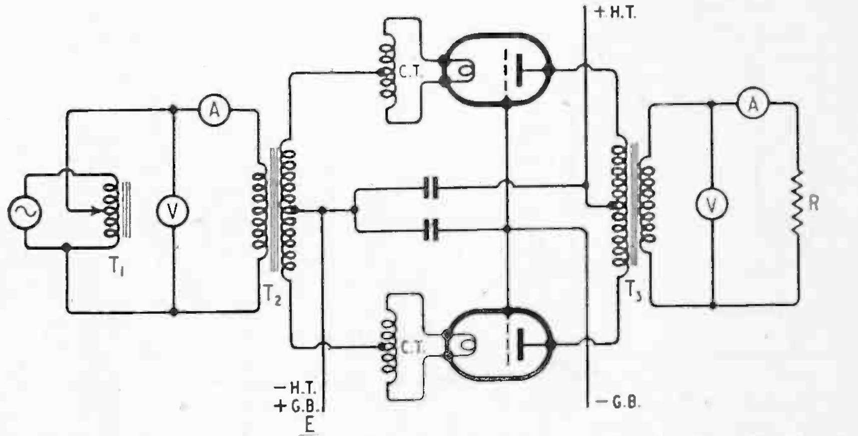


Fig. 4. *Circuit for the measurement of the input impedance of a Class B stage.*

5. Measurement of Input Impedance

A circuit suitable for the measurement of the input impedance of the output valves is shown in Fig. 4, and it is convenient to make measurements at 50 c/s. A "Variac" transformer T_1 is connected to the a.c. mains and its variable output applied to the primary of a coupling transformer T_2, of turns ratio $n : 1$. The secondary centre-tap is earthed and the outer ends of the winding connected respectively to the centre-taps of the separate filament transformer windings which supply the two output valves. Rectifier voltmeters and ammeters permit the measurement of input and output voltages and currents. Exceptionally good regulation of h.t. and grid bias supplies is essential, and impedance measurements should be made over a wide range of input signal level. It will be found that there is a definite drop in the input impedance once the peak driving voltage exceeds the fixed negative bias, but the percentage change remains relatively small, unless the valves are dangerously over-driven.

Let E = measured input voltage to T_2.
I = observed primary current.

Then, input impedance $Z_i = E/I$.

The actual input impedance between the valve cathodes, taking account of the effect of the coupling transformer, is :—

$$Z = n^2 Z_i$$

Once the value of Z is determined, either by measurement or by calculation, the correct ratio of the coupling transformer required to match the driver valves to the input impedance of the final stage is given by :—

$$n = \sqrt{R/Z}$$

where R is the optimum load of the driver stage. At first sight it might seem preferable to make direct input-impedance measurements on the secondary side of the coupling transformer. By using the method described, the transformer losses are included in the measurement and a more realistic figure for the impedance is obtained. If possible, the test transformer should be physically similar to the component which will eventually be employed as a driver.

6. Experimental Results

Using the circuit of Fig. 4, results are obtained which support the theoretical conclusions and indicate that the cathode-coupled audio-frequency amplifier has useful properties not possessed by normal Class B amplifiers. In making such measurements, the experimental procedure is as follows:—

(i) From valve data sheets select a pair of valves capable of delivering the desired output and choose an appropriate load resistance, allowing for the turns ratio of the output transformer.

(ii) Using a double-beam oscilloscope, display the input voltage to the driver transformer on one trace and the amplifier output voltage on the other.

(iii) Adjust the input and output signal levels to equality and bias the deflector plates until the two displays can be superimposed. Auxiliary potentiometers used for this purpose must be of high resistance in comparison with the load resistance and with the input impedance of the driver transformer. Alternatively, allowance must be made for power dissipated in them.

(iv) Raise the input signal level until the output shows signs of distortion. This is recognised by the fact that the two traces can no longer be brought into coincidence, and is an indication that the maximum useful power output has been attained.

(v) Record the input and output voltages and currents. Note the h.t. supply voltage and feed current.

Results are recorded below for the case of a particular amplifier using 25-watt triodes in the output stage. The valves are apparently being over-run, but although the h.t. voltage is 50 per cent above the makers' rating, the maximum permissible anode dissipation is not exceeded.

Type of valve: Marconi-Osram PX25
H.T. Voltage: 600 V
Feed Current: 170 mA
D.C. Grid Current: 3 mA
Input Power: 102 W
Input Transformer (T_2 in Fig. 4) turns ratio (Primary : Secondary): 1 : 2.
Output Transformer (T_3 in Fig. 4) turns ratio (Primary : Secondary): 2.8 : 1.
Load Resistance: 1,100 Ω
Equivalent load resistance at anodes of output valves:
$(2.8)^2 \times 1100 = 8{,}624\ \Omega$
Output Voltage: 245 V r.m.s.
Output Current: 0.223 A r.m.s.
Output Power (maximum undistorted): 54.63W
Input Voltage: 65 V r.m.s.
Input Current: 0.22 A r.m.s.
Driving Power: 14.3 W
Input Impedance at primary of $T2$: 295.5 Ω
Equivalent Input Impedance (between valve cathodes):
$2^2 \times 295.5 = 1{,}182\ \Omega.$

The high ratio of driving power to the final output power is clearly evident. As already pointed out, a proportion of the driving power of 14.3 W is transferred to the output and contributes to the total of 54.63 watts delivered to the load. The apparent conversion efficiency of the amplifier is 54.63/102 = 53.6 per cent. The actual value is lower than this since the excitation power has not been included in the calculations.

If an attempt is made to increase the output power by raising the excitation voltage, there is a rapid increase in grid current, due first to the increased positive grid voltage, and to the simultaneous reduction of mini-

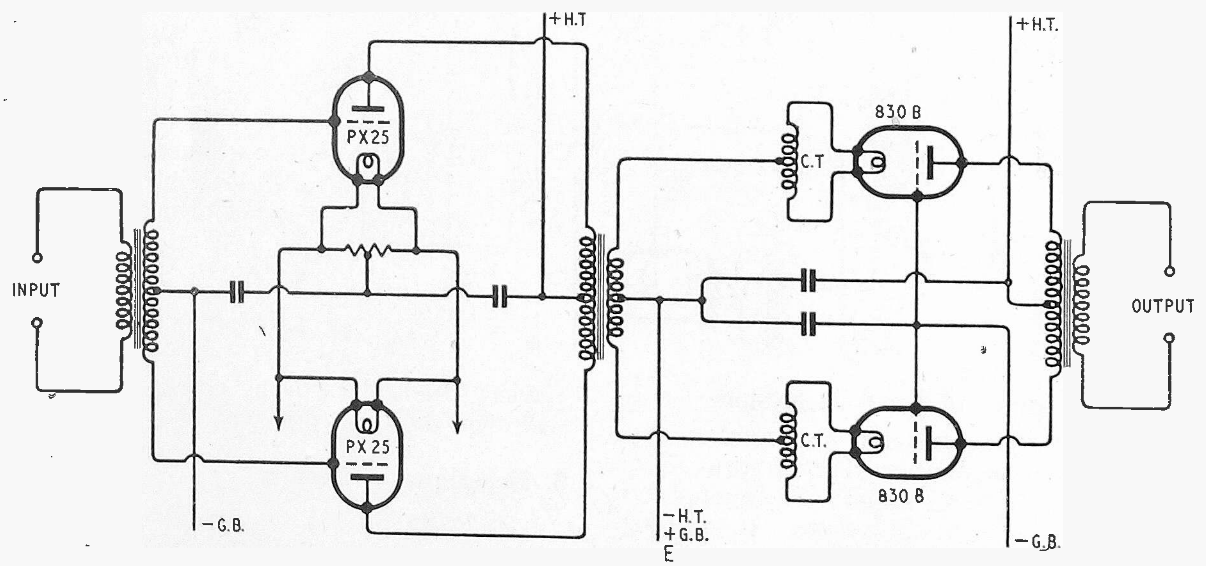


Fig. 5. *Push-pull driver and amplifier stages embodying the principle described.*

imum anode potential due to the larger alternating voltage developed across the load. This is exactly analogous to the case of Class B or Class C radio-frequency amplifiers, in which there is corresponding rapid rise of excitation power once a certain output is exceeded. Correct choice of load resistance minimizes this difficulty at the expense of reduced efficiency.

The above experimental figures do not reveal all the properties of the amplifier. The relationships between driving power and final output, between excitation voltage and grid current, and between input impedance, excitation voltage and load resistance should also be studied. These are straightforward investigations, and will not be described in further detail.

7. Practical Amplifier Circuit

Fig. 5 shows a push-pull driver and output stage employing the principles already described. With the valve types shown, operating at maximum ratings, the output should be nearly 200 watts. Conventional pre-amplifier stages may be added to raise the overall gain to any desired level. If desired, beam-tetrode driver valves may be employed. These have the advantage of higher efficiency, though distortion may be rather greater than with triodes.

To secure a large output with minimum distortion, it is essential that h.t. and grid bias supplies are well regulated. In high-power modulators, supplied from mercury-vapour rectifiers, the voltage regulation is satisfactory. It is usually possible to select gas-discharge regulators for stabilizing the grid-bias supplies of medium power amplifiers. Alternatively, electronic regulators must be used. It is also of importance to keep the driver-transformer secondary windings of low resistance, otherwise an undesired variable cathode bias will be introduced, due to the fluctuating anode current of the output valves.

No reference has been made to the use of negative feedback in connection with cathode-driven amplifiers. Any of the usual systems may be employed, though voltage feedback will normally be preferred, in order to reduce the output impedance of the amplifier. Reference to equation (1) shows that the normal valve resistance R_a is increased to a value :—

$$R_a + (\mu + 1)\frac{R_0 z}{R_0 + z}.$$

In transmitter modulators, this increase is not objectionable, but in public address amplifiers the loudspeaker damping is not quite so good as with the usual Class B system. Voltage feedback can correct this defect and simultaneously reduce the residual distortion.

As regards the response at high audio frequencies, the cathode-driven valve has an inherent advantage over other triode amplifiers, since the earthed grid avoids degenerative feedback due to anode-grid capacitance. To obtain a comparable performance with normal amplifiers, audio-frequency neutralizing would be necessary.

Professor G. W. O. Howe

On his retirement from the James Watt Chair of Electrical Engineering in the University of Glasgow, Professor Howe was presented with a cheque for £200, the presentation being carried out by Professor Bernard Hague.

Professor Howe has handed the gift to the Principal of the University for the endowment of a prize for the most outstanding graduate in electrical engineering.

R. N. Vyvyan

We regret to announce the death of R. N. Vyvyan on December 14th, at the age of 70. Formerly Engineer-in-Chief of the Marconi Company, Mr. Vyvyan was one of the pioneers of wireless. He supervised the erection of the famous Poldhu and Cape Cod stations.

While Managing-Engineer in Canada for the Marconi Company he built the Glace Bay station, which formed the western link of the first regular transatlantic wireless service. He returned to this country in 1908.

He was responsible for the construction and engineering design of the short-wave beam system of the Imperial wireless stations which came into being after the first World War. He retired in 1936.

Popov Medal

Instituted by a decree of the Soviet Government, the Popov Gold Medal is awarded annually on 7th May by the Academy of Sciences of the U.S.S.R. to a scientist of any nationality for outstanding scientific research or invention in the field of radio. The first award will be made in 1947 for work completed in the period 1933–1945 and work submitted must reach the Academy of Sciences Radio Physics and Radio Engineering Council not later than 1st February.*

Papers may be in any language, typed or printed, and there should be three copies. The address of the Council is Tretya Miusskaya d 3, Moscow, U.S.S.R

*Notification of this competition was not received until early December, and the early closing date would seem to make entries from this country difficult.—Ed.

TRANSIENT RESPONSE OF V.F. COUPLINGS*

Compensated Networks for Wide-Band Amplifiers

By W. E. Thomson, M.A.

(P.O. Radio Branch)

SUMMARY.—Formulae and curves are given for the response to the Heaviside unit function of a single resistance-capacitance coupled stage as commonly used in wide-band amplifiers, the nominally resistive load being compensated to offset the low- and high-frequency deficiencies associated with this type of coupling.

The analysis of the low-frequency response deals mainly with the compensation of grid coupling by anode decoupling. Two orders of compensation are dealt with.

For high-frequency compensation, one particular type of compensating circuit is analysed. Attention is confined to conditions which give "critical damping." The ultimate possible improvement with this type of circuit is demonstrated.

CONTENTS

1. Introduction

FOR amplifiers, such as oscillograph and video amplifiers, which must pass a signal with as little change of waveform as possible, the amplitude-frequency and phase-frequency responses do not give an entirely satisfactory criterion of performance. It is preferable, and now fairly general, to take as a criterion of performance the indicial response; that is, the response to a Heaviside unit function.

In this article, several commonly used types of coupling are analysed. The work is of a general nature and includes previous results as particular cases. Attention has been confined to the theoretical aspects, and the numerous practical considerations involved are not discussed.

The usual assumptions have been made. These are:—

(1) The valve concerned can be regarded as a constant-current generator.

(2) The high-frequency response can be considered separately from the low-frequency response. For each case an equivalent circuit is used which neglects certain impedances. As far as the indicial response is concerned, this distinction corresponds to a distinction between the initial build-up period, which is associated with the high-frequency response, and the maintenance of the response at the ideal level up to a certain maximum period, which is associated with the low-frequency response.

(3) The impedance of the d.c. supply is negligible.

The symbols used are either standard or are defined when required; certain symbols, such as α, β, have different meanings in different sections.

2. Low-Frequency Compensation

2.1. *General*

For an amplifier which cannot amplify d.c. the indicial response must tend to zero with increasing time. It is possible, however, to design the amplifier so that the response may approximate the ideal, within given limits, up to a certain maximum time. Let the indicial response be expressed in a Maclaurin series in t:

$$V = 1 + a_1 t + a_2\, t^2/2\,! + a_3\, t^3/3\,! + \ldots$$

* MS. accepted by the Editor, April 1946.

Then the response may be made more and more nearly ideal by choosing the design parameters so that

$$a_1 = a_2 = a_3 = \ldots = a_n = 0$$

for n as large as possible.

The corresponding expression for the operational gain is[1]

$$G = 1 + a_1/p + a_2/p^2 + a_3/p^3 + \ldots$$

The above condition may also be expressed in terms of derivatives, thus :

$$\left[\frac{dV}{dt}\right]_0 = \left[\frac{d^2V}{dt^2}\right]_0 = \ldots = \left[\frac{d^nV}{dt^n}\right]_0 = 0,$$

or

$$\left[\frac{dG}{d(1/p)}\right]_0 = \left[\frac{d^2G}{d(1/p)^2}\right]_0 = \ldots = \left[\frac{d^nG}{d(1/p)^n}\right]_0 = 0$$

In practical work it is usually sufficient to make $a_1 = 0$ only, and the conditions for this can be determined fairly readily from the operational gain. For example the normalized gain for the circuit of Fig. 1 is*

$$G = \frac{1 + Z_d/R_L}{(1 + g_cZ_c + \sigma_sZ_s)\,(1 + 1/pC_gR_g)}$$

where g_c = sum of grid-anode, grid-screen, and screen-anode mutual conductances,

σ_s = screen conductance,

Z_d, Z_c, Z_s are anode decoupling, cathode, and screen networks,

and it is assumed that the coupling capacitor and grid leak can be regarded as of high impedance.

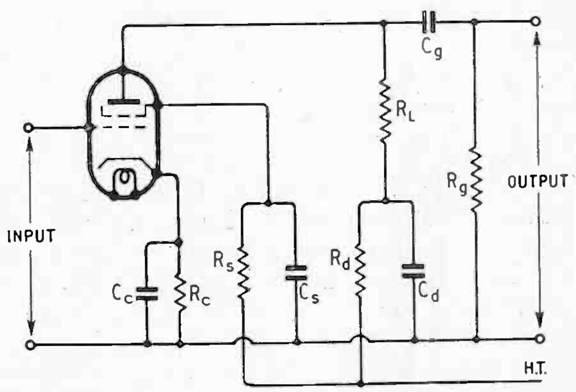


Fig. 1. *Typical resistance-capacitance coupled stage including decoupling, screen-feed, and cathode-bias networks.*

Each of the networks Z_d, Z_c, and Z_s consists effectively of a resistance in parallel with a capacitance. The operational impedance is of the form :

$$\frac{R/pC}{R + 1/pC} = \frac{1/pC}{1 + 1/pCR}$$
$$= (1/pC)\,[1 - 1/pCR + 1/(pCR)^2 \ldots]$$
$$= 1/pC \text{ to a first approximation.}$$

Hence, to a first approximation :

$$G = \frac{1 + 1/pC_dR_L}{(1 + g_c/pC_c + \sigma_s/pC_s)(1 + 1/pC_gR_g)}$$
$$= (1 + 1/p)\,(1/C_dR_L - g_c/C_c - \sigma_s/C_s - 1/C_gR_g)$$

Thus the condition which makes $\left[\frac{dV}{dt}\right]_0 = 0$ is

$$1/C_dR_L = g_c/C_c + \sigma_s/C_s + 1/C_gR_g$$

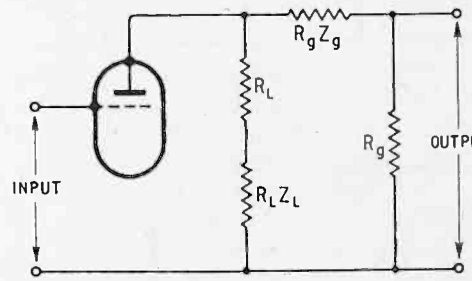


Fig. 2. *The basic coupling circuit ; R_LZ_L and R_gZ_g are networks which respectively do and do not pass direct current.*

2.2.—*Compensation of grid coupling by anode decoupling.*

A case frequently considered is illustrated in Fig. 2 in which R_L and R_g are the anode load and grid leak respectively ; R_LZ_L is a decoupling network which must pass d.c. and R_gZ_g is a coupling network which does not pass d.c. The impedance of both networks tends to zero with increasing frequency. The normalized operational gain of such a stage, assuming that the grid network can be regarded as a high-impedance potential divider across the effective load is :

$$G = \frac{1 + Z_L}{1 + Z_g}$$

Obviously if $Z_L = Z_g$ the gain will be independent of frequency and the indicial response perfect. The d.c. requirements, of course, make this condition impossible. It is possible, however, by making Z_L and Z_g almost equal, to satisfy the conditions stated in the previous section. Z_L and Z_g have the form shown in Fig. 3. These forms automatically fulfil the d.c. and high-frequency conditions stated above. In Fig. 3, Z_L and

---

* Adapted from a formula of Edwards and Cherry.[9]

Z_g have n components identical; in 3(a), Z_g has an extra component C_{n+2} and in 3(b), Z_L has the extra component R_{n+2}. In both cases, as is shown in Appendix 2.1,

$$\left[\frac{dV}{dt}\right]_0 = \left[\frac{d^2V}{dt^2}\right]_0 = \ldots = \left[\frac{d^nV}{dt^n}\right]_0 = 0$$

$$n = 0; \quad V = e^{-t/T}$$

$$n = 1; \quad V = \frac{1}{\alpha - 1}\left(e^{-t/\alpha T} - e^{-t/T}\right)$$

where $T = C_dR_L = C_gR_g$, and α, as is shown in Fig. 4, is the ratio of decoupling to load resistance.

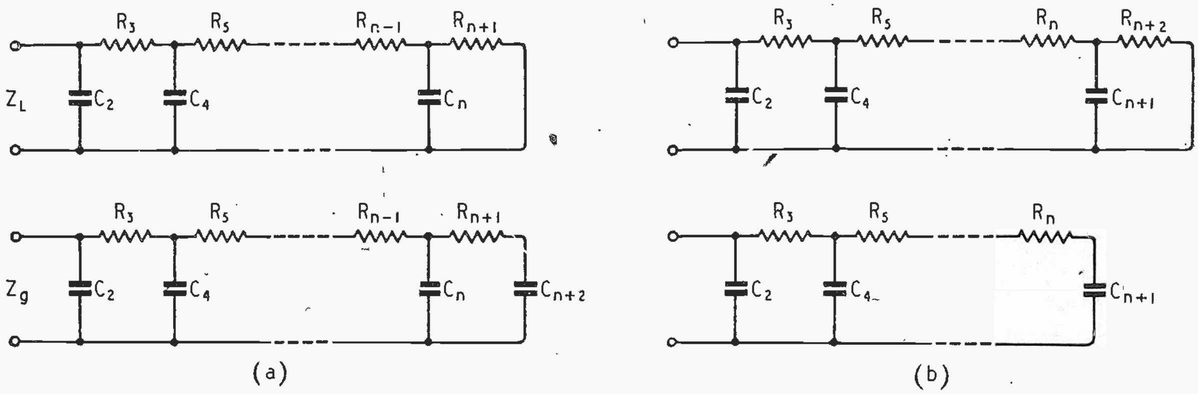


Fig. 3. *General forms of Z_L and Z_g of Fig. 2 are shown here.*

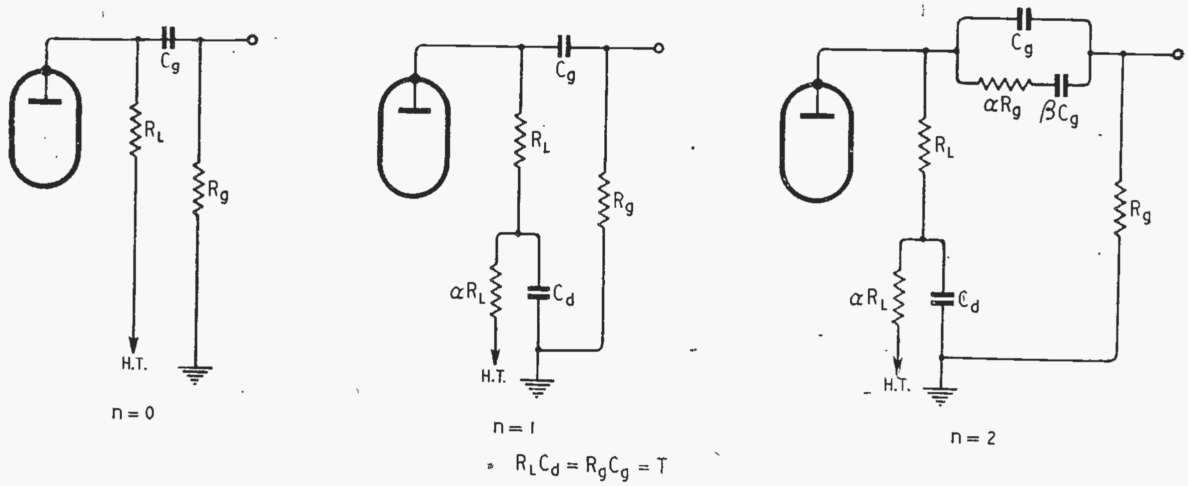


Fig. 4. *The three simplest practical forms of Fig. 2: $n = 0$, uncompensated; $n = 1$, first-order compensation; $n = 2$, second-order compensation.*

Fig. 4 shows the three simplest cases, for $n = 0$, 1, and 2 respectively; the components have been renamed for convenience. Case $n = 0$ is the basic resistance-capacitance coupled stage. For case $n = 1$, the anode load has been modified. The addition of the capacitor C_d alone, with value C_gR_g/R_L would give perfect response; the capacitor must however be bypassed by αR_L to allow h.t. current to flow. The correct relationship $C_dR_L = C_gR_g$ gives an indicial response for which $[dV/dt]_0 = 0$ as is shown in Fig. 5, which gives responses for various values of α; the curve $\alpha = 0$ corresponds to the case $n = 0$. This circuit has been adequately analysed elsewhere [10, 11], so only the formulae are given here;

The case $n = 2$ is not so widely known. This carries the process a stage further by adding components to the grid coupling network which compensates, to some extent, for the resistor αR_L which prevents the response being ideal in the case $n = 1$. Perfect compensation would be achieved by shunting C_g by a resistor of value αR_g, but, in order to block h.t., a capacitor βC_g must be added in series with αR_g. The addition of these two components does not affect $[dV/dt]_0$: provided $C_dR_L = C_gR_g$ we still have $[dV/dt]_0 = 0$. The correct value of resistor, namely αR_g, improves the indicial response still further, by making $[d^2V/dt^2]_0 = 0$. This is shown in Fig. 6 which gives responses for various values of α and β. The curves $\beta = 0$ correspond to the case $n = 1$. The curves also show that it is better to increase α, rather than β. The formula for this case is derived in Appendix 2.2 and given below:

$n = 2$;

$$V = \frac{\alpha\beta}{\alpha - \alpha\beta - 1}\left[\frac{1}{\beta}e^{-t/\alpha T} + \frac{q(1+mq-m)}{m-q}e^{-mt/T} - \frac{m(1+mq-q)}{m-q}e^{-qt/T}\right],$$

where $m = c + d \quad c = \frac{1}{2}(1 + 1/\alpha + 1/\alpha\beta)$

$q = c - d \quad d^2 = c^2 - 1/\alpha\beta$

$T = C_d R_L = C_g R_g$

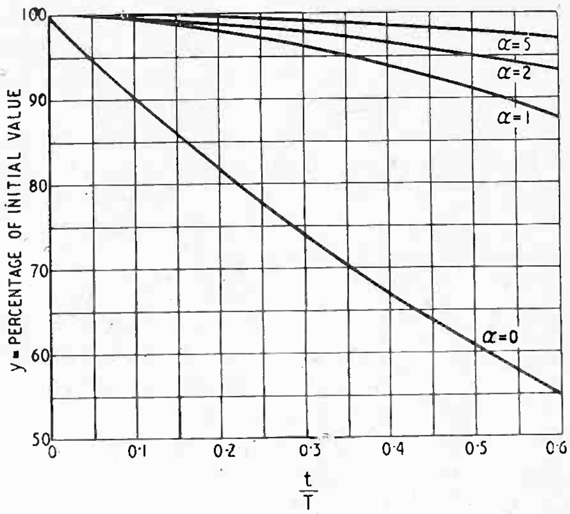


Fig. 5. *Indicial response for first-order l.f. compensation.*

3. High-Frequency Compensation

3.1.—*General*

Of the many types of compensated networks only one is considered here; it is shown in Fig. 7. The nominal load is R_L, and C_s is the total shunt capacitance across the effective load. The compensation is effected by a two-terminal network $R_L Z$ in series with R_L. This network is assumed to be purely reactive for the purposes of design; it may have as many components as desired, each extra component

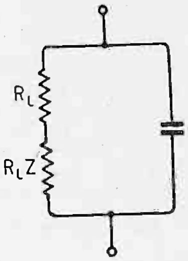


Fig. 7. *Basic network for high-frequency compensation.*

enabling some improvement in response to be obtained. The form of Z must be such that its impedance, as a function of frequency, has a zero at the origin, so that the low-frequency response is unaffected.

The change of shape of the indicial response due to the addition of such compensating networks consists, in general, of reduced build-up time, and oscillation about the steady-state value, this oscillation being generally referred to as "overshoot"* For a given network, improvement in build-up time is obtained at the expense of in-

* In certain types of response, overshoot occurs without oscillation.

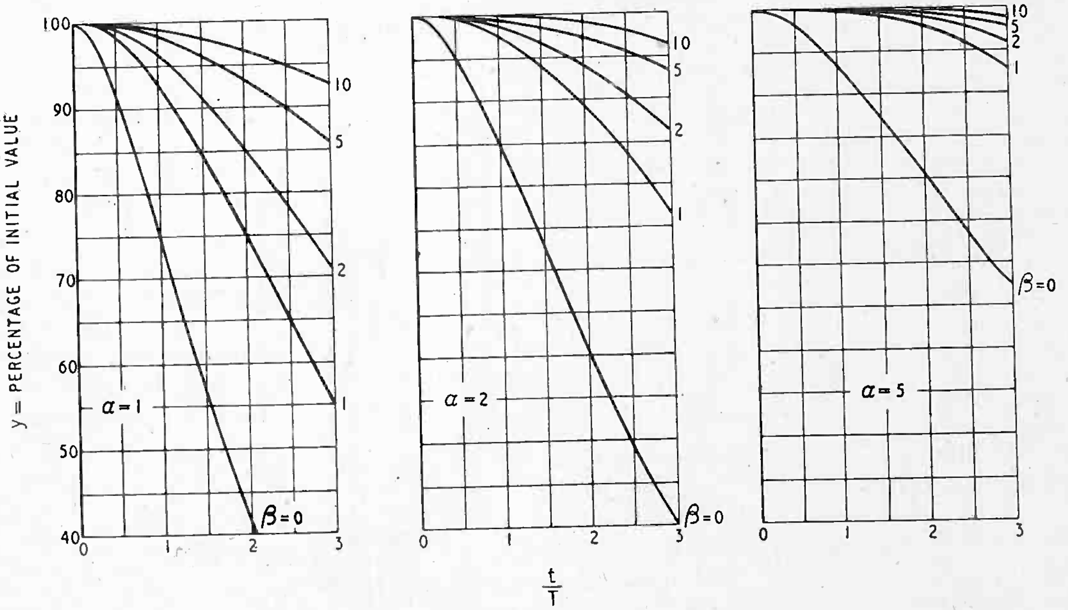


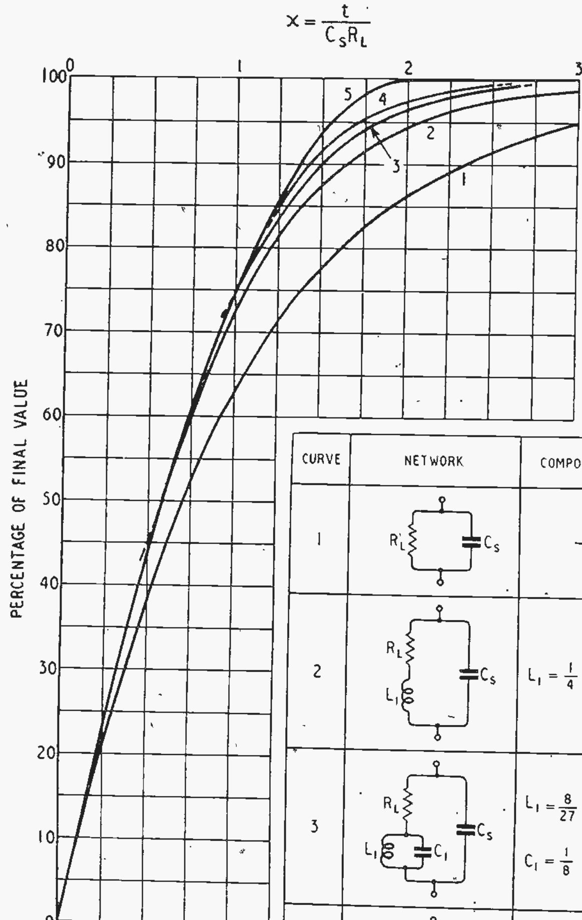
Fig. 6. *Indicial response for second-order l.f. compensation.*

creasing the overshoot once the critical value has been reached.

When Z consists of only a few components, it is possible to make an analysis of all the possible values to determine suitable values for the parameters[4,5,6,7,8,12].

A general treatment requires some simplifying principle. Such a principle is to be found in specifying that the response shall be "critically damped," that is to say the response will be the fastest possible without overshoot.

The conditions are similar to the well-known case of critical damping in a series-tuned circuit where the critical condition is associated with equal real roots of a quadratic equation. Here the nature of the response is determined by the roots of a polynomial equation which is of degree one higher than the number of components in the compensating network. A pair of complex roots gives an oscillatory term in the response, while a real root gives an aperiodic term. Critical damping requires that all the roots be real and equal, and the imposition of this condition is sufficient to determine all the components of the compensating network.



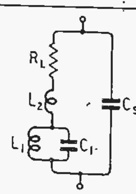
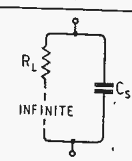
CURVE	NETWORK	COMPONENTS	EQUATIONS & VALUES
1		—	$x = 0.5 \quad 1 \quad 1.5 \quad 2$ $1 - e^{-x}$ 39·4 63·2 77·7 86·5
2		$L_1 = \frac{1}{4} C_S R_L^2$	$1-(1+x)e^{-2x}$ 44·7 72·9 87·6 94·5
3		$L_1 = \frac{8}{27} C_S R_L^2$ $C_1 = \frac{1}{8} C_S$	$1-(1+2x+2x^2)e^{-3x}$ 44·2 75·1 90·6 96·8
4		$L_1 = \frac{1}{4} C_S R_L^2$ $C_1 = \frac{1}{4} C_S$ $L_2 = \frac{1}{16} C_S R_L^2$	$1-(1+3x+4x^2+\frac{16}{3}x^3)e^{-4x}$ 43·6 75·6 91·9 97·8
5		—	$x - \frac{1}{4}x^2$ 43·8 75·0 93·8 100·0

Fig. 8. *Indicial response of the coupling networks shown for the condition of critical damping.*

3.2.—*Particular cases*

Fig. 8 gives response curves and component values for networks consisting of one, two, and three components. The derivation of the two-component case, which is typical, is given in Appendix 3.1.

3.3.—*Maximum possible improvement*

In order to determine whether any worth-while improvement can be got by increasing the number of components, it is useful to determine the response for a network with an infinite number of components. This has been done in Appendix 3.2 and the curve is included in Fig. 8. It will be seen that up to about 75 per cent of the final value the various curves do not differ greatly from one another. The differences are most pronounced for values of t round about $2C_SR_L$ where the values obtained are shown in the table below.

The form taken by the compensating network for the infinite-order case is rather interesting. The

Number of components in compensating network	0	1	2	3	∞
Percentage of final value	86.5	94.5	96.8	97.8	100

network is equivalent to a negative capacitance of value C_s in series with an open-ended transmission line, impedance R_L, delay C_sR_L. The line is thus matched at the input so that there is only one reflection. The combination of input and reflected waveforms give ideal response after a time $2C_sR_L$.

Fig. 9 shows one form of the infinite order compensating network.

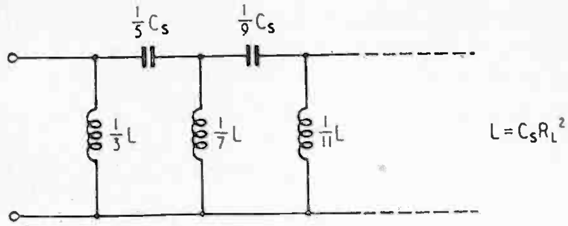


Fig. 9. *One form of the infinite-order compensating network.*

3.4.—*Attenuation and delay-frequency response.*

Figs. 10 and 11 give respectively the attenuation - frequency and delay-frequency characteristics for the five cases dealt with in Fig. 8. The formulae are given in Appendix 3.3.

Acknowledgments

The author is indebted to the Engineer-in-Chief, G.P.O., for permission to publish this article, and to those of his colleagues who have helped in its preparation.

APPENDICES

1. Operational Formulae [1]

General Formulae.

If $\phi(p)\, 1 = f(t)$

then $\phi(p/a)\, 1 = f(at)$ for a real > 0;

$e^{-ap}\phi(p)\, 1 = f(t-a)$ for $t > a$; $a > 0$.

Standard Forms.

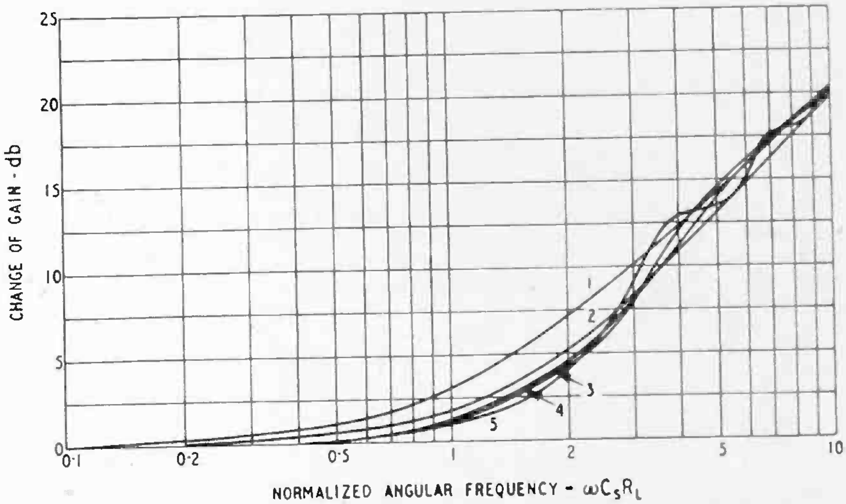
$$\frac{p}{(p-a)^n}\, 1 = \frac{t^{n-1}e^{at}}{(n-1)!} \qquad n > 0$$

$$p^{-n}\, 1 = t^n/n! \qquad n > -1$$

2. Compensation of grid coupling by anode decoupling

2.1. *General.*

The network of Fig. 3 (a) will be treated in detail;



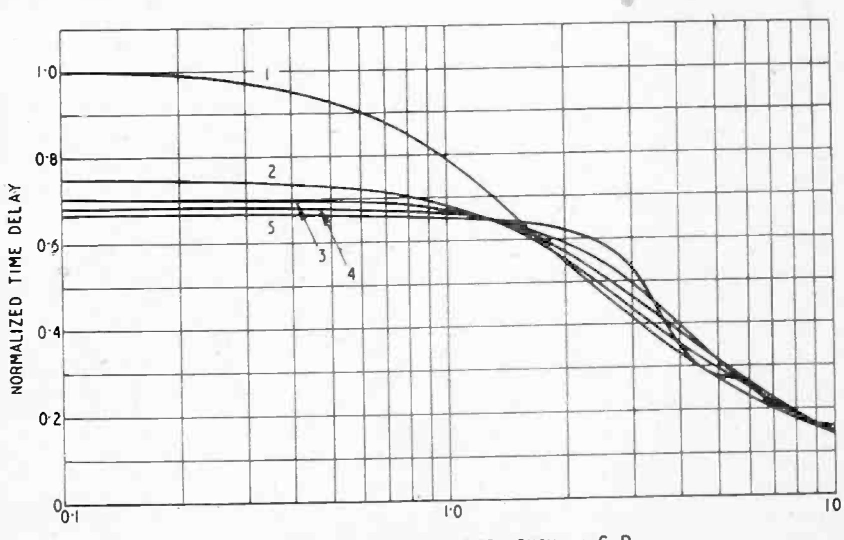


Fig. 10 (*above*). *Amplitude-frequency response curves of the networks of Fig. 8.*

Fig. 11 (*left*). *Delay-frequency characteristics of the circuits of Fig. 8. In each case—Curve 1, uncompensated; curve 2, first-order compensation; curve 3, second-order compensation; curve 4, third-order compensation; curve 5, infinite-order compensation.*

that of Fig. 3 (b) is essentially similar.

We are concerned with two impedances $1 + Z_L$ and $1 + Z_g$ each consisting of a unit resistance in series with Z_L or Z_g respectively. These may be expressed as the ratio of two polynomials in p^2.

$$1 + Z_L = \frac{a_0 + a_1 p + \ldots\ldots + a_{\frac{1}{2}n}\, p^{\frac{1}{2}n}}{1 + b_1 p + \ldots\ldots + b_{\frac{1}{2}n}\, p^{\frac{1}{2}n}}$$

$$1 + Z_g = \frac{1 + c_1 p + \ldots\ldots + c_{\frac{1}{2}n+1}\, p^{\frac{1}{2}n+1}}{d_1 p + \ldots\ldots + d_{\frac{1}{2}n+1}\, p^{\frac{1}{2}n+1}}$$

$$\therefore \quad \frac{1 + Z_L}{1 + Z_g} = \frac{a_0 d_1 p + \ldots\ldots\ldots + a_{\frac{1}{2}n}\, d_{\frac{1}{2}n+1}\, p^{n+1}}{1 + (b_1 + c_1)\,p + \ldots + b_{\frac{1}{2}n}\, c_{\frac{1}{2}n+1}\, p^{n+1}}$$

Apart from the unity term in the denominator, numerator and denominator are equal. This can be shown by general continued fraction theory.[3] $1 + Z_L$ and $1 + Z_g$ can be considered as the $(n + 1)$th and $(n + 2)$th convergents of the same continued fraction :

$$1 + \frac{1}{C_2 p +}\,\frac{1}{R_3 +}\,\frac{1}{C_4 p +}\ \cdots\ \frac{1}{R_{n+1} +}\,\frac{1}{C_{n+2} p +}\ \cdots$$

$$\therefore 1 + Z_L = p_{n+1}/q_{n+1}$$

where p_r and q_r are as defined in reference (3) and are, in this case, polynomials in p ; similarly

$$1 + Z_g = p_{n+2}/q_{n+2}$$

$$\therefore \quad \frac{1 + Z_L}{1 + Z_g} = \frac{p_{n+1}\, q_{n+2}}{p_{n+2}\, q_{n+1}}$$

Now, in general,

$$p_n\, q_{n-1} - p_{n-1}\, q_n = (-1)^n$$

$\therefore \quad p_{n+2}\, q_{n+1} - p_{n+1}\, q_{n+2} = (-1)^{n+2} = 1$, since, in this case, n is even.

$$\therefore \quad \frac{1 + Z_L}{1 + Z_g} = \frac{p_{n+1}\, q_{n+2}}{1 + p_{n+1}\, q_{n+2}}$$

Comparing this with the previous expression we see that $p_{n+1}\, q_{n+2}$ must be identified with the numerator and that the denominator must be equal the numerator plus one. Now convert to a ratio of polynomials in $1/p$:

$$\frac{1 + Z_L}{1 + Z_g} = \frac{F}{F + 1/p^{n+1}} \quad \text{where } F \text{ is a polynomial in } 1/p$$

$$= \frac{1}{1 + 1/Fp^{n+1}}$$

$$= 1 - 1/Fp^{n+1} + 1/(Fp^{n+1})^2 - \ldots$$

The lowest term in $1/p$ in this expression is of degree $n + 1$ therefore, for the corresponding indicial response :

$$\left[\frac{dV}{dt}\right]_0 = \left[\frac{d^2 V}{dt^2}\right]_0 = \ldots\ldots = \left[\frac{d^n V}{dt^n}\right]_0 = 0$$

2.2. *Special case* ; $n = 2$

normalized operational gain

$$= \frac{1 + Z_L}{1 + Z_g} = \frac{pT(1 + \alpha + \alpha pT)(1 + \beta + \alpha\beta pT)}{(1 + \alpha pT)[1 + (1 + \beta + \alpha\beta)pT + \alpha\beta p^2 T^2]}$$

after simplification where α, β, and T are as defined in Fig. 4.

The denominator may be factorized thus :

$$\alpha\beta(p + 1/\alpha T)(p + m/T)(p + q/T)$$

where $m = c + d \qquad c = \frac{1}{2}(1 + 1/\alpha + 1/\alpha\beta)$

$q = c - d \qquad d^2 = c^2 - 1/\alpha\beta$

the gain may therefore be split into three partial fractions each of the standard form $kp/(p + a)$ and hence interpreted to give the indicial response as quoted in section 2.2.

3. High-Frequency Compensation—Critically-Damped Response

3.1. *Second order.*

Normalized operational impedance of load

$$= \frac{\left(R_L + \dfrac{pL_1}{1 + p^2 L_1 C_1}\right)\cdot\dfrac{1}{pC_s}}{R_L\left(R_L + \dfrac{1}{pC_s} + \dfrac{pL_1}{1 + p^2 L_1 C_1}\right)}$$

$$= \frac{1 + \alpha x + \alpha\beta x^2}{1 + x + \alpha(1 + \beta)\,x^2 + \alpha\beta x^3}$$

where $pC_s R_L = x$, $L_1 = \alpha C_s R_L^2$, $C_1 = \beta C_s$

For critical damping, denominator $= (1 + bx)^3$

hence, comparing coefficients, we have $b = \frac{1}{3}$, $\alpha = \frac{8}{27}$, $\beta = \frac{1}{8}$

$$\text{So impedance} = \frac{1 + \frac{8}{27}x + \frac{1}{27}x^2}{(1 + x/3)^3}$$

$$= \frac{x^2 + 8x + 27}{(x + 3)^3} = 1 - \frac{x(x^2 + 8x + 19)}{(x + 3)^3}$$

$$= 1 - \frac{x}{x + 3} - \frac{2x}{(x + 3)^2} - \frac{4x}{(x + 3)^3}$$

Each of these terms may be interpreted by standard forms to give the indicial response

$$V = 1 - (1 + 2t/T + 2t^2/T^2)\, e^{-3t/T}$$

where $T = C_s R_L$

3.2. *Infinite order.*

For n components the operational impedance Z of the compensating network will be of the form

$$Z = \frac{a_1 x + a_3 x^3 + \ldots\ldots\ldots + a_{n-1}\, x^{n-1}}{1 + a_2 x + \ldots\ldots\ldots\ldots + a_n x^n}$$

where $x = pC_s R_L$ and the a's are coefficients specifying Z ; n has arbitrarily been taken as even. For this value of Z the normalized impedance of the load becomes

$$\frac{1 + a_1 x + a_2 x^3 + \ldots\ldots\ldots\ldots + a_n x^n}{1 + x + (a_1 + a_2)x^2 + a_2 x^3 + (a_3 + a_4)x^4 + \ldots\ldots\ldots\ldots + (a_{n-1} + a_n)x^n + a_n x^{n+1}}$$

$$\text{For critical damping, denominator} = \left(1 + \frac{x}{n + 1}\right)^{n+1}$$

As $n \to \infty$ the impedance becomes the ratio of two infinite series, and for critical damping the denominator equals e^x. This specifies the a's as follows :

$a_1 + a_2 = 1/2\,!$ whence $a_1 = 1/2\,! - 1/3\,!$

$a_2 = 1/3\,!$

$a_3 + a_4 = 1/4\,!$ $a_3 = 1/4\,! - 1/5\,!$

etc.

The numerator therefore is :

$$1 + x/2\,! - x/3\,! + x^2/3\,! + x^3/4\,! - x^3/5\,! + \ldots$$

This series can be shown to be absolutely convergent. It may therefore be rearranged :

$$1 + (x^2/2\,! + x^3/3\,! + \ldots\ldots)/x - (x^3/3\,! + x^5/5\,! + \ldots\ldots)/x^2$$

$$= 1 + (e^x - 1 - x)/x - (\tfrac{1}{2}e^x - \tfrac{1}{2}e^{-x} - x)/x^2$$

$$= e^x/x - e^x/2x^2 + e^{-x}/2x^2$$

So the impedance is

$$1/x - 1/2x^2 + e^{-2x}/2x^2$$

Each of these terms may be interpreted by standard forms. The factor e^{-2x} in the third term delays its contribution to the indicial response, which must be considered in two time intervals :

(1) $0 \leqslant t/T \leqslant 2 \qquad T = C_s R_L$

$$V = t/T - t^2/4T^2$$

(2) $2 \leqslant t/T \leqslant \infty$

$$V = t/T - t^2/4T^2 + (t/T - 2)^2/4 = 1$$

The curve is shown in Fig. 8

Z may be obtained from the equation

$$\text{impedance} = \frac{1 + Z}{1 + x(1 + Z)}$$

We find that $Z = \coth x - 1/x$;

$\coth x$ is the impedance of an open-ended transmission line and $-1/x$ that of a negative capacitance. The network of Fig. 9 is readily obtained from the continued fraction expansion of $\coth x$

$$\coth x = 1/x + \frac{1}{3/x+}\,\frac{1}{5/x+}\cdots$$

3.3. *Attenuation and Delay Characteristics.*

Order of Compensation	Attenuation Characteristics	Delay Characteristics
0	$10 \log_{10}(1 + x^2)$	$(\tan^{-1}x)/x$
1	$20 \log_{10}(4 + x^2) - 10 \log_{10}(16 + x^2)$	$\{2 \tan^{-1} \tfrac{1}{2}x - \tan^{-1} \tfrac{1}{4}x\}/x$
2	$30 \log_{10}(9 + x^2) - 10 \log_{10}\{(27 - x^2)^2 + 64x^2\}$	$\{3 \tan^{-1} \tfrac{1}{3}x - \tan^{-1} \dfrac{8x}{27 - x^2}\}/x$
3	$40 \log_{10}(16 + x^2) - 10 \log_{10}\{(256 - 16x^2)^2 + (80x - x^3)^2\}$	$\{4 \tan^{-1} \tfrac{1}{4}x - \tan^{-1} \dfrac{80x - x^3}{256 - 16x^2}\}/x$
∞	$20 \log_{10} 2x^2 - 10 \log_{10}\{(1 - \cos 2x)^2 + (2x - \sin 2x)^2\}$	$(\tan^{-1} \dfrac{2x - \sin 2x}{1 - \cos 2x})/x$

$$x = \omega C_s R_L$$

The delay formulae are for the normalized delay; i.e., ratio of delay to time constant $C_s R_L$.

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CORRESPONDENCE

Transient Response of Filters

To the Editor, "Wireless Engineer"

SIR,—In the March issue D. G. Tucker[1] attempts to find mathematically the transient response of a "Six-element symmetrical" filter section. The ordinary methods of solution prove to give only a guide as to the nature of the solution, and an indirect method of finding the constants involved in the final solution has to be resorted to. Further a number of approximations have to be made.

C. C. Eaglesfield[2] in an article in the November issue attempts and obtains a solution which relies on the bandwidth being small compared with the mid-band frequency.

Acting on a remark made by Tucker in a footnote (page 87) an attempt was made to line-up the transient response of a band-pass network with that of its low-pass analogue.

A low-pass network, which includes components $R_1, R_2, \ldots, L_1, L_2 \ldots, C_1, C_2 \ldots$, has a transfer impedance $Z(\omega)_{L.P}$ which may be written

$$Z(\omega)_{L.P} = f\left\{R_1,\ R_2,\ \ldots,\ j\omega L_1,\ j\omega L_2,\ \ldots,\ \frac{1}{j\omega C_1},\ \frac{1}{j\omega C_2},\ \ldots\right\}$$

Now if in series with every inductance L we place a capacitance $\dfrac{1}{\omega_0^2 L}$, and in parallel with every capacitance C we place an inductance $1/\omega_0^2 C$ we obtain the band-pass analogue. This is just a statement of Landon's Low-Pass—Band-Pass Analogue Theorem. For

$$j\omega L \text{ becomes } j\omega L + \frac{1}{j\omega\left(\dfrac{1}{\omega_0^2 L}\right)}$$

$$= j\omega L\left(1 - \frac{\omega_0^2}{\omega^2}\right)$$

$$\text{and } \frac{1}{j\omega C} \text{ becomes } 1 \Big/ \left\{ j\omega C + \frac{1}{j\omega\left(\dfrac{1}{\omega_0^2 C}\right)}\right\}$$

$$= 1 \Big/ \left\{ j\omega C\left(1 - \frac{\omega_0^2}{\omega_0^2}\right)\right\}$$

Thus if we replace ω for the low-pass case by $\omega\left(\dfrac{1 - \omega_0^2}{\omega^2}\right)$ we have the band-pass case. However, this just states the analogy for the amplitude and phase frequency characteristics. We may demonstrate the analogy for a transient.

By Fourier's Integral Theorem the frequency characteristic of unit step is

$$g(\omega) = \frac{1}{2\pi}\int_0^\infty e^{-j\omega t}\,.\,dt = \frac{1}{2\pi j\omega}$$

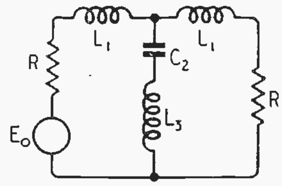
while the frequency response for the wave $\cos \omega_0 t.\ 1$ is

$$g(\omega) = \frac{1}{2\pi}\int_0^\infty \cos \omega_0 t\,.\,e^{-j\omega t}\,.\,dt = \frac{1}{2\pi j}\,\frac{\omega}{\omega^2 - \omega_0^2}$$

Comparing these two cases, we see again that the analogy is obtained by replacing ω by $\omega\left(1 - \dfrac{\omega_0^2}{\omega^2}\right)$

Thus we may state that if the wave $\cos \omega_0 t.\ 1$ is put into a band-pass network then the envelope of the output wave is exactly the same as the output wave obtained when unit step is placed into the low-pass analogue. This is a result (with no restrictions on bandwidth) that should prove of utmost use in the solving of transient problems.

For the "Six-element symmetrical" filter the equivalent low-pass network is clearly (using D. G. Tucker's notation)



where $n\omega_0 = \omega_1$

$$L_1 = \frac{mR}{\omega_1}$$

$$C_2 = \frac{2m}{R\omega_1}$$

$$L_3 = \frac{(1 - m^2)R}{2m\omega_1}$$

When E_0 is unit step, the voltage appearing across R is easily found to be given by

$$V_0 = \tfrac{1}{2}\left\{1 - e^{-\frac{\omega_1 t}{2m}} - \frac{m}{\sqrt{1 - \frac{m^2}{4}}} \cdot e^{-\frac{m\omega_1 t}{4}} \cdot \sin\left(\frac{\omega_1}{2} \cdot \sqrt{1 - \frac{m^2}{4}} \cdot t\right)\right\}$$

which corresponds to equation (7) in C. C. Eaglesfield's article.

London, E.14. E. T. EMMS.

FILTERS

[1] D. G. Tucker. "Transient Response of Filters." *Wireless Engineer*, March 1946, p. 84.

[2] C. C. Eaglesfield. "Transient Response of Filters." *Wireless Engineer*, November 1946, p. 306.

Thermistors at High Frequencies

To the Editor, "Wireless Engineer"

SIR,—I have been using directly-heated high-resistance thermistor* elements to assist impedance measurements on cyclotron models. The models have been operated at about 400 Mc/s and some consideration has been given to the performance of the high-resistance thermistor at these frequencies. Three points have been considered; lead reactance, skin effect and distributed capacitance. Simple calculation has shown that the first two of these cause very small changes in thermistor resistance for the frequency range 0–400 Mc/s. The distributed capacitance does, however, exert considerable influence on the performance of thermistors which have a resistance of over a few thousand ohms.

Using the equivalent transmission line method suggested by Howe[1,2] for carbon-rod resistors, an estimate has been made of the high-frequency resistance of a thermistor of known resistance to direct current. For calculation, the thermistor was regarded as a cylinder of diameter 0.075 cm and length 0.025 cm. These values were taken after measurement with a travelling microscope. With these dimensions the calculated capacitance per unit length of the thermistor becomes about 3.0 $\mu\mu$F.

To calculate the effective impedance we now regard the thermistor as equivalent to a shorted transmission line of length l, resistance per unit length r and capacitance per unit length c where

l = semi-length of thermistor bead
r = resistance per unit length of line; i.e., R/l
R = d.c. resistance of thermistor
c = capacitance per unit length of thermistor = 3.0 $\mu\mu$F per cm in our case.

The input impedance of such a line is quoted by Howe as

$$Z = Z_0 \tanh \alpha l$$

where

$$Z_0 = \sqrt{r/j\omega c}$$
$$\alpha = \sqrt{j\omega c r}$$

For αl large we can take

$$Z \approx Z_0 \approx \sqrt{r/j\omega c}$$

This impedance is equivalent to a resistance $\sqrt{2r/\omega c}$ and a capacitance of reactance $\sqrt{2r/\omega c}$ in parallel.

Thus, for sufficiently high values of resistance and frequency, the high-frequency resistance of the thermistor varies as the square root of the resistance to direct current. This condition is satisfied in our case of resistance greater than 30,000 ohms and frequency of about 400 Mc/s.

Calculated values of the high-frequency resistance for different d.c. resistance values are given in column 2 of the following table.

These can be compared with measured values given in column 3.

D.C. resistance (MΩ)	Calculated h.f. resistance (MΩ)	Measured h.f. resistance (MΩ)
8.9	0.39	0.30
0.073	0.035	0.025

The measured high-frequency resistance has been obtained by using the thermistor as a load on a shorted quarter-wave transmission line. The thermistor was actually soldered between the outer conductor and an insulated core of the inner conductor to enable the 50-c/s control voltage to be applied to the thermistor and to enable d.c. resistance measurements to be made. Screening of the thermistor to prevent radiation was obtained by continuing the outer conductor. The line was formed from brass tube and was damped by a small ring of plasticine attached to the inner conductor to bring the Q down to about 500. From the measured Q and the known dimensions of the line the effective resonant impedance at the open end was calculated. With the thermistor mounted across the open end, the Q of the line was again measured for different thermistor resistances. From the known Q and unloaded input impedance, the effective thermistor resistance was determined from the relation

$$\frac{Q_1}{Q_2} = \frac{R_E + R_i}{R_E}$$

Q_1 = Q without thermistor
Q_2 = Q with thermistor
R_E = Effective resistance of thermistor
R_i = Input resistance of unloaded line at resonance

The possibility of comparing calculated and measured capacitances for the thermistor was

[1] G. W. O. Howe, *Wireless Engineer*, Vol. 12, p. 291, 1935.
[2] G. W. O. Howe, *Wireless Engineer*, Vol. 17, p. 471, 1940.

* The thermistor is a small bead composed of a mixture of metallic oxides and enclosed in a small glass capsule with sealed-in leads. The main characteristic is a very large negative temperature coefficient of resistance. In the directly-heated type of thermistor, the resistance is varied by means of the current through the bead. An indirectly-heated type is also made in which the temperature and, hence, resistance are controlled by a separate heating wire. See G. L. Pearson, *Physical Review*, Vol. 57, p. 1065, 1940.

rejected as the calculated value was of the order 0.001 $\mu\mu$F, which was certainly far too low to measure with the equipment used. It was merely checked that the capacitance was small enough to produce no measurable change in resonant frequency when the effective thermistor impedance changed by a factor of 10. For the apparatus used this meant that the change in capacitance was less than 0.002 $\mu\mu$F.

I am grateful to Standard Telephones and Cables Ltd. for giving the thermistor elements which were specially prepared under the direction of H. Wolfson. Dr. J. H. Fremlin suggested the use of thermistors in cyclotron model measurements and has helped greatly with several stimulating discussions.

Nuffield Physics Laboratory, J. WALKER.
University of Birmingham.

Effect of Aircraft on Fading

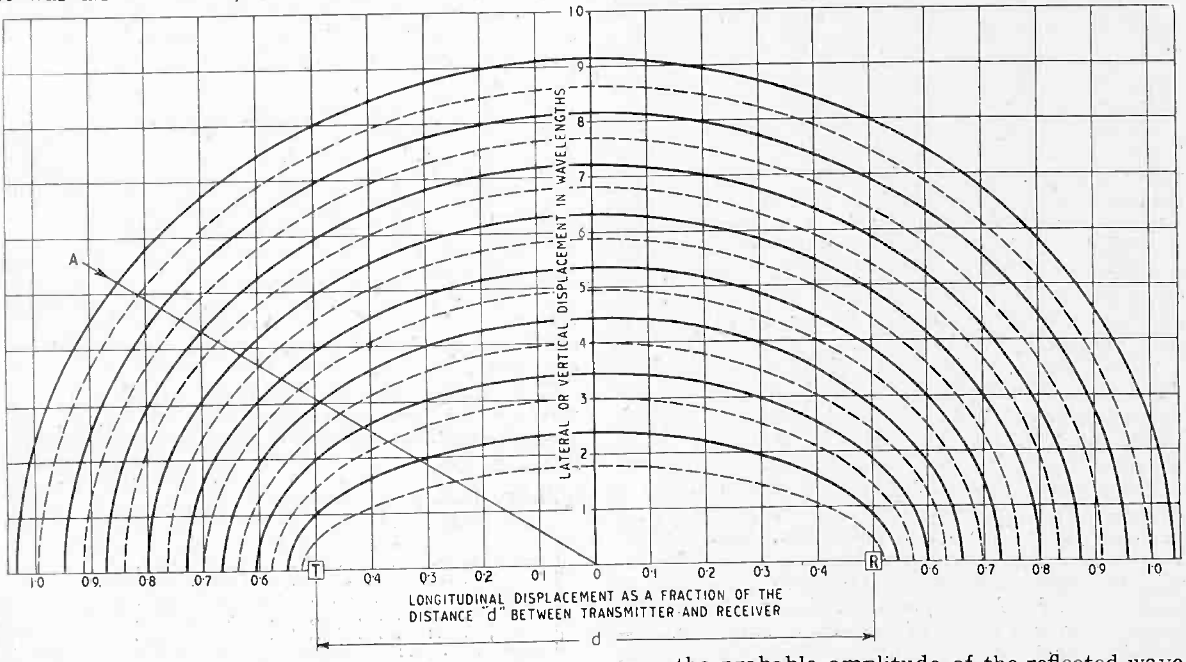
To the Editor, "Wireless Engineer."

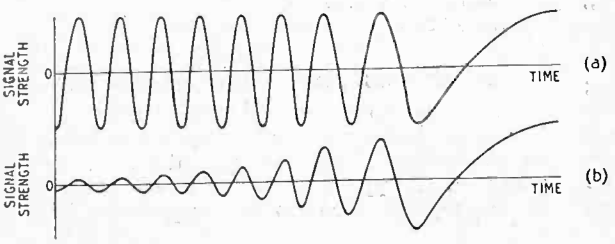
SIR,—While some beam communication trials on 30 to 40 Mc/s and 65 to 100 Mc/s were in progress in 1944, the attention of the writer was brought to frequent occurrences of fading signals in which the fading exhibited the characteristics of steadily changing amplitude and frequency of fade. It was subsequently proved that the phenomenon was due to the presence of aircraft in the transmitted beam, and although the effect was generally most noticeable when field strength measurements were in progress within a mile or two of the transmitter, it was also observed, but to a less extent, when transmitter and receiver were at, or a little beyond, the limits of optical range.

Under the latter conditions of transmitter-receiver separation, there is normally only the direct ray from the transmitter to the receiver ; but when there is an aircraft in the vicinity, there may also be at the receiver an indirect ray reflected from the aircraft. This indirect ray will have travelled over a greater distance than the direct ray, and consequently there will be a phase difference between the two signals at the receiver, the resultant being a minimum or maximum according to whether the phase difference is $(2n + 1)\pi$ or $2n\pi$. Thus, as the aircraft position changes, there is a variation of field strength at the receiver.

The accompanying figure shows the positions of an aircraft relative to the transmitter and receiver which will give signal reinforcement (shown by full lines) at the receiver, and also the positions which will result in a signal minimum at the receiver (shown by dotted lines). The loci of these positions, when plotted in a plane passing through the line joining the transmitter and receiver, are a set of confocal ellipses, with the transmitter and receiver as foci. The ellipses have been plotted with distances parallel to the line joining the transmitter and receiver measured in terms of "d," the distance between transmitter and receiver. Lateral displacements from this line are measured in terms of wavelengths.

The resulting variation in field strength due to an aircraft is shown for the particular case of an aircraft flying along the path AO. Curve (a) shows the rate of fading without taking into account





the probable amplitude of the reflected wave. Curve (b) shows the type of fading which might be expected due to the reduction of the reflected signal strength as the aircraft travels further from the transmitter. The latter curve agrees closely with observations which have been made in connection with the alignment of beam transmitters.

Potters Bar, Middlesex. J. W. WHITEHEAD.

(*Correspondence continued overleaf.*)

Spontaneous Fluctuations in a Double-Cathode Valve

To the Editor, "Wireless Engineer"

SIR,—W. Schottky[1] has indicated that the fluctuations of electricity arising in a "valve" with two similar emitting filaments at the same temperature in a state of thermal equilibrium must obey at once both the formula for the "shot effect" as derived by Schottky[2] himself and for thermal noise as derived by H. Nyquist[3]. D. O. North[4] has also discussed this problem; he has shown in detail that, since in such a valve the conductance g (i.e., $\frac{\delta I}{\delta V}$ where I is the mean resultant current throughout the "valve") must obey the formula :—

$$g = \frac{eJ}{kT} \qquad \text{(1)}$$

where

e = the magnitude of the electronic charge,
k = Boltzmann's constant (1.37×10^{-23} joules/°K.)
T = the temperature in degrees Kelvin,
J = the mean current traversing the valve, in either direction, assumed identical.

It follows that the current fluctuations should satisfy the relationship :

$$\overline{\triangle I^2} = 4gkT\triangle f \qquad \text{(2a)}$$
$$= 4eJ\triangle f \qquad \text{(2b)}$$

where

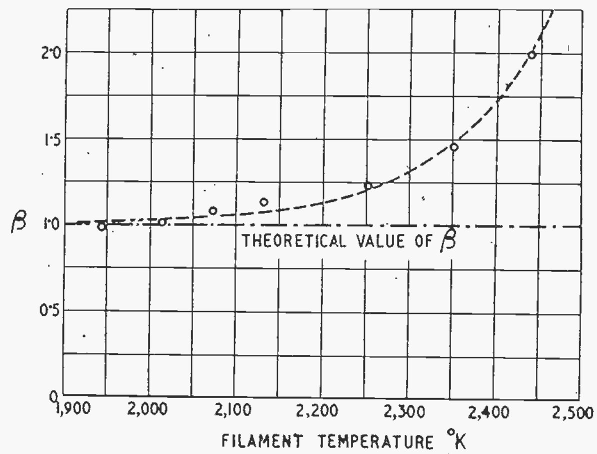
$\overline{\triangle I^2}$ represents the mean square fluctuation of I from its average zero value.
$\triangle f$ is the integrated bandwidth of the measuring instrument.

The latter expression may then be interpreted as the sum of the combined "shot effects" of the currents J.

This result depends on a demonstration that the "space-charge reduction factor" Γ^2, in such a valve is unity. North points out that such should be the case since any variation in the depth of the space-charge barrier arising from a fluctuation in the emission from either cathode should affect the mean current in either direction equally; no compensation of the primary fluctuations, as normally expressed by the factor Γ^2 in normal valves, should therefore arise.

As a result, originally, of discussions with R. Fürth, Edinburgh, on the fundamental connection between "shot" and "thermal" fluctuations, the writer considered that it would be of interest to attempt an experimental verification of this classical theorem. Valves with two tungsten filaments in close proximity—for which the writer is indebted to The M.O. Valve Co., Ltd., Osram Works, London—were employed. An amplifier with a mean response at about 1.6 Mc/s with a bandwidth of a few kc/s was utilized. The results obtained over a range of temperature about 2,000–2,500°K are shown in the figure, where the ordinate β is the ratio of measured "fluctuation temperature" to true temperature.

It appears probable therefrom that the theorem would be obeyed for sufficiently low temperatures; the experimental difficulties, however, associated with diminishing temperatures increase most seriously; the valve dynamic resistance increases rapidly—not only in an absolute sense—but also in comparison with the residual dynamic impedance of the amplifier input itself; further, from equation (2) the magnitude of the generated fluctuations diminishes, and finally it becomes increasingly difficult to guarantee maintenance of equilibrium in the valve. It is hoped later to confirm the behaviour of β at even lower temperatures.



Meanwhile, however the rapid increase of β at higher temperatures which has been carefully confirmed appears of interest. It has often been speculated (e.g., North[5], Moullin[6]) that "excess noise" observed with increasing cathode temperature in normal valves arises from the emission from the cathode of a small amount of positive ions. These, while directly contributing negligible noise, are presumed to influence the space-charge severely during their "leisurely transit." Such an explanation appears to the writer difficult to apply in this case since, according to the discussion mentioned in the second paragraph of this note, the effect of the ions on the space-charge should "balance out." Further, one would presumably expect such ions in any case to be emitted with a Maxwell-Boltzmann distribution at the temperature of the cathode(s) and thus to be incapable of altering the basic result (equation 2a) derived on general statistical grounds.

It is hoped later to carry out a similar investigation on a "valve" with oxide-coated cathodes; considerably lower temperatures can thus be examined.

The writer would like here to acknowledge warmly the invaluable assistance of Mr. E. W. Houghton of the Military College of Science in this investigation.

D. K. C. MACDONALD.
Military College of Science,
Shrivenham.*

1 Schottky, W.: *Z. Phys.*, Vol. 104, p. 248, 1936.
2 Schottky, W.: *Ann. d. Phys.*, Vol. 57, p. 541, 1918.
3 Nyquist, H.: *Phys. Rev.*, Vol. 32, p. 110, 1928.
4 North, D. O.: *R.C.A. Rev.*, Vol. 4, p. 466, 1940.
5 North, D. O.: *R.C.A. Rev.*, Vol. 5, p. 115, 1940.
6 Moullin, E. B.: "Spontaneous Fluctuations of Voltage," Clarendon Press. Cap. 6.

* Now of Clarendon Laboratory, Oxford.

Abstracts and References Index

The Index to the Abstracts and References published in 1946 is in course of preparation and will be available in February, priced 2s. 8d. (including postage). As supplies will be limited our Publishers ask us to stress the need for early application for copies.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

577 969.—Disposition of the carbon-content in a directional microphone of the velocity or pressure type.

H. Dewhurst. Application date 15th February, 1935. *(Secret patent: published 7th May,* 1946.)

AERIALS AND AERIAL SYSTEMS

577 750.—Flexible mounting for the base of a folded dipole aerial for use on the roof of a road vehicle.

The General Electric Co. Ltd., G. C. L. Campbell, W. Grimbaldeston. Application date 1st March, 1943.

578 018.—Dipole aerial, with coaxial reflectors mounted above and below, for broadcasting centimetre waves outwards and downwards over a restricted service-area.

D. I. Lawson, D. Weighton and Pye Ltd. Application date 8th April, 1943.

578 457.—Dipole aerial, comprising conical and disk-shaped elements, suitable for use on an aeroplane.

Standard Telephones and Cables, Ltd. (assignees of A. G. Kandoian). Convention date (U.S.A.) 15th May, 1943.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

577 453.—D. F. receiver in which the signals from two or more separate aerials are combined to generate an harmonic frequency which varies in amplitude or phase in accordance with the direction of the incoming signals.

J. Robinson. Application dates 19th April, 1939, *and 28th March,* 1940.

577 455.—Frequency-changing device for use in a direction finder receiving signals from two separate aerials, combined with means for selecting a given harmonic frequency (divided from 577 453).

J. Robinson. Application dates 19th April, 1939, *and 28th March,* 1940.

577 458.—Directional signalling system in which one or more gas-discharge tubes are utilized to modulate a narrow-central zone in an existing beam of radiation.

H. Hughes and Son Ltd., A. H. W. Beck, and A. J. Hughes. Application date 10th January, 1941.

577 460.—D. F. system wherein a directional aerial is constantly rotated in synchronism with an electromagnet which is controlled by the incoming signals to actuate a pointer.

H. E. Sjöstrand. Convention date (Sweden) 26th March, 1942.

577 527.—Blind-landing system in which one of two overlapping beams is more sharply directive in a vertical plane than the other (addition to 577 276).

Standard Telephones and Cables Ltd., C. W. Earp, and G. G. Samson. Application date 19th September, 1941.

577 672.—Spaced array of aerials, coupled to a common generator through phase-delay networks, for scanning a predetermined area in a radiolocation system.

Hazeltine Corporation (assignees of A. V. Loughren) Convention date (U.S.A.) 26th May, 1941.

577 674.—Radiolocation system in which the beam from a fixed aerial array is swung by electric control so as to scan systematically the area to be explored (divided out of 577 672).

Hazeltine Corporation (assignees of A. V. Loughren), Convention date (U.S.A.) 26th May, 1941.

577 742.—Arrangement for maintaining correct phase and amplitude relations between the signals received on the directional and omni-directional aerials in a sense-determining d.f. receiver.

Marconi's W.T. Co. Ltd., S. B. Smith, and J. F. Hatch. Application date 11th October, 1940.

577 824.—Navigational system in which two spaced transmitters radiate characteristic pulses, so that a mobile craft can follow a predetermined course by maintaining a given time-interval between their reception.

Standard Telephones and Cables, Ltd. (communicated by International Standard Electric Corporation). Application date 14th May, 1942.

577 853.—Blind-landing aerial in which the phase-excitation is arranged to offset distortion due to earth-reflections.

Standard Telephones and Cables, Ltd., and E. H. Ullrich. Application date 25th September, 1940.

577 854.—Radiolocation system in which each exploring pulse partly overlaps the resulting echo, the residual echoes being utilized to give a distance-indication in a graduated milliammeter.

Standard Telephones and Cables, Ltd. (communicated by International Standard Electrical Corporation). Application date 26th December, 1941.

577 855.—Radiolocation system in which auxiliary out-of-phase pulses are utilized to offset undesired reflections from the ground or from fixed objects.

Standard Telephones and Cables, Ltd., and H. G. Busignies. Convention date (U.S.A.) 5th March, 1941.

577 937.—Radiolocation system in which the distance is indicated by the charge developed in a thyratron-controlled capacitor during the interval between the outgoing pulse and the incoming echo.

J. Forman and Pye Ltd. Application date 17th April, 1941.

577 939.—Aerial and reflector system, located at or below ground level, for radiating a navigational beam of rotational symmetry.

H. M. Dowsett. Application date 6th June, 1941.

RECEIVING CIRCUITS AND APPARATUS

(See also under Television)

577 641.—Broadcast receiver in which the press-button tuning control also brings a rotary frame aerial into orientation with the selected transmitter.
E. K. Cole Ltd., and A. W. Martin. Application dates, 16th March, and 3rd November, 1944.

577 775.—Discriminator circuit, for stabilizing frequency-modulated signals on centimetre waves, comprising two velocity-modulating valves of the resonator type (divided out of 577 758).
E. L. C. White. Application date 25th March, 1944.

577 795.—Remote-control system for use say in radiolocation wherein pulsed signals are transmitted over a radio link to synchronize distant apparatus with the rotation of a local directional aerial.
W. Jones and Pye Ltd. Application date 22nd January, 1943.

577 796.—Remote-control system for aligning a local with a distant rotating indicator, applicable to radiolocation equipment.
L. W. Germany and Pye Ltd. Application date 19th February, 1943.

577 800.—Frequency-discriminator for developing a fixed beat-frequency from a variably-tuned circuit and any one of a series of standard oscillators.
The General Electric Co. Ltd., N. R. Bligh, and J. B. L. Foot. Application date 5th April, 1943.

577 801.—Cathode-follower valve circuit, arranged to give an output which is proportional in magnitude and sense to the difference between two input voltages.
D. G. O. Morris and Metropolitan-Vickers Electrical Co. Ltd. Application date 16th April, 1943.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

577 216.—Preventing the d. c. component of a saw-tooth waveform from entering the deflecting coils of a c.r. tube, used say, for television.
Marconi's W.T. Co. Ltd. (assignees of T. T. Eaton). Convention date (U.S.A.) 25th September, 1941.

577 670.—Television system in which, to ensure secrecy, certain of the synchronizing signals are provided by separately-phased local generators at the transmitter and receiver.
Scophony Ltd., and G. Wikkenhauser. Application date, 22nd April, 1938.

577 758.—Stabilizing device for frequency-modulated signals, such as television, where there is no definite or constant mean-frequency.
E. L. C. White. Application date 25th March, 1944.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

577 572.—Combined transmitter-and-receiver installation for carrier-wave signalling over a power-line through a common tuned input-output coupling and a frequency-changing circuit.
Standard Telephones and Cables Ltd. (communicated by International Standard Electric Corporation). Application date 6th November, 1943.

577 710.—Impulse-generator of the multivibrator type comprising means for varying the repetition-frequency, and also the duration of the impulses.
E. A. Newman. Application date 8th March, 1944.

577 713.—Pulse-generator of the blocking-oscillator type in which high-frequency stability is combined with a short build-up time.
Standard Telephones and Cables Ltd. (assignees of A. Alford). Convention date (U.S.A.) 12th January, 1943.

CONSTRUCTION OF ELECTRONIC DISCHARGE DEVICES

577 599.—The use of barium thorate as the activating material in a thermionic cathode surrounded by a perforated metal container.
The General Electric Co. Ltd., S. H. Noble, H. P. Rooksby, W. G. S. Branson, and J. S. Herriott. Application date, 19th January, 1944.

577 602.—Thermionic filament consisting of a plurality of bifilar windings in which the heating current flows in opposite directions through adjacent turns.
Standard Telephones and Cables Ltd (assignees of C. V. Litton). Convention date (U.S.A.) 30th January, 1943.

SUBSIDIARY APPARATUS AND MATERIALS

577 585.—Method of adding a thalliam content to the barrier on boundary layer of a rectifier of the selenium type.
Westinghouse Brake and Signal Co. Ltd., L. E. Thompson, and A. Jenkins. Application date 2nd March, 1944.

577 608.—Band-pass filter in which the mid-band frequency can be varied over a wide range, whilst preserving a constant impedance-ratio.
F. G. Clifford. Application date, 11th February, 1944.

577 613.—Capacitance arrangement for varying the natural frequency of a coaxial resonator, as used for ultra-short waves.
"Patelhold" Patentverwertungs & Elektro-Holding A.G. Convention date (Switzerland), 18th February, 1943.

577 616.—Spraying or sputtering process for preparing the counter-electrode of a rectifier of the selenium type.
Westinghouse Brake and Signal Co., Ltd., L. E. Thompson, and A. Jenkins. Application date, 2nd March, 1944.

577 662. Compensating for temperature-variations in cavity-resonators as used for signalling through waveguides.
Automatic Telephone and Electric Co. and T. H. Turney. Application date, 3rd May, 1944.

577 663.—Circuit which is triggered by a firing pulse of one microsecond to generate a sweep voltage, or an abrupt change of potential, for application to a c. r. indicator.
A. C. Cossor Ltd., B. C. Fleming-Williams, and A. Allen. Application date 4th May, 1944.

577 859.—Piezo-electric crystal with roller electrodes which make contact only at points, or along lines, in the nodal plane of oscillation.
The General Electric Co. Ltd. and E. A. Fielding. Application date, 9th September, 1942.

ABSTRACTS AND REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to the World List practice.

ACOUSTICS AND AUDIO FREQUENCIES

4.2–14 1

Transmission, Reflection, and Guiding of an ponential Pulse by a Steel Plate in Water : Part — Experiment.—M. F. M. Osborne & S. D. Hart. *. acoust. Soc. Amer.*, July 1946, Vol. 18, No. 1, . 170–184.) Part 1 (Theory) was abstracted in 60 of 1945, with brief reference to the present perimental work.

ı.22–13 2

The Velocity of Sound : a Molecular Property.— G. Richardson. (*Nature, Lond.*, 31st Aug. 1946, l. 158, No. 4009, pp. 296–298.)

ı.26 3

The Diffraction of Sound Pulses : Part 1 — Dif- ction by a Semi-Infinite Plane ; Part 2 — Dif- ction by an Infinite Wedge ; Part 3 — Note on Integral occurring in the Theory of Diffraction a Semi-Infinite Screen ; Part 4 — On a Paradox the Theory of Reflexion.—F. G. Friedlander. *oc. roy. Soc. A.*, 24th Sept. 1946, Vol. 186, No. 6, pp. 322–344, 344–351, 352–356 & 356–367.)

.321.9 : 621.396.9 4

Radar in Nature.—T. Roddam. (*Wireless World*, t. 1946, Vol. 52, No. 9, pp. 286–288.) The ure of the mechanism by which a bat produces supersonic pulses was described in 258 of 1946 (Hartridge). Its performance is here considered, with a theoretical discussion of the requirements of a radar model using supersonic waves in air.

534.84 5

Improvement in the Acoustics of Reinforced Concrete Buildings by the Oelsner Method.—(*Tech. wet. Tijdschr.*, Jan./March 1945, Vol. 14, Nos. 1/3, p. 14. In Flemish.) Summary of a paper in *Ingeniør-og-Bygningsvoesen*, 10th Feb. 1943.

621.317.75.029.3 : 532.593 6

A Frequency Analyser used in the Study of Ocean Waves.—Barber, Ursell, Darbyshire & Tucker. (*See* 203.)

621.395.61 : 534.43 7

Tuned-Ribbon Pickup.—W. F. Leidel, Jr., & N. E. Payne. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 67–69 . . 101.) Describes a variable-reluctance type of gramophone pickup in which the armature is supported by a tensioned ribbon. Extremely low mass and a flat response up to 10 000 c/s are claimed.

621.396.615.11 8

A Simple "Wien Bridge" Audio Oscillator.— H. T. Sterling. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 29–32.) Details of the theory of operation show that the oscillator has a good waveform and a uniform output amplitude with respect to frequency. Practical circuits are given for (*a*) an oscillator having a variable range from 15 c/s to 150 kc/s, and (*b*) an oscillator having ten spot frequencies between 20 c/s and 20 kc/s.

621.395.625.6 9

16-mm Sound-on-Film Recorders.—J. Neil. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 309–312.)

AERIALS AND TRANSMISSION LINES

621.315.2 : 621.317.33.029.63 10

The Measurement of Cable Characteristics at Ultra-High Frequencies.—Jones & Sear. (*See* 175.)

621.315.21.029.4/.6] : 621.317.33 11

Characteristics of R.F. Cables.—Stamford & Quarmby. (*See* 174.)

621.315 [.211.2 + .22 12

Mineral-Insulated Metal-Sheathed Conductors.— F. W. Tomlinson & H. M. Wright. (*J. Instn elect. Engrs*, Part II, Aug. 1946, Vol. 93, No. 34, pp. 325–335. Discussion, pp. 336–340.) Powdered magnesium oxide is a very satisfactory insulating material. Its dielectric constant is 3.6 and power factor 0.000 5. Copper-covered cable with this insulator is suitable for transmission at frequencies up to several hundred megacycles per second. Properties, applications, and methods of manufacture and installation of such cables are discussed.

621.315.232.014.1 **13**

[Current] **Rating of Cables in Ducts.**—C. C. Barnes. (*Elect. Times*, 31st Oct. 1946, Vol. 110, No. 2871, pp. 583–586.) Tables and graphs are given for calculating the rating, which is lower owing to heat effects than for other forms of installation.

621.317.336 **14**

Impedance Matching with an Antenna Tuner.—G. G. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 38–40.)

621.392 **15**

On the Eigen-Values of an Electromagnetic Waveguide.—T. Kahan. (*C. R. Acad. Sci., Paris*, 11th Feb. 1946, Vol. 222, No. 7, pp. 380–381.) The behaviour of a straight waveguide of rectangular cross-section is illustrated by means of the special case of a square cross-section. The form of the vector potential and nodal surfaces (which divide the guide into electrically distinct portions) is considered for some of the simpler possible modes.

621.392.2 **16**

On the Propagation of Waves in Curved Guides.—M. Jouguet. (*C. R. Acad. Sci., Paris*, 4th March 1946, Vol. 222, No. 10, pp. 537–538.) The method of perturbation can be used to determine the deformation of E_0 or H_0 waves in cylindrical guides even when the conductivity σ of the walls is finite. The perturbation terms remain small provided that the curvature C does not exceed a value independent of σ for E_0 waves and a much smaller value, decreasing to zero as σ increases indefinitely, for H_0 waves. The E_0 waves are stable; the phase velocity and attenuation factor (dependent on σ) are independent of curvature to a second-order approximation. For a perfectly conducting guide the H_0 wave is unstable; for large C and σ a solution still exists but differs radically from the H_0 wave. For rectangular guides, an analogous method yields an approximate formula easier to use than the exact formula obtained by the Bromwich-Borgnis method. If the walls are perfect conductors, the phase velocity is not modified (to a second-order approximation) by the curvature, and both E_0 and H_0 waves are stable.

621.392.43 **17**

Conditions of Termination in a Waveguide.—T. Kahan. (*C. R. Acad. Sci., Paris*, 4th March 1946, Vol. 222, No. 10, pp. 535–537.) A proof that if a waveguide is terminated by a semi-transparent cavity whose length is an odd number of quarter wavelengths, no standing waves are produced whatever the mode of vibration. For a definition of semi-transparent cavity see 1795 of 1946.

621.392.43.082.7 **18**

Note on a Reflection-Coefficient Meter.—N. I. Korman. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 657–658.) The reflection coefficient of a network N terminated by an impedance Z_L is defined as $r = \dfrac{Z_0 - Z_L}{Z_0 + Z_L}$ where Z_0 is an impedance such that the current through an indicating device V is zero. It is deduced that the indication of V will be proportional to r if (1) N is a linear, bilateral, passive network, (2) Z_0 is a physically realizable impedance and (3) the output impedance of N equals Z_0.

621.392.5 **19**

Equalized Delay Lines.—Kallmann. (*See* 41.)

621.396.67 **20**

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621.396.67 **21**

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621.396.67 **22**

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621.396.67.011.2 **23**

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621.396.674.011.2 25

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621.396.677 : 621.397.5 26

Rhombic Antennas for Television.—J. Minter. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 58–61.) Design data for aerials receiving horizontally polarized signals in the ranges 44–215 Mc/s and 480–920 Mc/s with polar diagrams for various lengths of side. A forward inclination of the rhombic is used to lower the main lobe.

621.396.678 27

A New Kind of Skyhook.—D. T. Ferrier & W. G. Baird, Jr. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 24–25.) A description of a "kytoon" having the advantages of both a kite and a balloon. The casing is made of light nylon fabric and the bladder of neoprene. The kytoon when deflated can be carried in a package 14 inches long and 3 inches in diameter. It is used for erecting aerials at mobile stations in a short time.

621.396.67 (02) 28

Currents in Aerials and High Frequency Networks. [Book Review]—F. B. Pidduck. Oxford Univ. Press, 97 pp., 8s. 6d. (*Phil. Mag.*, Oct. 1945, Vol. 36, No. 261, p. 730.) ". . . no one interested in aerial theory can afford to neglect this hitherto unpublished account." See also 1182 of 1946.

CIRCUITS AND CIRCUIT ELEMENTS

621.3 : 519.241.6 29

An Extension of Campbell's Theorem of Random Fluctuations.—R. S. Rivlin. (*Phil. Mag.*, Oct. 1945, Vol. 36, No. 261, pp. 688–693.) Campbell's theorem giving the mean square deviation of the response $\overline{(y - \bar{y})^2}$ of a linear system to a large number of identical random events is extended to the case in which the events are dissimilar, and the quantity $\overline{(y - \bar{y})^p}$ is also calculated. In an addendum the result obtained by Rice for $p = 3$ in the case of identical events is shown to agree with that of the present paper.

621.3.015.1 : 621.396.012.8 30

Differentiation of a Voltage.—J. Görner. (*Arch. tech. Messen*, July 1940, No. 109, pp. T77–78.) Consideration of RCL, RC, transformer and choke circuits and electromechanical methods.

621.316.313.025 31

A.C. Network Analyzer.—N. H. Meyers & N. R. Schultz. (*Gen. elect. Rev.*, Sept. 1946, Vol. 49, No. 9, pp. 34–40.) A general description of the analyser is given. It may be used for determining physical entities from equivalent circuits in problems of the following types: (*a*) short circuits, (*b*) transient stability, (*c*) steady state stability, (*d*) shunting and out-of-step operation of electrical machinery, (*e*) torsional oscillations, (*f*) electromagnetic cavity resonators and (*g*) Schrödinger equation of atom structure.

Once a particular equivalent circuit has been determined, it can be set up on the analyser; readings of current, voltage and power then correspond to certain physical entities in the physical system for which the equivalent circuit has been obtained.

621.316.86 : 546.281.26 32

Silicon Carbide Non-Ohmic Resistors.—Ashworth, Needham & Sillars. (*See* 141.)

621.316.974 : 621.318.4.017.31 33

Power Loss in Electromagnetic Screens.—C. F. Davidson, R. C. Looser & J. C. Simmonds. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 315–316.) Discussion of the formula derived by C. A. Siocos (2710 of 1946) for the eddy current density in a plane sheet due to an inductor of finite length. It agrees with the formula derived by the writers (see 1077 of 1946) and is more convenient when the inductor is a solenoid.

621.317.336 34

Impedance Matching with an Antenna Tuner.—G. G. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 38–40.)

621.318.323.2.042.15 35

Permeability of Iron-Dust Cores.—G. W. O. H.: H. W. Lamson: R. E. Burgess. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 291–292 & 313–315.) An editorial discussion of various permeability formulae proposed by Doebke and Howe, and the letters from H. W. Lamson and R. E. Burgess which inspired it. Lamson investigates theoretically the permeability for three arrangements of particles: (i) uniform distribution of aligned cubes, (ii) random distribution of aligned cubes, and (iii) uniformly distributed spheres. The results are tabulated and compared with the experimental values of Legg & Given (4424 of 1940) which are much higher at large iron concentrations. Reasons for this discrepancy are discussed.

Burgess comments on Howe's formula (1033 Abstracts, p. 173) at small iron concentrations, suggests reasons for the discrepancy between this theory and measurement, and draws attention to L. Page's work (1176 of 1942) which provides satisfactory agreement for large iron concentrations.

621.385 : 621.3.011.2 36

Negative Resistance Circuit Element.—G. A. Hay. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 299–305.) The possibility of making the dynatron available for dynamic resistance measurement up to 100 Mc/s is considered. After a review of previous work the optimum operating conditions and the equivalent circuit of the dynatron are considered. Measurements of the low-frequency value of the negative resistance (R_a) by a bridge method and of the high-frequency value (R_a) by a tuned circuit method are described. The ratio $\epsilon = R_a/R_a$ is investigated up to 100 Mc/s after removing a spurious effect due to resonances in the voltage supply leads. It is concluded that ϵ lies between 0.95 and 1.05 up to 50 Mc/s. At higher frequencies the effects of dielectric loss, transit time and valve lead inductances limit the usefulness of the dynatron.

621.385.831.012.8 37

Equivalent Noise Representation of Multi-Grid Amplifier Tubes.—R. Q. Twiss & E. J. Schremp.

pp. 325–335. Discussion, pp. 336–340.) Powdered magnesium oxide is a very satisfactory insulating material. Its dielectric constant is 3.6 and power factor 0.000 5. Copper-covered cable with this insulator is suitable for transmission at frequencies up to several hundred megacycles per second. Properties, applications, and methods of manufacture and installation of such cables are discussed.

621.315.232.014.1 **13**
[Current] **Rating of Cables in Ducts.**—C. C. Barnes. (*Elect. Times*, 31st Oct. 1946, Vol. 110, No. 2871, pp. 583–586.) Tables and graphs are given for calculating the rating, which is lower owing to heat effects than for other forms of installation.

621.317.336 **14**
Impedance Matching with an Antenna Tuner.—G. G. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 38–40.)

621.392 **15**
On the Eigen-Values of an Electromagnetic Waveguide.—T. Kahan. (*C. R. Acad. Sci., Paris*, 11th Feb. 1946, Vol. 222, No. 7, pp. 380–381.) The behaviour of a straight waveguide of rectangular cross-section is illustrated by means of the special case of a square cross-section. The form of the vector potential and nodal surfaces (which divide the guide into electrically distinct portions) is considered for some of the simpler possible modes.

621.392.2 **16**
On the Propagation of Waves in Curved Guides.—M. Jouguet. (*C. R. Acad. Sci., Paris*, 4th March 1946, Vol. 222, No. 10, pp. 537–538.) The method of perturbation can be used to determine the deformation of E_0 or H_0 waves in cylindrical guides even when the conductivity σ of the walls is finite. The perturbation terms remain small provided that the curvature C does not exceed a value independent of σ for E_0 waves and a much smaller value, decreasing to zero as σ increases indefinitely, for H_0 waves. The E_0 waves are stable; the phase velocity and attenuation factor (dependent on σ) are independent of curvature to a second-order approximation. For a perfectly conducting guide the H_0 wave is unstable; for large C and σ a solution still exists but differs radically from the H_0 wave. For rectangular guides, an analogous method yields an approximate formula easier to use than the exact formula obtained by the Bromwich-Borgnis method. If the walls are perfect conductors, the phase velocity is not modified (to a second-order approximation) by the curvature, and both E_0 and H_0 waves are stable.

621.392.43 **17**
Conditions of Termination in a Waveguide.—T. Kahan. (*C. R. Acad. Sci., Paris*, 4th March 1946, Vol. 222, No. 10, pp. 535–537.) A proof that if a waveguide is terminated by a semi-transparent cavity whose length is an odd number of quarter wavelengths, no standing waves are produced whatever the mode of vibration. For a definition of semi-transparent cavity see 1795 of 1946.

621.392.43.082.7 **18**
Note on a Reflection-Coefficient Meter.—N. I. Korman. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 657–658.) The reflection coefficient of a network N terminated by an impedance Z_L is defined as $r = \frac{Z_0 - Z_L}{Z_0 + Z_L}$ where Z_0 is an impedance such that the current through an indicating device V is zero. It is deduced that the indication of V will be proportional to r if (1) N is a linear, bilateral, passive network, (2) Z_0 is a physically realizable impedance and (3) the output impedance of N equals Z_0.

621.392.5 **19**
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621.396.67 **20**
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(*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 696.) North's results for noise generated by multi-grid amplifier valves are extended to the case where there are arbitrary impedances in all the electrode leads. Equations for a pentode are given. Summary of Amer. Phys. Soc. paper.

621.392 **38**

Unification of Linear Network Theory.—J. D. Weston. (*J. Brit. Instn Radio Engrs*, Jan./Feb. 1946, Vol. 6, No. 1, pp. 4–14.) "The concern of this paper is not so much to present new concepts as to achieve a unification of old ones in a consistent scheme." The axioms of projective geometry fall into pairs, so that either member of a pair can be converted into the other by interchanging certain words (correlatives). Associated, therefore, with each theorem derived from the axioms is a valid correlated theorem which may be written down from inspection of the first. There is similar duality in the six axioms (restatements of the laws of Kirchhoff, Ohm, Coulomb and Neumann) applicable to linear, invariant networks having lumped circuit parameters and in which the currents and potentials are steady or slowly varying. "In virtue of this [dual] correspondence it is possible to halve the number of axioms required, provided we add to them the Principle of Duality."

The correlative terms are listed (*e.g.* junction and mesh; admittance and impedance; current and fall of potential) and it is shown how they may be used to group the network theorems into pairs. An outline of the application of the above principles to field theory is given, in which $\mathbf{H}$ and $j\mathbf{E}$: $\mathbf{M}$ and $j\mathbf{P}$ are the pairs of correlatives. "It is important to notice . . . that the same general principles apply to any dynamical system, the axioms of which can be formulated in an analogous way."

621.392.001.1 **39**

Node-Pair Method of Circuit Analysis.—W. H. Huggins. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 661–662.) It is claimed that the confusion and duplication of the impedance and admittance treatments of circuit analysis could be avoided by greater use of the node-pair method, particularly for valve systems. "The node-pair quantities are the most natural and consistent with those already used in everyday measurements."

621.392.5 **40**

Note on a Parallel-T Resistance-Capacitance Network.—A. Wolf. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, p. 659.) Formulae are developed for the performance of a four-terminal parallel-T resistance-capacitance network which serves for the elimination of a given frequency. It is shown that an unsymmetrical form of the network is advantageous when a high degree of frequency discrimination is desired.

621.392.5 **41**

Equalized Delay Lines.—H. E. Kallmann. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 646–657.) The decrease in delay time due to decrease in effective inductance of the coil when the wavelength along the line becomes comparable with the length of the line may be compensated by an increase in the capacitance. This capacitance may be artificially increased by copper strips on the outside of the central coiled conductor, insulated from it, from each other and from earth. Automatic compensation due to the self-capacitance of the coil may suffice with high impedance lines.

The technique of measurement of delay and transmission characteristics is discussed and it is concluded that most of the transmission losses at the higher frequencies occur in the insulation of wire forming the coil. The measurement of input impedance and the matching of the line to source and load are also discussed.

Practical delay lines are described including some with sectional windings. There is a short discussion of the delay in low-pass filters with lumped parameters.

621.392.5 **42**

Network Synthesis, especially the Synthesis of Resistanceless Four-Terminal Networks.—B. D. H. Tellegen. (*Philips Res. Rep.*, April 1946, Vol. 1, No. 3, pp. 169–184.) For a two-terminal network, the "order" is defined as the order of the differential equation of the free vibrations when its terminals are connected by a resistance. This concept is extended to four-terminal networks, and possible types of resistanceless network for various orders are discussed.

621.392.5 **43**

Determination of a Class of Coupled Circuits with *n* Degrees of Freedom, having the Same Natural Frequencies as a Given Assemblage of *n* Coupled Circuits, and such that Each Mesh also has the Same Total and Coupling Self-Inductance as the Corresponding Mesh of the Given Assemblage.—M. Parodi. (*C. R. Acad. Sci., Paris*, 11th Feb. 1946, Vol. 222, No. 7, pp. 379–380.) A continuation of 2501 of 1946. The self-inductance condition is satisfied by requiring one of the matrices involved in the earlier theory to be of a special type.

621.392.5 : 621.316.722.078.3 **44**

The Theory of the Non-Linear Bridge Circuit.—G. N. Patchett. (*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, p. 343.) Discussion of 867 of 1946.

621.392.5 : 621.316.722.078.3 : 621.326[.1 + .3/.4 **45**

The Characteristics of Lamps as applied to the Non-Linear Bridge, used as the Indicator in Voltage Stabilizers.—G. N. Patchett. (*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, pp. 305–322.) The effect of ambient temperature and vibration on the usefulness of various types of lamp in d.c. bridges is considered. "Experimental and mathematical results are given for the response time of various lamps when used in this circuit. Methods of overcoming this delay by means of suitably designed capacitance-resistance networks are given, together with experimental results." See also 867 of 1946.

621.392.52 **46**

The Universal Characteristics of Triple-Resonant-Circuit Band-Pass Filters.—K. R. Spangenberg. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 629–634.) "The universal insertion-loss versus frequency characteristics of a band-pass filter composed of one, two, or three lossless resonant circuits in a loosely coupled cascade connection between a source and a load impedance are given. The effects of load and source coupling and of intermesh coupling upon pass-band insertion-loss variations and upon band width are discussed."

621.392.52 **47**

Double-Derived Terminations.—R. O. Rowlands. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 292–295.) Extension of an earlier method (872 of 1946) of terminating complementary filters at their common ends so that the real part of the image impedance is that of a double-derived section. The technique now described permits a wider choice of attenuation function for a prescribed impedance function.

621.392.52.015.3 **48**

Transient Response of Filters.—C. C. Eaglesfield. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 306–307.) Analysis of the response of a 6-element symmetrical filter section to a sudden application of cos $\omega_0 t$, on the assumption that the pass-band-width is small compared with the central frequency. Equation (7) for the envelope of the output voltage has the same form as Tucker's empirical equation (see 1188 of 1946) but there is a discrepancy between the numerical results which is not fully understood.

621.392.52.091 **49**

The Effect of Incidental Dissipation in Filters.—E. A. Guillemin. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 130–135.) A theoretical paper which presents "a method of ascertaining the effect of these losses on propagation factor, reflection factor and interaction factor". Approximate formulae are developed for the various portions of a low-pass filter characteristic. It is shown how to extend the results to more complicated networks.

621.394/.397].645.34 **50**

A Variation on the Gain Formula for Feedback Amplifiers for a Certain Driving-Impedance Configuration.—T. W. Winternitz. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 639–641.) An expression is obtained for the gain when the source impedance is the only one across which the feedback voltage is developed. This expression is then used to obtain the response, to a Heaviside unit step, of differentiating and integrating amplifiers and a sweep amplifier driving a cathode-ray tube with magnetic deflexion.

621.394/.396].645.015.33 **51**

Pulse Transmission in Amplifiers.—A. E. Newlon. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 116–121.) An experimental investigation carried out with long pulses at a carrier frequency of 460 kc/s. The results may be 'scaled' to cover much shorter pulses. The rounding of the leading edge and the increase in length of the pulse are determined for transmission through amplifiers having various steady-state response characteristics; simple expressions are derived empirically for the pulse distortion in terms of such characteristics. The transmission properties of amplifiers employing single- and double-tuned circuits are compared.

621.396.611 : 518.61 **52**

Calculation of the Electromagnetic Field, Frequency and Circuit Parameters of High-Frequency Resonator Cavities.—H. Motz. (*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, pp. 335–343.) The wave equation $\Delta\phi + (\omega^2/c^2)\phi = 0$ is replaced by a system of difference equations which for free vibrations are soluble only for the 'proper' values of ω/c. A method for finding the least value of ω/c without solving a determinantal equation is described. The sharp corners of klystron resonator boundaries present a special problem. The analytic behaviour of the fields near such sharp corners is allowed for, in a manner well suited to the relaxation method of solving the equations; the computation work is thereby reduced. Once the field components and the resonant frequency are found, the beam impedance and the damping constant are easily determined.

621.396.611 : 534.014.2 **53**

An Experimental Investigation of Forced Vibrations in a Mechanical System having a Non-Linear Restoring Force.—C. A. Ludeke. (*J. appl. Phys.*, July 1946, Vol. 17, No. 7, pp. 603–609.) The apparatus is capable of generating and recording forced vibrations; and the experimental waveforms are compared with the theoretical results given by three graphical methods due to Martienssen (*Phys. Z.*, 1910, Vol. 11, pp. 448–460), den Hartog (*J. Franklin Inst.*, Oct. 1933, Vol. 216, No. 4, pp. 459–473) and Ranscher (*J. appl. Mech.*, 1938, Vol. 5, pp. A169–A177). In all cases, the waveforms of the resulting motion are nearly sinusoidal as long as the frequency of the observed motion is the same as the frequency of the disturbing force, but for a certain kind of non-linearity the frequency of the forced vibration can be made a submultiple (*e.g.* 1/3 or 1/5) of that of the driving force.

621.396.611.3 **54**

Universal Optimum-Response Curves for Arbitrarily Coupled Resonators.—P. I. Richards. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 624–629.) Generalized analysis of the frequency response of a source and sink linked by n coupling elements and n resonators, with the restrictions that (i) all the elements are lossless, (ii) the coupling circuits are not resonant near the central frequency f_0, (iii) all the resonators have their resonant frequencies near f_0, and (iv) the desired bandwidth Δf is small compared with f_0.

The overall loss M has the form 10 log $(1 + Z)$ db where Z is a polynomial of degree $2n$ in x and where $x = (f - f_0)/\Delta f$ and f is any frequency. There is only one optimum form for Z giving maximum off-band rejection with $M \leqslant M_0$ (the allowed loss) in the pass band. This is $Z = d[T_n(x)]^2$ where $d = \text{antilog}_{10}(M_0/10) - 1$, and $T_n(x)$ is the Tchebycheff polynomial of degree n.

621.396.615.11 **55**

A Simple "Wien Bridge" Audio Oscillator.—Sterling. (*See* 8.)

621.396.615.17 **56**

Multivibrator Circuits.—N. W. Mather. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 136, 138.) Basic types collected for reference, showing waveforms at different points, methods of injecting synchronizing signals, and frequency-determining equations.

621.396.616.029.64 **57**

Buzzer Signal Generator for 3 000 Mc/s.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, p. 140.) A buzzer produces r.f. pulses of complex waveform which pass through a coaxial cable and a coupling loop to a resonator tunable from 1 000 to 3 500 Mc/s. The output is controlled by a piston attenuator.

621.396.645 58
Preventing Self-Oscillation in Tetrode Amplifiers.—P. D. Frelich. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 22–23 . . 112.) The grid must be driven from a source of low impedance, *e.g.* a cathode follower. Practical details of a typical circuit are given.

621.396.645 59
Transoceanic Radio Amplifier.—C. F. P. Rose. (*Bell Lab. Rec.*, Sept. 1946, Vol. 24, No. 9, pp. 326–329.) A description of the Western Electric D-158974 amplifier, the frequency of which can be rapidly changed between 4.5 and 23 Mc/s and which can operate for 22 hours a day.

The amplifier has one stage employing four 25-kW tubes connected in a push-pull bridge-neutralized circuit with the two tubes on each side of the circuit connected in parallel. The outdoor power plant transforms the three-phase input of 4 160 V to 230 V for the low-voltage equipment and to 13 000 V for the three-phase high-voltage rectifier.

621.396.645.35 60
Gas-Tube Coupling for D.C. Amplifiers.—F. Iannone & H. Baller. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 106–107.) "Combining a cathode follower with a gas diode gives a stable and efficient coupling network." The grid of the cathode follower is connected directly to the anode of the driver stage, while the isolating gas tube is in the cathode circuit of the cathode follower.

621.397.645 61
Video Amplifier H.F. Response : Part 1.—(*Wireless World*, Sept. 1946, Vol. 52, No. 9, pp. 301–302.) The procedure for determining the optimum circuit values for the shunt-corrector circuit is given, together with several examples. See part 2 below.

621.397.645 62
Video Amplifier H.F. Response : Part 2.—(*Wireless World*, Oct. 1946, Vol. 52, No. 10, pp. 333–334.) The procedure is explained for finding circuit values (*a*) for the flattest frequency response, (*b*) for critical damping, given the drop in response required at a known maximum frequency, or the response required at a known time after the onset of a pulse, and the total circuit capacitance. See part 1 above.

621.392 : 517.432.1 (02) 63
Heaviside's Electric Circuit Theory. [Book Review]—H. J. Josephs. Methuen, London, 115 pp., 4s. 6d. (*Wireless World*, Oct. 1946, Vol. 52, No. 10, p. 332.) See also 2535 of 1946.

621.396.67 (02) 64
Currents in Aerials and High Frequency Networks. [Book Review]—Pidduck. (*See* 28.)

GENERAL PHYSICS

530.162 65
On Onsager's Principle of Microscopic Reversibility.—H. B. G. Casimir (*Philips Res. Rep.*, April 1946, Vol. 1, No. 3, pp. 185–196.) Onsager's theory of reciprocal relations in irreversible processes is summarized and the fluctuations in the parameters of an adiabatic system are evaluated in a general form. The theory is applied to a number of simple examples : the thermon olecular pressure difference, the conduction of heat in crystals, and the conduction of electricity of solids, considered in terms of an arbitrary four-pole.

535.13 + 621.396.11 66
On an Interpretation of the Propagation of E.M. Waves and Its Consequences.—Haubert. (*See* 218.)

535.241.4 67
"Foot-Lambert" Unit of Picture Brightness.—(*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 57, 111.) Definition of the term, and its adoption in the television industry.

535.343.4 + 538.569.4.029.64 + 621.396.11.029.64]:551.57 68
The Absorption of 1-cm Electromagnetic Waves by Atmospheric Water Vapor.—Kyhl, Dicke & Beringer. (*See* 219.)

535.343.4 + 621.317.011.5 + 621.396.11.029.64] : 546.171.1 69
The Inversion Spectrum of Ammonia.—W. E. Good. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 213–218.) For a preliminary account see 3236 of 1946.

536 : 621.3.012.8 70
Thermal Inductance.—R. C. L. Bosworth. (*Nature, Lond.*, 31st Aug. 1946, Vol. 158, No. 4009, p. 309.) Earlier theory suggested that there are no inductances in the equivalent circuits of a thermal system because of the absence of oscillatory phenomena. Further theoretical and experimental investigation shows that transients in a fluid system with convection currents can only be explained by postulating an inductance corresponding to the kinetic energy of the currents.

537.122 : [537.312.62 + 538.224 71
Diamagnetism and Superconductivity of a Collective Electron Assembly.—W. Band. (*Proc. Camb. phil. Soc.*, Oct. 1946, Vol. 42, Part 3, pp. 311–327.)

537.122 : 538.3 72
The Classical Equations of Motion of an Electron.—C. J. Eliezer. (*Proc. Camb. phil. Soc.*, Oct. 1946, Vol. 42, Part 3, pp. 278–286.)

537.222.1 73
The Two-Dimensional Electric Field of a Single Semi-Infinite Rectangular Conductor.—N. Davy. (*Phil. Mag.*, Oct. 1945, Vol. 36, No. 261, pp. 694–705.) The equipotentials and lines of force inside and outside a charged semi-infinite rectangular conductor are investigated theoretically and experimentally and their distribution shown in diagrams. The electric intensities, surface densities, total charges, capacitances, and mechanical forces on an external object are discussed, and special cases of a semi-infinite thin plate and an infinitely narrow hollow conductor are considered.

537.226 : 62 74
Theoretical Physics [of dielectrics] **in Industry.**—Fröhlich. (*See* 124.)

537.291 75
Influence of Space Charge on the Bunching of Electron Beams.—L. Brillouin. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 187–196.)

The application of the Llewellyn method of integration to electron motion within plane structures, taking account of space charge effects, is described. The conditions for bunching, *i.e.* intersection of trajectories, are investigated by this method for the cases of (1) a conventional plane diode, (2) a diode with given initial electron velocity (with sinusoidal velocity modulation as a special case), (3) a plane magnetron, (4) a plane magnetron with velocity modulation. In the last case it is shown that multiple intersections occur with strong magnetic fields.

537.5 : 621.385.18.029.64 76
Conductivity of Electrons in a Gas at Microwave Frequencies.—H. Margenau. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 698.) "Using the energy distribution law the complex conductivity is calculated as function of electron density, gas pressure and frequency of the field." The results are applied to t.r. switches. See 3818 of 1946. Summary of Amer. Phys. Soc. paper.

537.531 + 539.165 77
Calorimetric Experiment on the Radiation Losses of 2-Mev Electrons.—W. W. Buechner & R. J. Van de Graaff. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 174–177.)

538.691 78
Apparatus showing the Path of an Electrified Particle in a Magnetic Field.—J. Loeb. (*C.R. Acad. Sci., Paris*, 25th Feb. 1946, Vol. 222, No. 9, pp. 488–490.) If a light flexible wire is placed in any given magnetic field, the ratio of the (direct) current carried to the tension can be chosen so that it assumes the shape of the trajectory of a given charged particle in the same field. For example, a silver wire of diameter 2.10^{-3} cm carrying a current of 500 mA and with a tension of 18 dynes represents a particle charged to 10 000 V. The method can be applied to the study of (*a*) the motion of charged particles in the earth's magnetic field, (*b*) finding pairs of image points for magnetic electron lenses, and (*c*) the movement of ions in apparatus used in nuclear chemistry.

539.16.08 79
A Report on the Wilson Cloud Chamber and Its Applications in Physics.—N. N. Das Gupta & S. K. Ghosh. (*Rev. mod. Phys.*, April 1946, Vol. 18, No. 2, pp. 225–290.) A comprehensive account of recent developments in the design and use of the apparatus, with particular reference to its application to the study of cosmic rays. The physics of drop formation is also discussed. There is an extensive bibliography.

546.3 + 669 80
Electrons and Metals : Part 2 — The Nature of a Metal.—Hume-Rothery. (*See* 145.)

530.145 (02) 81
Philosophic Foundations of Quantum Mechanics. [Book Review]—Reichenbach. (*See* 321.)

[537 + 538](075.3) 82
Principles of Physics. Vol. 2 : Electricity and Magnetism. [Book Review]—F. W. Sears. Addison-Wesley Press, Cambridge, Mass., 1946, 434 pp., \$5.00. (*Science*, 2nd Aug. 1946, Vol. 104, No. 2692, pp. 112–113.) Written for a two-year elementary course. "A genius for clear explanations runs through . . . the whole series . . . The usual welter of units and viewpoints is brought here into a lucid, teachable orderliness." Vols 1 and 3 were published in 1945.

537.591(042) 83
Kosmische Strahlung. [Book Review]—W. Heisenberg (Ed.). Springer, Berlin, 1943. (*Schweiz. Arch. angew. Wiss. Tech.*, March 1946, Vol. 12, No. 3, p. 104.) Reprints of lectures given at the Max Planck Institute, Berlin. ". . . in highly concentrated form the best review of the properties of cosmic radiation hitherto published."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA.

523.2 : 621.396.9 84
Radar Measurement of Inter-Planetary Distances.—(*See* 105.)

523.2 : 621.396.9 : 621.396.1 85
Astronomical Radar.—Clarke. (*See* 106.)

523.72 : 621.396.822.029.62 86
Circular Polarization of Solar Radio Noise.—E. V. Appleton & J. S. Hey. (*Nature, Lond.*, 7th Sept. 1946, Vol. 158, No. 4010, p. 339.) Application of the magneto-ionic theory to the radiation of electromagnetic waves from sunspots shows that such radiation (solar noise) should be circularly polarized. This result was experimentally verified for noise from sunspots on a frequency of 85 Mc/s. See also 323 of 1946 (Appleton) and 1825 of 1946 (Hey : Stratton) and 87 and 89 below.

523.72 : 621.396.822.029.62 87
Origin of Solar Radiation in the 1–6 Metre Radio Wave-Length Band.—K. O. Kiepenheuer. (*Nature, Lond.*, 7th Sept. 1946, Vol. 158, No. 4010, p. 340.) It is claimed that the oscillation of electrons in circular orbits under the influence of the magnetic field in the neighbourhood of a sunspot is sufficient to explain the increased radiation of solar noise from these regions. See also 86 and 89.

523.72 : 621.396.822.029.62 88
Polarization of Solar Radio-Frequency Emissions.—D. F. Martyn. (*Nature, Lond.*, 31st Aug. 1946, Vol. 158, No. 4009, p. 308.) Observations were made at Canberra on 200 Mc/s of the large sunspot group at the end of July 1946. Four Yagi arrays in two perpendicular sets spaced by $\lambda/4$ in the line of sight were used to accept only one sense of circular polarization. The (power) ratio of right-handed polarization to left-handed was about 7 : 1 before the group reached the solar meridian, but became 1 : 5 after the group had crossed it. Magneto-ionic theory shows that in both cases, the "extraordinary" ray is stronger than the "ordinary". See also 1823 of 1946 (Pawsey, Payne-Scott & McCready) and 3252 of 1945 (Southworth).

523.72 : 621.396.822.029.62 89
Solar Radiation on 175 Mc/s.—M. Ryle & D. D. Vonberg. (*Nature, Lond.*, 7th Sept. 1946, Vol. 158, No. 4010, pp. 339–340.) By the use of two aerial systems spaced several wavelengths apart, solar radiation on a frequency of 175 Mc/s has been observed even on days of low sunspot activity. The method is analogous to Michelson's inter-

ference method of measuring stellar diameters. On the occasion of the passage of a large sunspot, the approximate diameter of the source of radiation was found to correspond with that of the visual spot. The circular polarization of the radiation from the spot was confirmed. See also 86 and 87 above, and 1823 of 1946 (Pawsey, Payne-Scott & McCready).

523.74/.75] : 550.385 **90**

Magnetic Storms and Solar Activity, 1945.—(*Observatory*, Feb. 1946, Vol. 66, No. 830, p. 225.) Statement of sunspot numbers, positions, and sizes, with number and intensity of solar flares. Eleven geomagnetic storms were recorded; there was little tendency for recurrence at 27-day intervals as in the recent sunspot minimum epoch.

523.746 : [550.385 + 621.396 **91**

Sunspots and Radio.—H. Spencer Jones. (*Observatory*, Aug. 1946, Vol. 66, No. 833, pp. 326–327.) Lecture by the Astronomer Royal on the history and characteristics of sunspots, and the consequent radio fade-out, radio noise, and magnetic storms.

523.747 : 550.38 **92**

On the Coincidence of Short Period Magnetic Activity and the Appearance of Faculae on the Sun.—M. Burgaud. (*C. R. Acad. Sci., Paris*, 4th March 1946, Vol. 222, No. 10, pp. 563–564.) It is suggested, as a result of examining past records, that geomagnetic disturbances of gradual commencement are connected with the growth, or rather the birth, of faculae. Tables are given comparing geomagnetic disturbances with the appearance of calcium flocculi for Jan./Feb. 1925, and for 1920–1931.

537.591 **93**

Cosmic Rays and Their Origin.—A. C. B. Lovell. (*Endeavour*, April 1946, Vol. 5, No. 18, pp. 74–79.) "This article gives a brief description of recent work and of previous experiments." The nature of the incident radiation from space is discussed together with the ensuing collision processes.

537.591 **94**

Cosmic Radiation above 40 Miles.—S. E. Golian, E. H. Krause & G. J. Perlow. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 223–224.) Cosmic-ray data at heights of 200 000 to 350 000 ft have been obtained by apparatus contained in a V2 rocket. A number of counters in the warhead were made to transmit results to the ground by multichannel radio equipment. The counters were arranged to provide data on showers and coincidences, one set being shielded by lead. Provisional results are given.

537.591 **95**

On the Production of Penetrating Ionizing Particles by the Non-Ionizing Component of Cosmic Radiation.—P. J. G. de Vos & S. J. du Toit. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 229–230.)

537.591 **96**

The Power Spectrum of the Cosmic-Ray Cascade Component.—E. P. Ney. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 221–222.)

539.16.08 **97**

A Report on the Wilson Cloud Chamber and Its Applications in Physics.—Das Gupta & Ghosh. (*See* 79.)

550.38 **98**

Induction Effects in Terrestrial Magnetism: Part 2—The Secular Variation.—W. M. Elsasser. (*Phys. Rev.*, 1st/15th Aug. 1946, Vol. 70, Nos. 3/4, pp. 202–212.) For part 1 see 1834 of 1946; see also 958 of 1942.

550.38 **99**

On the Origin of Terrestrial Magnetism.—Y. P. Bulashevich. (*Bull. Acad. Sci. U.R.S.S., sér. géogr. géophys.*, 1944, Vol. 8, Nos. 2/3, pp. 93–95. In Russian.) According to Haalck's theory the existence of temperature and pressure gradients in the crust of the earth causes a partial movement of electrons from the central part of a metallic nucleus towards its periphery. The rotation of the redistributed charges, owing to the diurnal rotation of the earth, causes the appearance of the magnetic field. A formula is given for the magnetic moment M in which a coefficient β is determined by equating M to an observed value. Haalck's theory is based on erroneous conceptions of the behaviour of electrons in a metal and if β is calculated from theoretical considerations instead of an empirical comparison, the formula for M will give a value 10^{14} times too low.

551.509 **100**

Weather Forecasting.—H. B. Brooks. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 84–87.) A brief account of the methods and equipment used by the U.S. Army for the measurement of wind velocity at high levels in the atmosphere, and of the application of centimetre-wave radar apparatus to storm detection. The importance of these techniques in weather forecasting and in reducing flying risks is indicated.

551.510.535 : 621.396.11 **101**

The Effect of the Ionosphere on Radio Communication.—McNicol. (*See* 219.)

624.13 **102**

Soil Compaction, Moisture, and [load] **Bearing Value.**—A. H. Gawith. (*J. Instn Engrs Aust.*, June 1946, Vol. 18, No. 6, pp. 109–115.)

LOCATION AND AIDS TO NAVIGATION

519.2 **103**

On the Location of a Point on a Plane by Cross Bearings from Three Known Points.—M. I. Yudin. (*Bull. Acad. Sci. U.R.S.S., sér. géogr. géophys.*, 1944, Vol. 8, Nos. 2/3, pp. 96–102. In Russian.) Experience has shown that in plotting on a chart the position O of a point in a plane by taking bearings from different known points, the most effective results are obtained when these observations are taken from three such points. On account of errors in observations, the straight lines drawn from the three points will not intersect at a single point, but form a small triangle. In the present paper equations are derived for determining the most probable position O′ of the observed point within the area of the triangle, and for calculating the quadratic error of such a determination. This is compared with the quadratic error for the case when two observations only are made. The advantages of using three observation points are discussed and methods are indicated for the most rational selection of the observation points. In conclusion, a graphical method for determining O′ is described.

The method discussed has wide applications in artillery practice and in geodesic surveys. It is also used in radio direction finding and for various other purposes, such as the determination of the wind from three angles of drift of an aeroplane.

For an extension of this paper, see 2959 of 1946.

621.396.9 **104**

The Evolution of Radiolocation.—R. A. Watson-Watt. (*J. Instn elect. Engrs*, Part I, Sept. 1946, Vol. 93, No. 69, pp. 374–382.) A general historical survey of British radar development, delivered at the I.E.E. Radar Convention, March 1946. Fundamental scientific research, the British radio industry, and the needs of the Royal Air Force all vitally affected radar development; the interplay of operational and technical experience and opinion, and close collaboration between scientists and Service users might be called our real secret weapon. A few 'technical milestones' briefly described were: monostatic working; first radar responder system; rotating beams; precision range- and direction-finding; airborne and shipborne radar; the plan position indicator; common aerial for transmission and reception; the 'memory tube'; hyperbolic navigation; development of centimetre technique; terrain discrimination; unorthodox visibility and radar detection of clouds; and location of V2 sites.

In conclusion, tribute is paid to the radio industry's achievements under 'crash programme' conditions, and some outstanding land, sea, and air uses of radar as a weapon of war are mentioned.

621.396.9 : 523.2 **105**

Radar Measurement of Inter-Planetary Distances.—(*Observatory*, Feb. 1946, Vol. 66, No. 830, pp. 193–194.) Successful radar detection of the moon suggests the use of radar to measure interplanetary distances more accurately. Milne's kinematic relativity theory assumes in effect that astronomical distances are thus measured.

621.396.9 : 523.2 : 621.396.1 **106**

Astronomical Radar.—A. C. Clarke. (*Wireless World*, Oct. 1946, Vol. 52, No. 10, pp. 321–323.) A description of U.S. Signals Corps work on radar echoes from the moon using a parabolic aerial array, and a discussion of the possibility of obtaining echoes from the nearer planets or the sun by pulse methods using either microwave or optical frequencies. Future applications of lunar reflection for point-to-point radio transmission are suggested.

621.396.9 : 534.321.9 **107**

Radar in Nature.—Roddam. (*See* 4.)

621.396.9 : 621.385.832 **108**

Skiatron.—(*See* 302.)

621.396.93 **109**

Naval Airborne Radar.—L. V. Berkner. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, pp. 671–706.) "Requirements in the early months of the war were fulfilled by meter-wave radar using lobe-switching techniques. The development of microwave radar gave enormous impetus to airborne applications and produced advanced types for air-to-air interception, high- and low-altitude bombing, reconnaissance, submarine search, and many specialized applications. The sharp beam produced by microwave radiation with relatively small antennas makes microwave techniques particularly adaptable to aircraft use and provides a vastly improved display permitting installation with low aerodynamic drag. Several types of airborne radar are briefly described and illustrated. Fundamental problems of design are reviewed. Related problems such as size, weight, and performance at high altitude are considered and solutions are discussed. Several types of display particularly suited to aircraft use, such as PPI, B, O, and G are illustrated. Utilization, applications, and advantages of auxiliary devices, such as computers, beacons, delay circuits, etc., are discussed. Solutions to systems problems introduced by use of a multiplicity of electronic gear within the aircraft are reviewed. The limitations and advantages of airborne radar as a solution to future aircraft problems are briefly considered."

621.396.93 **110**

Airborne Search Radar.—J. H. Cook. (*Bell Lab. Rec.*, Sept. 1946, Vol. 24, No. 9, pp. 321–325.) The American ASH or AN/APS-4 airborne radar equipment has a bomb-shaped unit containing the aerial, power circuits, transmitter and receiver. It weighs 150 lb (with associated units) and hangs in a standard bomb-rack under the wing of the aircraft. The wavelength is 3.2 cm, and beam width 6°. The beam is scanned horizontally over 150° once or twice per second. The equipment is used for ground scanning or to determine the relative elevation of another aircraft. Reports of its operational use are included.

621.396.93 **111**

Radar for Carrier-Based Planes.—C. B. Barnes. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 100–105.) Details of the APS-4 light-weight radar equipment operating on about 9 375 Mc/s. The equipment is suitable for search and interception roles while provision is also made for beacon operation. In each application type B display is used (slant range/azimuth on a Cartesian graticule); in interception operation there is special provision for simultaneous indication of the angle of elevation of the target on the display tube.

621.396.933 **112**

Radio Aids to Civil Aviation.—(*Engineering, Lond.*, 13th Sept. 1946, Vol. 162, No. 4209, p. 254.) An editorial dealing with the visit of delegates of the Provisional International Civil Aviation Organization to this country to inspect British equipment already in operation for controlling aircraft, and other experimental apparatus. Methods demonstrated included short-range supervision of flight and control of aircraft outside the approach zone, using very high frequency voice telephony and long-distance communication associated with teleprinters. Demonstrations were also given of radar methods of aircraft control, Babs and Gee. For another account see *Electrician*, 13th Sept. 1946, Vol. 137, No. 3563, p. 718.

621.396.933.2 **113**

On the Error in the Determination of the Median Plane of a Radio Beacon in a Tilted Airplane.—K. F. Niessen. (*Philips Res. Rep.*, April 1946, Vol. 1, No. 3, pp. 161–168.) The electric field from a vertical radiator situated on the ground has a radial as well as a vertical component at an elevated receiving point such as an aircraft. The effect of this on the accuracy of air navigation systems having spaced vertical radiators is considered. It is shown that bearing errors will occur if the aircraft aerial responds to non-vertical electric fields, *e.g.* when the aircraft banks.

621.396.9(02) **114**

Introduzione alla Radiotelemetria (Radar). [Book Review]—U. Tiberio. Editore Rivista Marittima, Rome, 1946, 277 pp., 300 lire. (*Nature, Lond.*, 31st Aug. 1946, Vol. 158, No. 4009, pp. 288–289.) ". . . an account of the Italian research on radar from 1935 until the end of hostilities. The treatment [of elementary principles] is simple and lucid."

621.396.932(02) **115**

Radio Aids to Navigation. [Book Notice]—[U.S.] Hydrographic Office Publ. No. 206, 1946, 463 pp., $2.00. (*U.S. Govt Publ.*, June 1946, No. 617, p. 667.)

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 **116**

An Easily Constructed All-Metal Vacuum Gauge.—R. T. Webber & C. T. Lane. (*Rev. sci. Instrum.*, Aug. 1946, Vol. 17, No. 8, p. 308.)

531.788.7 **117**

Radio-Active Ionization Gauge.—National Research Corporation. (*J. sci. Instrum.*, Oct. 1946, Vol. 23, No. 10, p. 247.) The replacement of the usual hot filament by a radium pellet as a source of ionization results in a smaller liability to zero drift and the coverage of a greater range of pressure. There is a loss of sensitivity at low pressures.

533.5 **118**

A Vacuum 'Lead-in'.—H. Herne. (*J. sci. Instrum.*, Oct. 1946, Vol. 23, No. 10, p. 244.) A current of up to 50 A had to be passed through a metal base-plate into a high vacuum chamber. A metal-rubber anti-vibration bush was used as an insulator; the outer metal cylinder of the bush was soft-soldered to the base plate, and the inner one to a solid conductor which would carry the required current.

533.5 : 621.3.032.53 **119**

Theory and Practice of Glass-Metal Seals: Parts 1-3.—A. J. Monack. (*Glass Ind.*, Aug.–Oct. 1946, Vol. 27, Nos. 8–10, pp. 389..420, 446..476 & 502..528.)

535.37 : 535.61–15 **120**

Development of Infra-Red Sensitive Phosphors.—B. O'Brien. (*J. opt. Soc. Amer.*, July 1946, Vol. 36, No. 7, pp. 369–371.)

535.37 : 535.61–15 **121**

On Infra-Red Sensitive Phosphors.—F. Urbach, D. Pearlman & H. Hemmendinger. (*J. opt. Soc. Amer.*, July 1946, Vol. 36, No. 7, pp. 372–381.)

535.37 : 535.61–15 **122**

Preparation and Characteristics of Zinc Sulfide Phosphors Sensitive to Infra-Red.—G. R. Fonda. (*J. opt. Soc. Amer.*, July 1946, Vol. 36, No. 7, pp. 382–389.)

535.371.07 : 621.385.832 **123**

Long Persistence C.R. Tube Screens.—R. Feldt. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 70–71.) A comparison of the persistence of the trace in DuMont tubes with the P^2 screen (green single layer) and the P^7 screen (blue ZnS.Ag and yellow ZnCdS.Cu) under various conditions of voltage, writing speed and ambient illumination. Curves and a table of results are given.

537.226 : 62 **124**

Theoretical Physics [of dielectrics] **in Industry.**—H. Fröhlich. (*Nature, Lond.*, 7th Sept. 1946, Vol. 158, No. 4010, pp. 332–334.) The behaviour of electrons in crystalline solids is discussed theoretically with particular reference to the properties of dielectrics. In insulators the occupied energy bands in the atoms are completely filled, while conduction in metals is associated with incompletely filled bands. Dielectric breakdown occurs when electrons are continuously raised to higher energy levels until internal ionization occurs. Dielectric strength increases with temperature up to a critical temperature at which collisions between electrons become important; dielectric strength decreases with further temperature increase. Dielectric loss is due to the phase lag between the motion of elementary dipoles and the applied field.

537.226.3 + 621.315.611.011.5 **125**

The Relation between the Power Factor and the Temperature Coefficient of the Dielectric Constant of Solid Dielectrics: Part 1.—M. Gevers. (*Philips Res. Rep.*, April 1946, Vol. 1, No. 3, pp. 197–224.) "The ratio of the temperature coefficient to . . . $\tan\delta$ at a given temperature and frequency is nearly the same for most solid dielectrics." This cannot be explained by existing theories, such as those of Pellat, von Schweidler, Wagner, Debye, Gyemant and others which are summarized. A bibliography of 44 items is given. Later articles will describe the experimental technique, and give a theoretical explanation of the result quoted.

538.22 : 546.77 **126**

Magnetic Anisotropy of Molybdenite at Different Temperatures.—A. K. Dutta. (*Indian J. Phys.*, Dec. 1945, Vol. 19, No. 6, pp. 225–234.)

538.652 : 62 **127**

Magnetostriction in Industry Processes.—F. Sloane. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 74–76, 101.) A general survey of the physical principles, and of magnetostriction oscillators.

538.662.13 : 537.228.1 **128**

The Lower Curie Point of Ferro-Electric Salts.—H. M. Barkla. (*Nature, Lond.*, 7th Sept. 1946, Vol. 158, No. 4010, pp. 340–341.) In a constant electric field the electric moment of ferro-electric salts analogous to potassium dihydrogen phosphate remains unchanged in passing through the lower Curie point.

546.287 **129**
Take a Grain of Sand.—H. C. E. Johnson. (*Sci. Amer.*, Sept. 1946, Vol. 175, No. 3, pp. 105–107.) Elementary survey of silicones and their uses in liquid, rubber or resin form.

546.621 : 620.193.2 **130**
Study of the Oxidation of Aluminium by Air at Ordinary Temperatures, by measuring the Potential of [electrolytic] **Dissolution.**—P. Morize & P. Lacombe. (*C. R. Acad. Sci., Paris*, 18th March 1946, Vol. 222, No. 12, pp. 658–659.) For pure electrolytically-polished aluminium the potential is − 1.20 V. Exposure to air causes a fall which increases with time at a rate dependent on atmospheric humidity: the value of the potential indicates the thickness of the oxide layer.

546.72 **131**
Preparation and Magnetic Properties of the Compound Fe_4N.—C. Guillaud & H. Creveaux. (*C. R. Acad. Sci., Paris*, 13th May 1946, Vol. 222, No. 20, pp. 1170–1172.)

551.582 : 620.19 **132**
Climate and the Deterioration of Materials.—C. E. P. Brooks. (*Quart. J.R. met. Soc.*, Jan. 1946, Vol. 72, No. 311, pp. 87–97.) The relation of air temperature to that of exposed surfaces or containers is explained. Temperature and moisture effects on materials are considered in general terms, and a rough index of their relative seriousness in various regions is plotted on a world chart.

620.197(213) : 621.396.6.045 **133**
Tropicalizing [transformers and chokes].—O. P. Scarff. (*Wireless World*, Sept. 1946, Vol. 52, No. 9, pp. 312–313.) See also 134 below.

620.197(213) : 621.314.045 **134**
Impregnated Windings [for tropicalizing transformers].—T. Williams: R. Burkett. (*Wireless World*, Oct. 1946, Vol. 52, No. 10, pp. 345–346.) Correspondence on 133 above.

621.314.63 **135**
Metal Rectifier Developments — Possible Applications of Titanium Dioxide.—Henisch. (*See* 248.)

621.315[.211.2 + .22 **136**
Mineral-Insulated Metal-Sheathed Conductors.—Tomlinson & Wright. (*See* 12.)

621.315.6 + [621.39 : 371.3 **137**
I.E.E. Radio Section Address: Part 1 — Training Courses; Part 2 — Dielectric Developments.—Jackson. (*See* 314.)

621.315.61 **138**
The Transformation of Anatase into Rutile.—T. Nguyen. (*C.R. Acad. Sci., Paris*, 13th May 1946, Vol. 222, No. 20, pp. 1178–1179.)

621.315.612.3.029.6 **139**
Steatite for High Frequency Insulation.—J. M. Gleason. (*J. Brit. Instn Radio Engrs*, Jan./Feb. 1946, Vol. 6, No. 1, pp. 20–32.) Reprint of 3059 of 1945.

621.315.617.3 **140**
Film-Forming Materials used in Insulation.—(*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, p. 344.) Report of I.E.E. Radio Section discussion led by C. R. Pye: for other accounts see 641 and 352 of 1946.

621.316.86 : 546.281.26 **141**
Silicon Carbide Non-Ohmic Resistors.—F. Ashworth, W. Needham & R. W. Sillars. (*J. Instn elect. Engrs*, Part I, Sept. 1946, Vol. 93, No. 69, pp. 385–401. Discussion, pp. 401–405.) An integrating paper, discussing the properties and construction of these resistors, and the characteristics of single contacts between silicon carbide crystals. A method of calculating currents and voltages in circuits involving these resistors is given in an appendix, and their uses, limitations, and specification are considered.

621.357.1 : 620.192.43 **142**
Electrolytic Detection of Small Amounts of Lead in Brass or Zinc.—D. McLean. (*Nature, Lond.*, 31st Aug. 1946, Vol. 158, No. 4009, p. 307.) The local cell set up between the lead and the groundmass during electrolytic polishing produces a 'moat' around the lead particles which can be identified microscopically.

621.357.8 : 669.018.2.21 **143**
Anodizing and Its Uses in Engine Construction.—N. D. Tomashov. **After Treatment of Aluminium-Alloy Castings.**—E. Carrington. **Anodizing — A Commentary.**—P. Smith. (*Light Metals*, Aug. & Oct. 1946, Vol. 9, Nos. 103 & 105, pp. 429–438, 439–450 & 515–516.)

666.1 : 62 **144**
New Glass Compositions for Industry.—(*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, p. 299.) Note on recent work by the British Thomson-Houston Co. on glasses for the lamp and radio valve industry, including "C40" for sealing to Kovar, a leadless glass for sealing to iron, and phosphate glasses. Expansion viscosity and annealing properties are also being fundamentally investigated.

669 + 546.3 **145**
Electrons and Metals: Part 2 — The Nature of a Metal.—W. Hume-Rothery. (*Metal Ind., Lond.*, 25th Oct. 1946, Vol. 69, No. 17, pp. 343–346.) Twenty-fifth instalment of a series; written as a discussion between a young scientist and an older metallurgist on co-valent bonds.

669 : 061.6 **146**
Metallurgical Research.—(*Electrician*, 25th Oct. 1946, Vol. 137, No. 3569, pp. 1139–1140.) An account of a new laboratory built at the Bilston works of Joseph Sankey Ltd. for research on cold-rolling processes, the fundamental nature of ferromagnetism, testing of sheet steels, and production control.

669.295/.296 **147**
Titanium and Zirconium.—W. J. Kroll & A. W. Schlechten. (*Metal Ind., Lond.*, 18th Oct. 1946, Vol. 69, No. 16, pp. 319–322.) Properties and extraction methods.

669.35.5.55 **148**
Electrical Resistance Alloy.—(*Engineering, Lond.*, 30th Aug. 1946, Vol. 162, No. 4207, p. 211.) The

physical properties of a new copper base resistance alloy 'Kumanal'. Its specific resistance is 41×10^{-6} Ω.cm and varies little between 20°C and 350°C.

679.5 149
Polytetrafluoroethylene.—M. M. Renfrew & E. E. Lewis. (*Industr. Engng Chem.*, Sept. 1946, Vol. 38, No. 9, pp. 870–877.) A full account of the electrical, mechanical and chemical properties of this new plastic. It will withstand 300°C and is not brittle at low temperatures. The dielectric constant is 2.0 and the power factor less than 0.000 2 at frequencies up to 3 000 Mc/s. When subject to an electric arc the plastic does not form a conducting carbon track but is reduced to a volatile gas. It is used for h.f. electrical insulation and in equipment for handling hot corrosive liquids.

679.5 150
Progress in Plastics: Parts 1 & 2.—A. E. Williams. (*Engineer, Lond.*, 6th & 13th Sept. 1946, Vol. 182, Nos. 4730 & 4731, pp. 206–207 & pp. 229–232.)

681.2.085 : 621.972.6 151
Improved Methods of Illuminating Instrument Dials.—H. Huxley. (*J. sci. Instrum.*, Oct. 1946, Vol. 23, No. 10, pp. 234–237.) Methods involving the leading of light in perspex can be used to give maximum uniformity of dial illumination with minimum extraneous light.

621.315.614.72(02) 152
Varnished Cloths for Electrical Insulation. [Book Review]—H. W. Chatfield & J. H. Wreddon. J. & A. Churchill, London, 1946, 255 pp., 21s. (*Gen. elect. Rev.*, Sept. 1946, Vol. 49, No. 9, p. 62; *Wireless World*, Sept. 1946, Vol. 52, No. 9, p. 307.) One of the authors is employed by a varnish manufacturer, and the other by an electrical manufacturer. See also 3648 of 1946.

MATHEMATICS

512.52 153
Interpolation with the Aid of a Plot of First Differences.—G. S. Fulcher. (*J. appl. Phys.*, July 1946, Vol. 17, No. 7, pp. 617–628.)

518.5 154
Differential Analyzer.—(*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 62–66..100.) A general description, with photographs, of the Massachusetts Institute of Technology equipment. Mechanical integrators are used with servo-operated capacitor bridges for transmitting the information electrically between integrators. Punched tape is used for automatic control of the equipment. For a full description see 361 of 1946.

518.61 155
Calculation of the Electromagnetic Field, Frequency and Circuit Parameters of High-Frequency Resonator Cavities.—Motz. (*See* 52.)

518.61 156
Calculation of the Magnetic Field in Dynamo-Electric Machines by Southwell's Relaxation Method.—H. Motz & W. D. Worthy. (*J. Instn elect. Engrs*, Part II, Aug. 1946, Vol. 93, No. 34, pp. 379–382.) Discussion of 658 of 1946.

519.2 157
On the Location of a Point on a Plane by Cross Bearings from Three Known Points.—Yudin. (*See* 103.)

519.2 158
The Resultant of a Large Number of Events of Random Phase.—Domb. (*See* 278.)

519.241.6 : 621.3 159
An Extension of Campbell's Theorem of Random Fluctuations.—Rivlin. (*See* 29.)

534.014.2 : 621.396.611 160
An Experimental Investigation of Forced Vibrations in a Mechanical System having a Non-Linear Restoring Force.—Ludeke. (*See* 53.)

621.317.727 : 514 161
A Simple Potentiometer Circuit for Production of the Tangent Function.—R. Hofstadter. (*Rev. sci. Instrum.*, Aug. 1946, Vol. 17, No. 8, pp. 298–300.)

51(075) : 621.396 162
Basic Mathematics for Radio Students. [Book Review]—F. M. Colebrook. Iliffe & Sons, London, 1946, 270 pp., 10s. 6d. (*Electronic Engng*, Sept. 1946, Vol. 18, No. 223, p. 293.) ". . . the subject is presented with the lucidity we expect from this well-known contributor to radio technical journals." See also 3338 of 1946.

518.2(083) : 517.564.4 163
Tables of Associated Legendre Functions. [Book Review]—Mathematical Tables Project. Columbia Univ. Press, New York, 1945, 303 pp., $5.00. (*Phil. Mag.*, Oct. 1945, Vol. 36, No. 261, pp. 729–730.) See also 1279 of 1946.

519.21(02) 164
An Experimental Introduction to the Theory of Probability. [Book Review]—J. E. Kerrich. Einar Munksgaard, Copenhagen, 1946, 98 pp., 8.50 kroner. (*Nature, Lond.*, 14th Sept. 1946, Vol. 158, No. 4011, p. 360.)

MEASUREMENTS AND TEST GEAR

538.12 : 621.3.08 165
Production of Uniform and Constant Magnetic Fields for Measurement Purposes: Parts 1 & 2.—H. Neumann. (*Arch. tech. Messen*, Nov. & Dec. 1940, Nos. 113 & 114, pp. T128–129 & T138–139.) Parts 3 & 4 were noted in 2823 of 1942.

538.214.082.1 166
Apparatus for Measuring Magnetic Moments.—G. N. Rathenau & J. L. Snoek. (*Philips Res. Rep.*, April 1946, Vol. 1, No. 3, p. 239.) The specimen whose susceptibility is required is suspended in a magnetic field such that the restoring force on the specimen for small displacements is proportional to the displacement. Measurement of the period of oscillation with and without the magnetic field enables the susceptibility to be deduced.

549.514.51 : 620.1 167
Quartz Crystal Testing Instrumentation.—D. S. Dickey. (*Instruments*, Jan. 1946, Vol. 19, No. 1, pp. 9–11.) Describes the use of recording instruments for production testing of the variation of crystal frequency and activity with temperature.

621.317 **168**
The History and Development of the British Scientific Instrument Industry.—Barron. (*See* 315.)

621.317 : 621.396.6.004.67(73) **169**
Measuring Equipment in American Radio Repair Workshops : Parts 1 & 2.—G. Keinath. (*Arch. tech. Messen*, Nov. & Dec. 1940, Nos. 113 & 114, pp. T120 & T132–133.) The noteworthy features of this equipment are (i) flexibility, (ii) limited accuracy, (iii) convenience in use, (iv) portability, and (v) cheapness. The article considers in some detail with diagrams and circuits : (*a*) high resistance d.c. voltmeters, (*b*) capacitance meters, (*c*) valve testers, (*d*) vibrator testers, (*e*) test oscillators, (*f*) universal receiver tester and (*g*) oscilloscopes.

621.317.2 : 621.396.623 **170**
Notes on Field Laboratory Design.—Matthews. (*See* 228.)

621.317.2 : 621.397.5 **171**
TV [television] **Test Equipment.**—Hunter. (*See* 268.)

621.317.2 : 621.397.5 **172**
A Television Pattern Test Generator.—Inskip. (*See* 267.)

621.317.3 : 621.392 **173**
Waveguide Measurements.—G. Ashdown. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 318–319.) A description of a 10 000-Mc/s waveguide test-bench whose main components are : a short length of waveguide carrying a klystron oscillator, a launching aerial, a 10-db attenuator, a coaxial-line wavemeter and a standing-wave detector.

621.317.33 : 621.315.21.029.4/.6 **174**
Characteristics of R.F. Cables.—N. C. Stamford & R. B. Quarmby. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 295–298.) A technique of measurement at 600 Mc/s, for coaxial or twin cables, developed at Manchester University. The cable is magnetically coupled to an oscillator at one end and left open at the other. The variations of input current, measured by a thermojunction, are observed as short lengths are cut from the open end. The phase constant is then determined from the lengths at which successive minima of current occur. The attenuation constant is found by comparing the outputs of short and long line lengths for equal inputs when the power measuring device is matched to the line. Characteristic impedance is measured by substituting a $\lambda/4$ section of air-spaced coaxial line whose characteristic impedance is varied (by variation of the diameter of the inner conductor) until the matched condition is restored.

621.317.33.029.63 : 621.315.2 **175**
The Measurement of Cable Characteristics at Ultra-High Frequencies.—F. Jones & R. Sear. (*J. Brit. Instn Radio Engrs*, Aug./Sept. 1945, Vol 5, No. 4, pp. 154–169. Discussion, pp. 170–172.) "The paper describes the two main methods of impedance measurement at frequencies above 100 Mc/s which are in general use, and their application to the determination of cable characteristics, with particular attention to the work of the authors.

"A preliminary account is given of the various conditions in which cables can be measured, and which are applicable to both of the methods described.

"An outline of the theory of the standing wave method is given, together with a general description of the equipment required. Attention is paid to coned connectors for the attachment of the cable to the measuring line, and the errors liable to be incurred by their use.

"The theory and equipment for the resonance line method are described, and an account is given of several investigations that have been conducted in order to extend its usefulness. These include the measurement of twin cables, the effect on cable attenuation values of reactive discontinuities at the junction of measuring line and cable and at internal supports in the line, the direct determination of the characteristic impedance of measuring lines, and the radiation and reactive effects which occur at their open ends.

"The application of the resonance line method to the determination of dielectric power factor is discussed, and a method is described which permits greater accuracy than has been obtained up to the present."

621.317.372 : 621.315.2 **176**
End Leakage in Cable Power-Factor Measurement.—A. Rosen. (*J. Instn elect. Engrs*, Part II, Aug. 1946, Vol. 93, No. 34, pp. 383–386.) In d.c. measurements of the insulation resistance of cables a simple guard wire is used to eliminate the effects of leakage at the cable ends. This guard wire has been modified to make it effective for a.c. measurements when used in conjunction with a suitable bridge ; end leakage error is thereby completely avoided.

621.317.374.029.6 **177**
A Microwave Dielectric Loss Measuring Technique.—W. R. MacLean. (*J. appl. Phys.*, July 1946, Vol. 17, No. 7, pp. 558–566.) The dielectric loss is measured by determinations of the *Q*-factor of a resonator partially filled with the sample. The method is restricted to the determination of small loss factors, and requires a preliminary approximate determination of the dielectric constant. The sample is placed inside the resonator as a dielectric core which confines the field almost entirely to the dielectric. Double sample technique is used to eliminate dominant spurious losses. Detuning for the half-power points in the determination of the *Q*-factor is accomplished by large movement of a small rod, whose characteristics as a tuning element can be calculated.

621.317.374.029.6 **178**
A New Method for Measuring Dielectric Constant and Loss in the Range of Centimeter Waves.—S. Roberts & A. von Hippel. (*J. appl. Phys.*, July 1946, Vol. 17, No. 7, pp. 610–616.) The closed end of a rectangular waveguide is covered by a slab of the material, the dielectric constant of which is deduced from measurements of the standing-wave pattern in the guide, determined by means of a sliding crystal-valve probe. Certain previous sources of error are eliminated in this method. The paper gives a description of the apparatus, the theory of the method, and results for various solid and liquid dielectrics at 25°C for $\lambda = 6$ cm.

621.317.374 : 519.283 179

Quality Control [of dielectric material] **by means of H.F. Currents.**—P. Toulon. (*C. R. Acad. Sci., Paris*, 4th March 1946, Vol. 222, No. 10, pp. 543–544.) Describes methods of measurement of dielectric loss by means of a Q-meter.

621.317.384 : 621.314.2 180

A Device for the Measurement of No-Load Losses in Small Power Transformers.—L. Medina. (*Proc. Instn Radio Engrs, Aust.*, Sept. 1946, Vol. 7, No. 9, pp. 13–16.) The magnetizing component of the exciting current is cancelled by means of a variable shunt capacitance, and the no-load loss current is measured by a rectifier-type microammeter in conjunction with a filter arrangement tuned to the fundamental frequency of the supply voltage. It is a method for production testing and the accuracy is 5–10% ; circuit details are given for a practical instrument.

621.317.7 181

Developments in Electrical Measuring Instruments.—(*Engineering, Lond.*, 6th Sept. 1946, Vol. 162, No. 4208, pp. 234–235.) Comment on 3351 of 1946.

621.317.7.082.7.029.63 : 621.315.212 182

Standing-Wave Indicator.—G. E. Feiker. (*Gen. elect. Rev.*, Sept. 1946, Vol. 49, No. 9, pp. 43–46.) The indicator consists of a slotted section of coaxial line with an adjustable outer conductor so that a probe may be driven by a rack and pinion arrangement along the line. Its operating frequency is of the order of 3 000 Mc/s. Methods are explained for using it to measure (*a*) standing wave ratio, (*b*) complex impedance, (*c*) net power flow, (*d*) attenuation.

621.317[.72 + .784 183

A Precision A.C./D.C. Comparator for Power and Voltage Measurements.—G. F. Shotter & H. D. Hawkes. (*J. Instn elect. Engrs*, Part II, Aug. 1946, Vol. 93, No. 34, pp. 314–319. Discussion, pp. 320–324.) The sources of error common to dynamometer wattmeters are briefly reviewed, and a new instrument for measuring a.c. power and voltage by direct comparison with a standard d.c. potentiometer is described.

621.317.727 184

Self Balancing Potentiometer.—(*Electronic Ind ustr.*, Oct. 1946, Vol. 5, No. 10, p. 79.) A 'slide back' arrangement using a galvanometer and a double photocell in a valve bridge circuit, drawing less than 0.01 μA from the source. It measures potentials between 100 μV and 1 V ; it will produce a maximum drop of 10 V across an output load of 2 000 Ω.

621.317.733 : 621.326 185

A Method of Measuring the Current Distortion and Phase-Angle due to a Non-Linear Impedance.—G. M. Petropoulos. (*Beama J.*, Sept. 1946, Vol. 53, No. 111, pp. 320–323.) A sinusoidal voltage is applied and a suppression method using a Wheatstone bridge arrangement is employed. The non-linear impedance of incandescent lamps causes current distortion, the distortion factor and phase-angle increasing with applied voltage, but its magnitude is not considered of practical importance.

621.317.76 186

A Standard of Frequency and Its Applications.—C. F. Booth & F. J. M. Laver. (*J. Instn elect. Engrs*, Part I, Sept. 1946, Vol. 93, No. 69, pp. 417–418.) Summary of 2973 of 1946.

621.317.784.029.6 187

Air-Flow U.H.F. Watt-Meter.—Z. W. Wilchinsky & R. H. Kyser. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 128–129.) " Description of a laboratory-type instrument, suitable for calibration of general-purpose wattmeters. A tungsten-filament dissipative element [in an evacuated envelope] is inserted as the central conductor in a section of coaxial transmission line." The operating wavelength is about 30 cm at power levels up to about 20 W. The temperature rise of the air passing the load is measured by thermocouples. Calibration at mains frequency is recommended. The chief drawback of the instrument is " that several minutes may be required for a steady state to be obtained ".

621.317.79 : 621.396.9 188

Production of Airplane Radar speeded by New Testing Technique.—F. P. Wight. (*Bell Lab. Rec.*, Sept. 1946, Vol. 24, No. 9, pp. 330–334.) Description of test apparatus and methods designed by the Western Electric Laboratories as an integral part of the production programme for certain airborne radar units. The tests included determination of wave shapes, amplitudes, frequencies, and bandwidths, and checks of general circuit performance and mechanical alignment.

621.317.794.029.6 189

Bolometers for V.H.F. Power Measurement.—E. M. Hickin. (*Wireless Engr*, Nov. 1946, Vol. 23, No. 278, pp. 308–313.) " In this method the power is dissipated in a resistor having a large temperature coefficient (an 'indicator') which forms one arm of a Wheatstone bridge. By direct-current power substitution the indicator may be calibrated and then one measurement of resistance will give the power in the load.

" Some details are given of indicators and circuits to deal with powers from a few microwatts to a few watts at frequencies up to 10 000 Mc/s. The limitations of the method and possible sources of error are discussed."

The construction of power indicators is treated in detail : the CV 95 using a 0.01-mm tungsten wire *in vacuo*, or in an inert gas to a given increased power rating, and the possible use for the filament of Wollaston wire (Pt coated with Ag), carbon or iron is described.

621.362 190

Schwarz Thermopiles.—A. Hilger Ltd. (*J. sci. Instrum.*, Oct. 1946, Vol. 23, No. 10, p. 246.) A new design, said to have unusually high sensitivity and speed and to be more robust than previous types.

621.384.5.08 191

The Properties of Glow Tubes and their Applications for Measurement Purposes.—A. Glaser. (*Arch. tech. Messen*, Dec. 1940, No. 114, pp. T136–137.)

621.392.43.082.7 192

Note on a Reflection-Coefficient Meter.—Korman. (*See* 18.)

621.396.6[illegible] 193

Specification of Receiver Sensitivity and Transmitter Power Output at Ultra-High Frequencies.—[illegible] Schwartz. (*Proc. Inst. Radio Engrs, W. & E.*, Sept. 1946, Vol. 34, No. 9, p. [illegible].) Plea for specifying receiver sensitivity in terms of decibels below 1 W and for calibrating signal generators in terms of available power. Similarly transmitter power may, for convenience, be specified in decibels above 1 W.

621.[illegible] 194

The Measurement of Time.—H. Spencer Jones. (*Wireless World*, Oct. 1945, Vol. [illegible], No. 10, pp. [illegible]–130.) Includes an account of the applications of crystal clocks at the Royal Observatory. Their advantages and disadvantages compared with other types are discussed. The need for operating crystal clocks in groups with regular intercomparison is stressed. A decimal counting chronometer with a 10 μs unit of time is briefly described. See [illegible] of 1946 (Booth).

621.396.615.11 195

Thermistor-Regulated Low-Frequency Oscillator.—[illegible] (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. [illegible].) Design considerations and detailed description of a phase-shift oscillator covering the frequency range 0.9 to 10 000 c/s in four bands. A direct-coupled cathode follower is included in the feedback chain and a separate cathode follower is used at the output stage. The feedback thermistor has a time constant of about one second, which sets the low limit to the frequency range covered.

[illegible] 196

Alternating Current Bridge Methods.—Book review. B. Hague. Pitman, London, 6th edn [illegible], 616 pp., 40s. (*Beama J.*, Sept. 1946, Vol. 53, No. 111, p. [illegible].) "This new edition, like its predecessors, should be in the hands of all engineers and students who have an interest in bridge methods."

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

[illegible] : 621.317.39 197

Electronic Micrometer for Thin Materials.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 106..108.) Permits continuous production measurement of wire diameter and strip thickness. The material passes an aperture illuminated by a scanning spot which throws a shadow on a photocell. The size of the shadow determines the voltage signal in the load resistor of the photocell.

535.61-15 198

Radioactive Infrared Detector.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 142, 146.) Radioactive material contained in the detector charges a viewing screen and makes it sensitive to incoming infrared radiation reflected on to it by a periscope device. It permits the detection of infra-red radiation but not the identification of objects.

621.314.67 : 621.385.38 : 621.365 199

Grid-Controlled Rectifiers for R.F. Heating.—[illegible] Boyd. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 125–127.) Control of the d.c. output voltage is obtained by varying the phase of an auxiliary a.c. supply on the thyratron grid. The application of this control system to a 100-kW power supply is described.

621.316.578.1 : 774 200

A Timer for Photo-Printing.—N. Phelp & F. Tappenden. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 300–301.) A two-valve electronic switch "controlled by the exponentially decreasing potential difference across the terminals of a capacitance shunted by a discharging resistance," whose value determines the exposure time.

621.316.923 : 623.95 201

Fuzes for Electronic Magnetic Mines.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 162, 166.)

621.317.39 202

The "Aquatector".—(*Electrician*, 18th Oct. 1946, Vol. 137, No. 3568, pp. 1073–1074.) Brief description of the performance of two instruments for the accurate detection and measurement of the moisture content in a wide range of solid materials and emulsions.

621.317.75 : 020.3 203

A Frequency Analyser used in the Study of Ocean Waves.—N. F. Barber, F. Ursell, J. Darbyshire & M. J. Tucker. (*Nature, Lond.*, 7th Sept. 1946, Vol. 158, No. 4010, pp. 329–332.) The records of water pressure with which the analyser is fed are in the form of a black trace of variable width on a white background. Such a record is attached to the perimeter of a wheel 30 inches in diameter which is rotated at a speed of several revolutions per second and scanned by a photocell. The amplified output from the photocell is fed to a vibration galvanometer, the motion of which is recorded on a trace forming the frequency spectrum. As the speed of rotation of the wheel is allowed to diminish continuously and naturally under the action of friction, the various frequency components present in the original record "glide" in turn through the resonant frequency of the galvanometer. Satisfactory resolution of harmonics up to the order of sixty has been achieved.

621.365.5 204

Volman-Stivin High-Frequency Induction Hardening Machines.—(*Machinery, Lond.*, 17th Oct. 1946, Vol. 69, No. 1775, pp. 498–500.) Automatic or semi-automatic machinery for hardening special alloy-steel and carbon-steel components by induction heating methods. Plant is available with powers up to 200 kW for handling work ranging from small components up to workpieces four feet in length such as crankshafts and camshafts.

621.365.5 : 621.3.018.41 205

The Effect of Frequency in Induction Heating.—R. A. Nielson. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 320–322.) A simple approximate formula is derived for the power dissipation in a cylinder heated in a long solenoid. Increasing frequency decreases the necessary current or number of turns. The limitation is occurrence of arcing.

621.38.001.8 206

Electronics — Servant or Fad?—Weiller. (See 318.)

621.38.001.8 **207**
[Electronic] **Machine-Tool Contour Controller.**—J. M. Morgan. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 92–96.)

621.38.001.8 : 551.576 **208**
Remote Cloud Indicator.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, p. 162.) Instrument developed in Britain for attaching to a barrage balloon to indicate when the balloon is flying in cloud. Strips of metal foil separated by very thin mica sheets are bridged by moisture when in cloud, thus firing a thyratron operating a transmitter.

261.38.078 : 538.56.029.6 **209**
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621.383.001.8 : 535.61–15 **210**
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621.385.38.078 **211**
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621.396.611.21 : 529.78 **212**
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621.396.619.018.41 : 621.384 **213**
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621.396.9 : 623.26 **214**
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621.396.9 : 629.135.053.2 **215**
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622.19 : 621.396 **216**
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621.365.5(02) **217**
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PROPAGATION OF WAVES

621.396.11 + 535.13 **218**
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Formulae for reflection and transmission coefficients at a boundary between two media are derived, which by use of complex quantities can be extended to conducting media, but only reduce to the Fresnel formulae when the media are dielectrics of equal permeability. The conception of the apparent permittivity of an ionized gas can be replaced by that of a uniformly distributed shunt admittance.

621.396.11.029.64 + 535.343.4 + 538.569.4.029.64] : 551.57 **219**
The Absorption of 1-cm Electromagnetic Waves by Atmospheric Water Vapor.—R. L. Kyhl, R. H. Dicke & R. Beringer. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 694.) The position (1.34 cm) and width (0.11 cm^{-1}) of the absorption line were determined using a radiometer due to R. H. Dicke. These results agree with Van Vleck's

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621.396.11 : 551.510.535 **220**

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RECEPTION

21.396/.397].621.004.67 **222**

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621.396.622 : 621.396.619.018.41 **227**

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621.396.623 : 621.317.2 **228**

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621.396.822 : 621.396.619.16 **230**

Pulse Distortion : the Probability Distribution of Distortion Magnitudes due to Inter-Channel Interference in Multi-Channel Pulse-Transmission Systems.—D. G. Tucker. (*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, pp. 323–334.) If the probability distribution of pulse distortion magnitudes in a multichannel pulse-transmission system is adequately considered, considerable economies in design may be made. An analysis of inter-channel interference distortion is given, and it is shown how to determine the probability distribution for interference of equal amplitudes from two adjacent channels on two or more links in tandem. On a typical multichannel v.f. telegraph system with two links in tandem, only about one pulse in a thousand is distorted more than half the maximum amount; the corresponding probabilities for other typical systems are much less. It is therefore evident that the basis of design should be not the maximum distortion but a fraction of it, perhaps between $\frac{1}{4}$ and $\frac{3}{4}$.

621.397.823 **231**

The Noise Suppressor in the V114 [television set].—Fairhurst. (*See* 274.)

STATIONS AND COMMUNICATION SYSTEMS

621.396.029.63/.64 **232**

Hyper-Frequency Radio.—J. M. A. Lenihan. (*J. Brit. Instn Radio Engrs*, Oct./Dec. 1944, Vol. 4, No. 3, pp. 178–186. Discussion, pp. 186–189.) A survey of problems and techniques involved at wavelengths below 30 cm, where conventional oscillators fail. The devices used include positive grid triodes, and cavity resonators. The principles of operation of these oscillators are outlined ; the most efficient and widely used are the klystron and the magnetron for which methods of modulation are given. Reception and the application of waveguides are briefly discussed, and future u.h.f. developments forecast. A selected bibliography of 33 items is given (p. 189).

621.396.1.029.6 : 523.2 **233**

Astronomical Radar.—Clarke. (*See* 106.)

621.396.4.029.6 **234**

336 Channels for V.H.F.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 150, 154.) Brief description of equipment shown to the Physical Society in 1946. 336 channels were frequency-controlled to $\pm$10 kc/s during transmission and reception by only three crystals, selection being made remotely by means of numbered or lettered dials.

621.396.619.018.41 **235**

A Review of Wide Band Frequency Modulation Technique.—C. E. Tibbs. (*J. Brit. Instn Radio Engrs*, June/Sept. 1944, Vol. 4, No. 2, pp. 85–119. Discussion, pp. 119–129.) The basic theory of frequency modulation is first outlined and the form of the frequency spectrum derived. The improvement in signal/noise ratio resulting from the use of frequency modulation is discussed and illustrated graphically, and the effects of increasing the frequency deviation, the use of pre-emphasis and the suppression of a weaker signal by a stronger one are explained.

It is shown that selective fading during propagation of a frequency-modulated signal can lead to serious distortion and that a system operating over an ionospheric path would therefore be unsatisfactory. Most frequency-modulated systems operate in the v.h.f. band with a stacked dipole or ' turnstile ' transmitting aerial array and a simple dipole receiving aerial.

Transmitter features peculiar to this form of modulation are discussed, including the modulator originally designed by Armstrong and one incorporating a reactance valve. Two types of station monitor are described, with a signal generator suitable for test work. The general design of a suitable receiver is given, with details of a typical limiter circuit, double tuned circuits, phase difference discriminators, and a tuning indicator.

A bibliography of 28 items is appended.

621.396.619.16 + 621.396.61.029.64 **236**

Army No. 10 Set.—(*Wireless World*, Sept. 1946, Vol. 52, No. 9, pp. 282–285.) Description of the u.h.f. sender, a split-anode magnetron with the segments arranged cylindrically about the cathode as axis, the receiver and the aerial system. A miniature triode is used as a local oscillator, with a crystal as the first detector, in the superheterodyne receiver. The aerial system comprises a waveguide matching section connected to a flexible waveguide, with a reflector placed before its open end, brought through the centre of a parabolic mirror. For previous articles on this set see 470 and 2706 of 1946.

621.396.619.16 **237**

Pulse Terminology.—W. A. Beatty : " Cathode Ray ". (*Wireless World*, Sept. 1946, Vol. 52, No. 9, pp. 311–312.) Discussion of the proper terminology for various types of pulse modulation, arising out of 2006 of 1946 (" Cathode Ray ").

621.396.619.16 : 621.396.822 **238**

Noise and Pulse Modulation.—T. Roddam. (*Wireless World*, Oct. 1946, Vol. 52, No. 10, pp. 327–329.) Pulse position modulation, with constant amplitude pulses and a constant number of pulses per second, would appear to give a noise-free communication system, since reception depends only on the position of the leading edge of each pulse. Considerations of bandwidth, however, show that this is untrue, and a value for the ratio of bandwidth to highest modulation frequency is obtained for the signal-to-noise ratio to be an improvement over a.m. systems. Examples of typical pulse systems show an improvement of up to 30 db. The effect of impulsive noise is also considered. See also 2006 of 1946 (" Cathode Ray ") for basic principles of pulse modulation.

621.396.619.16 : 621.396.97 **239**

Pulse Time Multiplex System tested at New York Demonstration.—(*Telegr. Teleph. Age*, Oct. 1946, Vol. 64, No. 10, pp. 15–18.) Outline description of a system employing pulse-time modulation, with notes on a recent demonstration in which eight programmes of various types (ordinary broadcast, facsimile, teleprinter, recording of music, etc.) were dealt with simultaneously. The pulse repetition frequency for each channel is 24 000 per sec. A width of 9 kc/s is available on each channel for modulation purposes; each channel may carry low definition information (such as high-speed morse) on a number of sub-channels defined by appropriate tone filters. The demonstration was carried out at a frequency of 930 Mc/s (peak power 800 W), omnidirectional and paraboloid type aerials being used respectively at the transmitter and the receiver. See also 3040 of 1946 (Grieg).

621.396.7 **240**

H.M.S. "Boxer".—G. M. Bennett. (*Wireless World*, Oct. 1946, Vol. 52, No. 10, pp. 324–326.) Fitted with equipment designed for fighter direction over sea and shore, this British warship has six high-power radar sets of various ranges with associated interrogators and beacons. Other installations include transmitters and receivers for use in all frequency bands, direction-finding equipment and a W/T homing beacon. Sets can be operated without mutual interference and data are collated in a central control room.

621.396.82 : 621.396.1 **241**

Interference Considerations affecting Channel-Frequency Assignments.—M. Reed & S. H. Moss. (*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, pp. 355–361.) A study is made of the mutual interference problems which arise when a number of stations transmitting c.w. signals and having the same frequency tolerance share a given frequency band. On the assumption that the transmitters are grouped into a number of channels spread over the frequency band, it is shown that, for a given receiver selectivity specified by its gate width, no practical advantage is gained by having a spacing of the channel frequencies less than about 75% of the nominal transmitter tolerance bandwidth, although (except over a limited region when the receiver gate is wider than the transmitter tolerance) a spacing equal to this bandwidth should not be exceeded. For a given separation between channels it is demonstrated that, in general, the interference falls with reduction of the receiver gatewidth.

621.396.97(4) **242**

Broadcasting in Europe.—(*J. Brit. Instn Radio Engrs*, Jan./Feb. & March/May 1946, Vol. 6, Nos. 1 & 2, pp. 33–40 & 41–46.) A summary of discussions held by various sections of the British Institution of Radio Engineers on a plan suggested by the Radio Industry Council in July, 1945, for a complete reallocation of broadcast frequencies throughout Europe (see 3667 of 1945).

621.396.97(058) **243**

"Broadcasting" Year Book, 1946. [Book Review]—Broadcasting Publications Inc., Washington, 580 pp. (*Wireless World*, Oct. 1946, Vol. 52, No. 10, p. 332.) A reference book containing the F.C.C. broadcasting regulations and directories of U.S., Canadian, and S. American stations.

SUBSIDIARY APPARATUS

538.652 : 62 **244**

Magnetostriction in Industry Processes.—Sloane. (See 127.)

621–526 **245**

Theory of Servo Systems, with particular reference to Stabilization.—A. L. Whiteley. (*J. Instn elect. Engrs*, Part II, Aug. 1946, Vol. 93, No. 34, pp. 353–367. Discussion, pp. 368–372.) Methods are described for achieving stability in continuous-control servo systems. Stability may be improved by the insertion of passive networks at the input end of the system to give approximations to derivatives and/or integrals of error which may be used to modify the performance characteristics. Feedback methods may often be used similarly. To assist calculations of constants of the added stabilizing networks, standard forms, which have been found to apply to widely different electric servos, are tabulated. A summary of this paper was noted in 2013 of 1946.

621–526 **246**

Dynamic Behavior and Design of Servomechanisms.—G. S. Brown & A. C. Hall. (*Trans. Amer. Soc. mech. Engrs*, July 1946, Vol. 68, No. 5, pp. 503–522. Discussion, pp. 522–524.)

621–526 : 621.313.28 **247**

A New Torque Motor.—A. E. Adams & D. Waloff. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, p. 308.) For servo applications. The normal rotating armature is replaced by gyratory motion of a low-inertia armature so that starting and stopping are almost instantaneous. The gyratory motion is transformed into rotation in the output shaft by a single stage of planetary gearing.

621.314.63 **248**

Metal Rectifier Developments — Possible Applications of Titanium Dioxide.—H. K. Henisch. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 313–315.) The three main problems are: (*a*) producing the conducting material consistently, (*b*) making the semiconductor as a thin film, and (*c*) the nature of the best electrodes.

621.315.21.029.4/.6] : 536.4 **249**

The Power Rating (Thermal) of Radio-Frequency Cables.—R. C. Mildner. (*J. Instn elect. Engrs*, Part I, Sept. 1946, Vol. 93, No. 69, p. 414.) Summary of 3066 of 1946.

621.315.668.2 **250**

Steel Tower Economics.—P. J. Ryle. (*J. Instn elect. Engrs*, Part I, Sept. 1946, Vol. 93, No. 69, pp. 407–409.) Summary of I.E.E. paper. See also 3067 of 1946.

621.316.86 : 546.281.26 **251**

Silicon Carbide Non-Ohmic Resistors.—Ashworth, Needham & Sillars. (See 141.)

621.317.755 **252**

A New Oscilloscope with D.C. Amplification.—J. H. Reyner & F. R. Milsom. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 297–299.) A general purpose instrument designed so that (*a*) all frequencies from zero to the megacycle region can be handled, (*b*) the image is sharply defined and completely steady, (*c*) instantaneous positioning at any part of the screen is provided, (*d*) any part of the trace can be expanded, (*e*) the various controls are independent, and (*f*) performance is consistent, and the instrument has long life. Methods of achieving these objectives are explained.

621.318.323.2.042.15 253
Permeability of Iron-Dust Cores.—G. W. O. H.: Lamson: Burgess. (*See* 35.)

621.318.4.017.31 : 621.316.974 254
Power Loss in Electromagnetic Screens.— Davidson, Looser & Simmonds. (*See* 33.)

621.394.652 : 621.394.141 255
A Deluxe Electronic Key.—W. R. De Hart. (*QST*, Sept. 1946, Vol. 30, No. 9, pp. 17–23.) An electronic circuit for automatic keying and monitoring purposes. The apparatus consists of two units: (*a*) a multivibrator which gives an output of dots or dashes depending on the position of an operating key; (*b*) A keying amplifier, monitoring oscillator and loudspeaker. The system has the advantage that no mechanical device is included.

621.396.615.17 256
Laboratory Pulse Generator with Variable Time Delay.—D. R. Scheuch & F. P. Cowan. (*Rev. sci. Instrum.*, June 1946, Vol. 17, No. 6, pp. 223–226.) A biased multivibrator ('flip-flop') can be adjusted so that the output pulse occurs at any desired interval between 2 μs and 850 μs after the input. Output pulse width is adjustable between 1 μs and 40 μs. Input signals of arbitrary waveform may be used at frequencies up to 100 kc/s. At 10 kc/s the minimum sine-wave input signal to operate the instrument is 0.2 V. The delayed pulse is delivered at low impedance with maximum amplitude 150 V. Block and circuit diagrams are given and fully explained.

621.396.68 : 621.397.5 257
30 kV Power Supply.—H. C. Baumann. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 77–78.) Details of a voltage trebler circuit giving 100 μA at 30 kV. It has a push-pull oscillator at 300 kc/s with a 350-V plate supply, and a r.f. transformer with separate secondaries for the high voltage and the rectifier filaments.

621.396.681 258
A Simple Battery Operated High Voltage Supply.—L. E. Williams. (*Rev. sci. Instrum.*, Aug. 1946, Vol. 17, No. 8, pp. 296–297.) The audio frequency from a blocking oscillator is transformed to a high voltage and is rectified by a diode to give an output of more than 1 000 V at 100 μA. The battery supply is 90 V at about 15 mA and a variable resistance in series with this battery adjusts the voltage output. (Appears similar to 3539 of 1940 [Burgess].)

TELEVISION AND PHOTOTELEGRAPHY.

535.241.4 259
"Foot-Lambert" Unit of Picture Brightness.—(*See* 67.)

621.385.832.032.2 260
Magnetic Focusing and Deflection.—Rawcliffe & Dressel. (*See* 303.)

621.385.832.032.2 261
Comparison of Electrostatic and Electromagnetic Deflection in Cathode-Ray Tubes.—(*See* 304.)

621.397.26 262
Electronic Newspaper.—(*Gen. elect. Rev.*, Sept. 1946, Vol. 49, No. 9, pp. 49–50.) In a trial next year, four 9½ inch by 12 inch pages of text or photographs will be relayed by f.m. broadcasting stations to facsimile receivers during transmissions lasting 15 minutes. See also 3444 of 1946.

621.397.26 263
Method of Transmitting Sound on the Vision Carrier of a Television System.—D. I. Lawson, A. V. Lord & S. R. Kharbanda. (*J. Televis. Soc.*, June 1946, Vol. 4, No. 10, pp. 239–250.) See 3091 of 1946 and back references.

621.397.26 264
A Method of Transmitting Sound on the Vision Carrier of a Television System.—D. I. Lawson, A. V. Lord & S. R. Kharbanda. (*J. Instn elect. Engrs*, Part I, Sept. 1946, Vol. 93, No. 69, pp. 415–416.) Summary of 3091 of 1946.

621.397.4 : 621.394.64.029.64 265
Facsimile over 4 000-Mc/s Relay System.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 146, 150.) An experimental two-way radio relay system, with repeater stations, has been used for transmitting various types of intelligence including facsimile transmission of text and photographs using a 4.8 kc/s bandwidth.

621.397.5 266
System Standards.—(*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 72–73.) Reference data and standards currently in use for the information and guidance of television design engineers.

621.397.5 : 621.317.2 267
A Television Pattern Test Generator.—F. A. Inskip. (*J. Televis. Soc.*, June 1946, Vol. 4, No. 10, pp. 255–256. Discussion, p. 257.) The unit is portable and is modulated to give a simple pattern on a c.r. tube screen. The tuning range is 30–60 Mc/s with calibration points at 45 and 41.5 Mc/s so that the sound section of the receiver can also be checked. A circuit diagram is given, and the procedure for testing receivers and sound explained.

621.397.5 : 621.317.2 268
TV [television] **Test Equipment.**—P. H. Hunter. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 49–51 .. 108.) There is a particularly urgent demand for a ". . . synthetic video pattern generator capable of producing various types of test patterns on television receiver screens for the evaluation of their over-all performance". Design trends are discussed, and existing equipment reviewed.

621.397.5 : 621.396.677 269
Rhombic Antennas for Television.—Minter. (*See* 26.)

621.397.5(44) 270
Television in France.—(*J. Televis. Soc.*, March 1946, Vol. 4, No. 9, pp. 224–225.) An abstract of a report by the Combined Intelligence Objectives Sub-Committee, which describes visits to the Compagnie des Compteurs, Montrouge, and to the studios of the R.D.F. At the former a 400-line projection on a screen 6 ft by 4 ft was seen, the quality being comparable with that from Alexandra Palace. Iconoscopes were employed for all cameras:

owing to the shortage of mica the mosaic was deposited on oxidized aluminium sheet. Demonstrations of both a 1050- and a 450-line system were seen, the increase in entertainment value with the 1050-line system being most marked. Electrostatic lenses were used in all the iconoscopes, but magnetic lenses were used in the projection tubes.

The transmitter in the Eiffel Tower belonging to 'R.D.F.' was damaged by the Germans before they left, and will probably not be in operation for two years.—'R.D.F.' has a large television studio built to the order of the Germans and in which all the equipment is German and made by Fernseh A.G. Demonstrations of film transmission with a 441-line interlaced system gave very good definition and a quality comparable with film transmission from Alexandra Palace. Three additional studios were under construction. See also 2741 and 1105 of 1946.

621.397.621 **271**

Line Scanning Systems for Television.—A. M. Spooner & E. E. Shelton. (*Electronic Engng*, Oct. 1946, Vol. 18, No. 224, pp. 302–307.) Formulae are derived for the time-base wattage of electrostatic and electromagnetic deflecting systems. A 'figure of merit' is obtained for deflecting coils, and its experimental measurement considered. The importance of each variable involved, such as illumination, anode voltage, beam current, and tube shape, is considered, and the relative merits of various scanning coils deduced.

621.397.621 : 621.397.645 **272**

Electromagnetic Frame Scanning.—W. T. Cocking. (*Wireless World*, Sept. 1946, Vol. 52, No. 9, pp. 289–291.) A linear frame-scan can be obtained economically only by compensating for the non-linearity of the coupling to the scan coils by the valve curvature. See also 3080 of 1946 (Cocking).

621.397.645 **273**

Video Amplifier H.F. Response : Parts 1 & 2.—*See* 61 & 62.)

621.397.823 **274**

The Noise Suppressor in the V114 [television set]. —H. A. Fairhurst. (*Murphy News*, Oct. 1946, Vol. 21, No. 10, pp. 244–246.) Suppression for noise pulses shorter than the periodic time of the highest audio frequency is obtained by a series diode in which a backing potential is derived from the audio signal through a circuit of time constant less than that corresponding to the maximum audio frequency but greater than the noise pulse duration.

TRANSMISSION

621.396.13 : 621.394.65 **275**

Frequency Shift Keying Techniques.—C. Buff. (*Radio, N.Y.*, Aug. 1946, Vol. 30, No. 8, pp. 14 . . 55.) The change in bias of a reactance tube, connected across the tuned circuit of a 200 kc/s oscillator, produces the desired frequency shift. The gain in signal-to-noise ratio is experimentally estimated as 11 to 20 db better than for on-off keying. Circuit and design data are included. See also 2306 of 1946 (Peterson *et al.*).

621.396.61 **276**

A Medium-Power Bandswitching Transmitter.—J. M. Smith. (*QST*, Oct. 1946, Vol. 30, No. 10, pp. 13–21 . . 108.) A crystal oscillator, arranged for five alternative crystals, is connected directly to a type 807 amplifier on the 80- and 40-metre amateur bands or through a type 6N7 frequency multiplier into the 807 amplifier on the 20- and 10-metre bands. The final amplifier uses a type 4-125A beam tetrode with an input of 375 W for c.w. or 270 W for telephonic operation.

621.396.619.018.41 : 621.385.5 **277**

Phasitron F.M. Transmitter.—F. M. Bailey & H. P. Thomas. (*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 108–112.) Describes in detail the mode of operation of the phasitron (see also 1405 and 2767 of 1946) and its application to a 250-W transmitter covering the frequency range 88–108 Mc/s. The phasitron output frequency is multiplied by 432 to give carrier frequency. Inductive tuning is employed in the output tank circuit of the transmitter, the entire frequency range being covered without changing coils or taps.

VALVES AND THERMIONICS

519.2 **278**

The Resultant of a Large Number of Events of Random Phase.—C. Domb. (*Proc. Camb. phil. Soc.*, Oct. 1946, Vol. 42, Part 3, pp. 245–249.) "Rayleigh's method of deducing the probability distribution of the amplitude of the sum of n equal vibrations of random phase is generalized to the case when the amplitude of each vibration is a definite function of its phase. The same method is applied to the shot effect and it enables the distribution of random noise to be obtained. Campbell's theorem and its generalizations can then be deduced from this."

537.291 **279**

Influence of Space Charge on the Bunching of Electron Beams.—Brillouin. (*See* 75.)

537.533.8 : 621.385.1.032.216 **280**

Dissociation Energies of Surface Films of Various Oxides as determined by Emission Measurements of Oxide Coated Cathodes.—H. Jacobs. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, pp. 692–693.) Summary of Amer. Phys. Soc. paper.

537.533.8 : 621.385.1.032.216 **281**

Enhanced Thermionic Emission from Oxide Cathodes.—J. B. Johnson. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 702.) After bombardment by a short pulse of electrons, the thermionic activity of a cathode in the temperature range for emission can remain abnormally high for many microseconds. Summary of Amer. Phys. Soc. paper.

537.533.8 : 621.385.1.032.216 **282**

Secondary Emission of Thermionic Oxide Cathodes. —J. B. Johnson. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 693.) The number, δ, of secondary electrons emitted per primary, rises with primary energy from 1 at about 30 eV to a maximum of 4–10 at 1 200–1 500 eV. When cold the oxide has a higher δ but acquires a surface charge which limits the escape of secondaries. δ decreases with increased temperature and the reported exponential increase with temperature (see 1925 of 1939, Morgulis & Nagorsky) is attributed to a temporary increase in thermionic activity. Summary of Amer. Phys. Soc. paper.

537.533.8 : 621.385.1.032.216 283

The Poisoning of Oxide Cathodes by Gold.—J. Rothstein. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 693.) The experimental technique is described. It is concluded that "Au readily migrates (diffuses) over (BaSrCa)O and Ni, that Au inhibits emission, and that suitable thermal gradients can alter Au concentration and restore emission. It seems likely that Au exerts its maximum inhibiting effect when present at the outer surface of the oxide." Summary of Amer. Phys. Soc. paper.

537.533.8 : 621.385.1.032.216 284

Dissociation Energies of Surface Films of Various Oxides as determined by Emission Measurements of Oxide Coated Cathodes.—H. Jacobs. (*J. appl. Phys.*, July 1946, Vol. 17, No. 7, pp. 596–603.) When an electron achieves a critical kinetic energy in moving from cathode to an oxide-coated anode, a dissociation of the oxide results. Liberated oxygen returning to the cathode reduces emission. This critical energy was found to be equivalent to the heats of formation of the oxides bombarded.

621.314.632 285

Small Deviations from Diode Behavior in Crystal Rectification.—K. F. Herzfeld. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 683.) Assuming that the mean free path of the electrons in the blocking layer is large, but not infinite, compared with the thickness of the layer, it is found that the dependence of current on voltage is slightly less than in the pure diode theory. Summary of Amer. Phys. Soc. paper.

621.326[.1 + .3/.4] : 621.392.5 : 621.316.722.078.3 286

The Characteristics of Lamps as applied to the Non-Linear Bridge, used as the Indicator in Voltage Stabilizers.—Patchett. (*See* 45.)

621.384.5.08 287

The Properties of Glow Tubes and Their Applications for Measurement Purposes.—Glaser. (*See* 191.)

621.38.216 288

A High Power Rising Sun Magnetron.—A. Ashkin. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 701.) A mode separation of a rising sun magnetron is independent of anode length. Tubes with long anodes have been constructed to give a peak power of 1 MW for λ 3 cm with output efficiency of about 45%. Summary of Amer. Phys. Soc. paper.

621.385.16 289

"Crown of Thorns" Tuning of Magnetrons.—S. Sonkin. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 701.) Description of a mechanically tunable vane-type magnetron for which linear tuning up to 10% was obtained. The causes and elimination of variations in power output with wavelength are discussed. Summary of Amer. Phys. Soc. paper.

621.385.16 290

Space-Charge Frequency Dependence of a Magnetron Cavity.—M. Phillips & W. E. Lamb, Jr. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 701.) A correction to the resonant frequency due to a thin layer of space charge surrounding the cathode is given "by a resonance type formula about the cyclotron frequency $eB/2\pi m$". A small amplitude theory is used based on a single stream steady state, and is limited to low level oscillations. Summary of Amer. Phys. Soc. paper.

621.385.16 291

Energy Build-Up in Magnetrons.—L. P. Hunter. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 700.) Analysis of the law of build-up and its dependence on load, initial noise level and cavity *Q*. Summary of Amer. Phys. Soc. paper.

621.385.16 292

One Centimeter Rising Sun Magnetrons with 26 and 38 Cavities.—A. V. Hollenberg, S. Millman & N. Kroll. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 701.) Summary of Amer. Phys. Soc. paper.

621.385.16 : 621.396.615.14.029.63/.64 293

The Magnetron as a Generator of Centimeter Waves : Parts 1 & 2.—J. B. Fisk, H. D. Hagstrum & P. L. Hartman. (*Bell Syst. tech. J.*, April 1946, Vol. 25, No. 2, pp. 167–263 & 264–348.) In Part 1 the fundamentals of the magnetron are discussed. A general picture is given of the nature of the electronic mechanism, and of the role played by the r.f. circuit and load. The second part gives a description of tests on a British 10-cm, 10-kW, 8-resonator magnetron which led to the development of radar magnetrons for λ 20 to 45 cm with peak power up to 700 kW ; the introduction of strapping improved efficiency greatly at the higher powers. Tunable magnetrons were first investigated at this wavelength, but later also for λ 3 cm. Variations of the British design resulted in a series of magnetrons with powers up to 1 MW for λ 10 cm. Then came magnetrons with λ 3 cm for airborne and marine applications, while finally for λ 1 cm the 3J21 'rising sun' magnetron was developed.

Details of design, production problems, and general performance of all these magnetrons are given, together with a short account of work on cathode design.

621.385.16.029.63 294

New Magnetron Designs for Continuous Operation in the Decimeter Wave Range.—D. A. Wilbur. (*Phys. Rev.*, 1st/15th July 1946, Vol. 70, Nos. 1/2, p. 118.) Two devices to eliminate cathode back heating from high-frequency operation of the split-anode magnetron are described, in which multi-gap operation is attained together with an easily tunable external tank circuit. (1) Only two segments of the multi-gap are used, the remainder being replaced by a single electrode maintained at zero r.f. potential. (2) A multiplicity of electrodes is used mounted internally on a conducting helix.

Abstract of an Amer. Phys. Soc. paper.

621.385.16.032.21.029.63 295

A 10-Kilowatt Magnetron with Water-Cooled Cathode.—R. V. Langmuir & R. B. Nelson. (*Phys. Rev.*, 1st/15th July 1946, Vol. 70, Nos. 1/2, p. 118.) A simple split-anode magnetron delivering over 10 kW c.w. at 60% efficiency and tuning from 560 to 625 Mc/s was developed for radar jamming. The efficiency is comparable with that of multiple-anode types. The cathode operates by secondary emission, and is water-cooled ; the best secondary-emission surface found was made of a magnesium alloy.

Abstract of an Amer. Phys. Soc. paper.

621.385.16.032.22 **296**

Development of the Rising Sun Magnetron Anode Structure.—S. Millman & A. Nordsieck. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 701.) With very short waves the mode separation of magnetrons by strapping becomes difficult and expensive. An alternative method is to make the resonator cavities alternately large and small with resonant wavelength ratio between 1.5 and 2.5. The mode spectrum, and the advantages and limitations of these magnetrons are discussed. Summary of Amer. Phys. Soc. paper.

621.385.16.032.22 **297**

Theory of the Rising Sun Magnetron Anode.—N. Kroll & W. E. Lamb, Jr. (*Phys. Rev.*, 1st/15th June 1946, Vol. 69, Nos. 11/12, p. 701.) The method is analogous to that of Clogston for a symmetric anode. Maxwell's equations can be solved for both the cathode-anode space and the side resonators for a specified boundary tangential electric field. Resonance is determined by the continuity of the average magnetic field at the junctions. The method may be extended to other structures. Summary of Amer. Phys. Soc. paper.

621.385.18.029.64 : 537.5 **298**

Conductivity of Electrons in a Gas at Microwave Frequencies.—Margenau. (*See* 76.)

621.385.831.012.8 **299**

Equivalent Noise Representation of Multi-Grid Amplifier Tubes.—Twiss & Schremp. (*See* 37.)

621.385.832 + 621.397.331.2 **300**

Simplification of Cathode Ray Tube Design by the Application of the Theory of Similitude.—I. Moss. (*J. Televis. Soc.*, March 1946, Vol. 4, No. 9, pp. 206–219. Discussion, p. 228.) A discussion of "the application of the general theories of scale, dimensional homogeneity, and energy conservation to cathode-ray tube designing. From these simple bases it is shown that many important deductions can be drawn about the general form which the tube geometry should assume."

The following four rules form the basis of the 'relaxation' methods of cathode-ray-tube design. The first three are quite rigorous within their limiting postulates. The fourth has some theoretical justification, but the main support is experimental.

(1) Principle of voltage similitude: in any electron optical system, in which space charge is negligible, and in which the electrons start from rest, the electron trajectory is unaltered by multiplication of all electrode potentials by a constant factor k. The transit time between any two fixed points in the system varies as $1/\sqrt{k}$.

(2) Principle of geometrical similitude: in any electron optical system in which the total current flow is constant, the shape of the field and of the electron trajectory is unaltered by multiplication of the size of all the bounding electrodes by a constant factor k. The transit time between corresponding points in the two systems is proportional to k.

(3) Spot size/crossover size relationship: if the crossover and spot are formed in regions of the same potential, then spot size = crossover size × geometrical magnification (M). More generally, if V_1 is the crossover potential, and V_2 the spot potential, then spot size = crossover size × M × $\sqrt{V_1}/\sqrt{V_2}$.

(4) Dependence of crossover size on voltage on crossover forming electrode: to a close approximation, in any system where the space charge is negligible, the crossover diameter is inversely proportional to the square root of the potential on the crossover-forming electrode.

Proofs of these principles are given, and they are applied to some specific problems including the design of projection tubes.

621.385.832 : 535.371.07 **301**

Long Persistence C.R. Tube Screens.—Feldt. (*See* 123.)

621.385.832 : 621.396.9 **302**

Skiatron.—(*Electronics*, Oct. 1946, Vol. 19, No. 10, pp. 216..220.) A short account of a lecture by P. G. R. King at the I.E.E. Radiolocation Convention, already noted in 2404 of 1946.

621.385.832.032.2 **303**

Magnetic Focusing and Deflection.—D. Rawcliffe & R. W. Dressel. (*Electronic Industr.*, Oct. 1946, Vol. 5, No. 10, pp. 52–56..111.) Magnets have the advantages of light weight and stability with respect to temperature, but generally give a focus inferior to that obtained with coils. Air-core and iron-core deflection coils are designed to overcome the distortion in a display pattern due to the curvature of the tube screen, non-uniformity of the magnetic field and inductive and capacitive coupling between the various points of the coils.

621.385.832.032.2 **304**

Comparison of Electrostatic and Electromagnetic Deflection in Cathode-Ray Tubes.—(*J. Instn elect. Engrs*, Part III, Sept. 1946, Vol. 93, No. 25, p. 364.) Summary of I.E.E. Radio Section discussion led by E. W. Bull and V. A. Stanley. It was pointed out that defocusing with beam deflexion was greater with electrostatic deflexion because the electron beam underwent energy changes in the deflecting field. Electrostatic tubes suffered from trapezium distortion. The effects of gaseous ions were less with electrostatic deflexion. Less energy was required to scan an electrostatic tube.

The use of mixed magnetic and electrostatic deflection, the difficulties of aligning deflexion and focusing coils when replacing magnetic tubes and the question of deflexion methods for oscillograph purposes were also discussed. For another account see *J. Televis. Soc.*, June 1946, Vol. 4, No. 10, p. 264.

621.396.622 : 621.396.619.018.41 **305**

Single-Stage F. M. Detector.—Bradley. (*See* 227.)

621.385 (02) **306**

Inside the Vacuum Tube. [Book Review]—J. F. Rider. J. F. Rider Publisher Inc., New York, $4.50. (*Proc. Instn Radio Engrs, Aust.*, Sept. 1946, Vol. 7, No. 9, p. 37.) "The aim in his [the author's] numerous well-known works has been to cover one subject at a time, dealing mainly with fundamentals and starting from ground level."

MISCELLANEOUS

001.891 **307**

Research Problems of the Smaller Firm.—(*Electrician*, 25th Oct. 1946, Vol. 137, No. 3569, p. 1140.) Brief summary of address by Sir Edward Appleton.

002 : 001.81 **308**
The Preparation of Technical Papers.—W. E. Clegg. (*J. Instn Engrs Aust.*, June 1946, Vol. 18, No. 6, pp. 135–139.) Paper presented before the Juniors' and Students' Section of Newcastle division. A bibliography of 11 items is given.

061.6 **309**
The National Physical Laboratory at Teddington.—(*Nature, Lond.*, 14th Sept. 1946, Vol. 158, No. 4011, pp. 361–363.) A review of its work as seen on the first post-war open day.

614.825 **310**
Dangerous Electric Currents.—C. F. Dalziel. (*Trans. Amer. Inst. elect. Engrs*, Aug./Sept. 1946, Vol. 65, Nos. 8/9, pp. 579–584.) A general account of the physiological effects, and the factors determining their magnitude, of the passage of electric currents through the body. Illustrative results of particular tests applied to certain animals and to man are included.

620.197 (213) : 621.396.6.045 **311**
Tropicalizing [transformers and chokes].—Scarff. (*See* 133.)

620.197 (213) : 621.314.045 **312**
Impregnated Windings [for "tropicalizing" transformers].—Williams : Burkett. (*See* 134.)

621.3(091) : 51 **313**
Notable Electrotechnical Forecasts.—C. Grover. (*Distrib. Elect.*, Oct. 1946, Vol. 19, No. 164, pp. 180–182.) An historical review of the contribution of mathematics to the development of electrical theory and forecasting of phenomena.

621.39 : 371.3] + 621.315.6 **314**
I.E.E. Radio Section Address : Part 1 — Training Courses ; Part 2 — Dielectric Developments.—W. Jackson. (*Elect. Rev., Lond.*, 18th Oct. 1946, Vol. 139, No. 3595, p. 609 ; *Electrician*, 18th Oct. 1946, Vol. 137, No. 3568, pp. 1065–1066.) Summaries of Chairman's inaugural address.

621.317 **315**
The History and Development of the British Scientific Instrument Industry.—S. L. Barron. (*Beama J.*, Sept. 1946, Vol. 53, No. 111, pp. 325–328.) Abridged version of a lecture delivered at the Stockholm Exhibition of British Scientific Instruments, 1946.

621.317.785 **316**
Trends in Measurements.—L. J. Matthews. (*Elect. Times*, 31st Oct. 1946, Vol. 110, No. 2871, p. 603.) Summary of inaugural address of the Chairman to the I.E.E. Measurements Section dealing particularly with recent developments in consumers' a.c. meters.

621.327.43 : 628.971.6 **317**
Dimming Fluorescent Lighting.—H. A. Miller. (*Elect. Rev., Lond.*, 20th Sept. 1946, Vol. 139, No. 3591, pp. 457–458.) The possible circuits discussed are (*a*) series resistance, (*b*) variable-voltage auto-transformer, (*c*) saturable reactor control, (*d*) thyratron control. Diagrams are given for each.

621.38.001.8 **318**
Electronics — Servant or Fad ?—P. G. Weiller. (*Instruments*, Jan. 1946, Vol. 19, No. 1, pp. 2–8.) Verbatim account of a lecture in which the advantages and disadvantages of electronic control and measuring devices for industrial applications are discussed. It is stressed that very careful prior consideration is necessary to decide whether such a device is likely to provide a practicable and economic solution to the problem in hand. Several instructive examples are described.

621.396 : 371.3 **319**
Aids to [radio] **Training.**—M. G. Scroggie. (*Wireless World*, Sept. 1946, Vol. 52, No. 9, pp. 303–304.) A review of an exhibition of equipment used in the teaching and training of R.A.F. personnel in the operation and maintenance of radio and radar apparatus.

654.19(41) **320**
Broadcasting in Great Britain.—(*Nature, Lond.*, 31st Aug. 1946, Vol. 158, No. 4009, pp. 314–315.) Abstract of the White Paper on Broadcasting policy which justifies the renewal of the B.B.C.'s Royal Charter for 5 years from 1st January, 1947.

530.145(02) **321**
Philosophic Foundations of Quantum Mechanics. [Book Review]—H. Reichenbach. University of California Press, Berkeley & Los Angeles: Cambridge University Press, London, 1944, 182 pp., $3. (*Nature, Lond.*, 14th Sept. 1946, Vol. 158, No. 4011, pp. 356–357.)

621.3(031) **322**
Whittaker's Electrical Engineer's Pocket Book. [Book Review] — R. E. Neale (Ed.). Pitman, London, 7th edn 1946, 938 pp., 30s. (*Electrician*, 12th July 1946, Vol. 137, No. 3554, p. 100 ; *Wireless World*, Oct. 1946, Vol. 52, No. 10, p. 332.) New edition, almost completely rewritten to bring it into line with modern practice.

621.394/.395](021) **323**
A Handbook of Telecommunication (Telephony & Telegraphy over Wires). [Book Review] —B. Cohen. Pitman, London, 437 pp., 30s. (*Elect. Rev., Lond.*, 25th Oct. 1946, Vol. 139, No. 3596, p. 652.)

621.396(075) **324**
An Experimental Course in the Fundamental Principles of Radio. [Book Review] —R. H. Humphrey. Pitman, London, 194 pp., 12s. 6d. (*Electronic Engng*, Sept. 1946, Vol. 18, No. 223, p. 292.)

621.396(075) **325**
Radio Communications. [Book Review] —W. T. Perkins & R. W. Barton. George Newnes, London, 312 pp., 12s. 6d. (*Elect. Rev., Lond.*, 23rd Aug. 1946, Vol. 139, No. 3587, p. 296 ; *Wireless World*, Sept. 1946, Vol. 52, No. 9, p. 307.) ". . . primarily a manual of the 'question and answer' type, intended for students. . . ."

621.396.3(075) **326**
Handbook of Technical Instruction for Wireless Telegraphists. [Book Review] —H. M. Dowsett & L. E. Q. Walker. Iliffe & Sons, London, 8th edn, 660 pp., 30s. (*Electronic Engng*, Sept. 1946, Vol. 18, No. 223, pp. 292–293.) ". . . It is an essential requirement for passing the P.M.G. Certificate and can be confidently recommended to all prospective wireless telegraphists."