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The Inventor of Radio Telegraphy

IN the January Editorial we discussed the efforts being made in Russia to boost the claim that Popov anticipated Marconi in the invention of radio. We pointed out that no support for this claim could be found in the Russian publications of 1896–1897 nor in the Popov memorial number of *Electritchestvo*, published in 1925. Our attention has been drawn, however, to the fact that in a recent Russian publication, "The Invention of Radio by Popov," published in 1945, three letters are reproduced, written in December 1925 and January 1926 by three Russian Professors who were present at the demonstration given by Popov on 12th March, 1896.* These letters, together with a booklet by Rybkin published in 1945, maintain that Popov was prevented from publishing the results of his experiments by the Russian Naval Authorities. In May 1895 Popov demonstrated the reception of waves from a Hertz vibrator by means of a coherer with a relay and automatic tapper. In March 1896 he is alleged to have given another demonstration; this time the transmitter was in another part of the University about 250 metres away and provided with a sending key. At the receiving apparatus the slowly rotating drum used for recording lightning flashes had been replaced in September 1895 by a Morse inker. After an introductory explanation the experiment began and as the letters appeared on the moving tape they were written on the blackboard, where the audience soon saw the words "Heinrich Hertz." It is reported by P. N. Rybkin, who was Popov's assistant and who was at the transmitter, that a naval captain warned Popov before the demonstration that he should exercise great caution as his work was of military importance and that Popov consequently insisted on the experiment being described in the minutes merely as a demonstration of Hertzian waves. Lebedinsky says in his letter that he preserved the tape with the words "Heinrich Hertz" until 1918–1919 when it perished with the rest of his library at Riga.

It must be remembered that Popov was attached to the Torpedo School and was under the control of the Naval Authorities. It is difficult to understand, however, why a detailed description of this important demonstration was not given in the 1925 memorial number of *Electritchestvo* to which Lebedinsky contributed an article without any mention of the transmission of "Heinrich Hertz" and its recording on a tape. In his article he did say, however, that Popov improved on the apparatus employed by Lodge by making the action of the tapper more regular and by using the more technical Morse apparatus in place of the galvanometer used by Lodge.

Popov's Letter to Ducrétet

This, however, is not the only difficulty in connection with this demonstration. When one reads the following letter written by Popov in 1897 to Ducrétet, who wished, if possible, to establish Popov's priority, one cannot but wonder if the memories of these old colleagues after thirty years may have played them false with regard to the date of the "Heinrich Hertz" demonstration. The letter was as follows† :—"I have no

* We are indebted to V. L. Rastorgoueff, B.Sc., of Standard Telephones and Cables, Ltd., for kindly supplying translations from the Russian.

† Translated from the French. *Electritchestvo*, Popov Memorial Number, 1925, p. 13.

written record which could fix my participation in the practical solution of the problem of wireless telegraphy, other than that which you already know. Nevertheless I regard that article as sufficient to prove the conformity of the integral parts of my apparatus and their disposition with the apparatus of the receiving station of Marconi. The receiving conductor, the connection of the tube to the earth, the disturbing effect of the spark, the value of the self-induction are all explained. In January 1896 I demonstrated my apparatus to the Cronstadt section of the Russian Technical Society; I said that it was desirable to test my apparatus over great distances and I demonstrated the action of the apparatus under the following conditions: The Hertz vibrator was installed at one end of one of the rooms and the acoustic apparatus was mounted on a portable stand. To ensure resonance the terminals of the tube [coherer] were connected to zinc plates of the same size as those fitted to the vibrator. The apparatus, which was carried across all the lecture rooms and which was finally removed to the most distant part of the building, responded all the time to the signals of the bell (sonnette). In March I showed an apparatus for optical experiments with electromagnetic rays; reflection, refraction, effect of a grating and rotation of the plane of polarization.

[N.B.—It was in March that the "Heinrich Hertz" transmission is alleged to have taken place.] In September 1896 the first notices of Marconi's experiments appeared in the daily press, but the nature of the apparatus was not divulged and the technical journals were full of conjectures about this new discovery. I then published a letter in a local journal, in which, after recalling to the readers the nature of my apparatus, I showed that in the registration of storms there were some records of discharges as far away as 30 kilometres, that the recording of artificial discharges up to a mile was possible and that in all probability Marconi's apparatus was similar to mine. This letter was published in the journal *Kotline* in October 1896."

It seems almost incredible that Popov could have written this letter to a leading French scientist anxious to obtain material to establish Popov's priority as the inventor of wireless telegraphy, giving him details of experiments demonstrated in March on reflection and refraction, without any reference whatever to the demonstration of actual wireless telegraphy, also alleged to have been made in March. If the demonstration did take place as alleged, then Popov was either so impressed by the need for secrecy that he was afraid even to refer to the demonstration, or he was so concentrated on the coherer and its auxiliaries that he regarded the actual transmission and reception of Morse signals and their recording on a telegraphic apparatus as minor details not worthy of mention.

Another difficulty arises from the statement made by his colleague Professor Guéorguievsky that "during the winter 1895–96 Popov continued to improve his storm indicator, tried his apparatus at the meteorological station of the Forestier Institute, and at the same time experimented with Röntgen rays which had recently been discovered. The idea of continuing his researches on the practical applications of his apparatus and the study of better conditions of operation of this new receiver were for the moment laid aside." Yet this was the winter during which we are told that he made such progress that the Naval Authorities warned him to "keep it dark." On p. 7 of the memorial number of *Electritchestvo* we are told that in 1896 he made his first trials of radio-telegraphy in the Russian Navy; on p. 12 we are told that he spent the summer at Nijny-Novgorod where in September the news of the Marconi invention shook him brusquely out of his torpor. In view of all this, it is not surprising that, in a letter to us, a well-known American radio engineer says "I think there was some date juggling to get him in the front rank."

Marconi Comes to London

It was in February 1896, that is, a month before the alleged demonstration, that Marconi, having brought his experiments at Pontecchio to a successful conclusion, came to London in order to apply for a patent, which he did in June. In the specification he does not go into details of the ordinary telegraphic apparatus but mentions the Morse key at the transmitter and "at the receiving instrument, a local battery circuit containing an ordinary receiving telegraphic or signalling instrument." To him the Morse inker and its tape were normal adjuncts, and it may be that Popov adopted much the same attitude. Assuming that Popov did make the experiment as alleged and transmitted the two words from one part of the University to another in March, after which he put his apparatus aside and spent the summer "in a torpor" at Nijny-Novgorod, one cannot possibly give him priority over Marconi who had completed his experiments in Italy and had come to London in February to take out a patent and to interest the British Post Office in his invention.

One may, however, very reasonably ask what one means by "inventing" wireless telegraphy and what it was actually that either Marconi or Popov invented. In 1879 Hughes had demonstrated the transmission of signals up to several hundred yards; as someone wrote in 1899 "Hughes's experiments of 1879 were virtually a discovery of Hertzian waves before Hertz, of the coherer before Branly, and of wireless telegraphy before Marconi and others." In a letter to Fahie in April 1899 Hughes said "Hertz's experiments were far more conclusive than mine, although he used a much less effective receiver than the microphone or coherer. I then felt it was now too late to bring forward my previous experiments . . . and I have been forced to see others make the discoveries I had previously made as to the sensitiveness of the microphone contact and its useful employment as a receiver for electric aerial waves." He then wrote in the highest terms of the achievements of Marconi.

In June 1894 at the Royal Institution Lodge demonstrated the transmission of signals over a distance of 150 yards, using a form of Branly detector. He himself said later (1897) "stupidly enough no attempt was made to apply any but the feeblest power so as to test how far the disturbance could really be detected."

We also know that "in Dec. 1895 Captain H. B. Jackson [who was then Commander of the torpedo school ship Defiance] commenced working in the same direction, and succeeded in getting Morse signals through space before he heard of Marconi. His experiments, however, were treated as confidential at the time and have not been published." There was no question of priority, for the experiments were only commenced about two months before Marconi's arrival in London to patent his invention. In his report for 1896 Captain Jackson said that "the principles on which Signor Marconi's apparatus was constructed were similar to those employed by the Defiance, but more fully developed, and the instruments themselves were much more sensitive."

Practically every element of the transmitting and receiving apparatus had been previously described. The one that gave most trouble to Marconi and to Popov was the coherer, and the greater part of Marconi's patent specification was devoted to the various devices introduced in order to stabilize its operation.

The Edison elevated aerials, the Hertz oscillator, the Branly coherer and the Morse inker, were all available, although the first-named was probably unknown to either Marconi or Popov, but their combination into a practical reliable system of wireless telegraphy capable of operation over a considerable distance was the goal at which Marconi aimed, and its accomplishment was his great achievement.

The first claim of Marconi's first patent is as follows: "The method of transmitting signals by means of electrical impulses to a receiver having a sensitive tube or other sensitive form of imperfect contact capable of being restored with certainty and regularity to its normal condition substantially as described." This indicates clearly what Marconi regarded as the *pièce de résistance* of his achievement, as does the fact that, of the nineteen claims, the first thirteen are concerned entirely with the coherer and its adjuncts.

During 1895 Marconi in Italy and Popov in Russia were each experimenting with the coherer and its tapper, Marconi with the sole object of using it as a telegraphic receiver, Popov with the primary object of perfecting his lightning recorder, but with its possible application to telegraphy also in his mind. The one, a young man of means, fired with enthusiasm and devoting himself entirely to the one object, the other, busy with lectures and laboratory work at the Naval College, and having to spend the summers as a technician at the Nijny-Novgorod exhibition where he was in charge of the electric lighting plant. It was here in September 1896 that, as his colleague said, he was brusquely shaken out of his torpor by the news of the achievement of the young Italian, which had been kept secret until the Patent Application was lodged in June.

G. W. O. H.

MAGIC-TEE WAVEGUIDE JUNCTION

By Godfrey Saxon, M.Sc. and C. W. Miller, M.Sc.

1. Introduction

IN this paper the problem of a junction of four rectangular waveguide legs in the form known variously as the hybrid-tee, magic-tee or side-outlet tee is investigated in detail both mathematically and experimentally. The general form taken by the junction is shown in Fig. 1 (a). This system can be represented by four transmission lines of unit characteristic impedance coupled together by means of an ideal four-element transformer consisting of two pairs of coils at right angles, as shown in Fig. 1 (b).

The authors have been greatly assisted in their problem by the work on T-junctions in rectangular waveguides carried out by Allanson, Cooper and Cowling and recently published.[1] In their work they define "characteristic points" at which the coupling reactances are considered to be situated, the position of these points depending on the wavelength of the propagated wave and on the waveguide dimensions. For any given three-way junction these points are fixed; but in a four-way junction there is another degree of freedom so that a characteristic point may be chosen *arbitrarily* in any one leg and the three other points determined in relation to it. Referring to Fig. 1 (a), the characteristic points are chosen to have the significance that if short-circuiting pistons are placed at these points in legs 1 and 2 then radiation along either leg 3 or leg 4 will be totally reflected at the junction, and the electric field will vanish at the characteristic point of the feeding leg (and, of course, at half-wavelengths away from this point). A similar property holds if the short-circuits are in legs 3 and 4 and radiation is fed down either leg 1 or leg 2.

The basic properties of the magic-tee, which appear to have been first described by Tyrell[2] of the Bell Telephone Laboratories, may be expressed as follows:

(a) If a generator is placed in leg 3 (the E-leg) and legs 1 and 2 are terminated equally and symmetrically with respect to the junction, then no power enters leg 4. The waves in legs 1 and 2 will be oppositely phased at equal distances from the junction.

(b) If a generator is placed in leg 4 (the H-leg) and legs 1 and 2 are terminated equally and symmetrically with respect to the junction, then no power enters leg 3. The waves in legs 1 and 2 will be in the same phase at equal distances from the junction.

(a) and (b) hold good for all normal operating frequencies in a junction of the type shown in Fig. 1 (a). They are properties to be expected from a consideration of the orthogonality of the field vectors in legs 3 and 4. The simple junction may be modified, however, so that for a particular frequency the magic-tee possesses further properties:

(c) If a generator is placed in leg 1, legs 3 and 4 being equally terminated with respect to characteristic points to be defined later, then the transmitted power divides

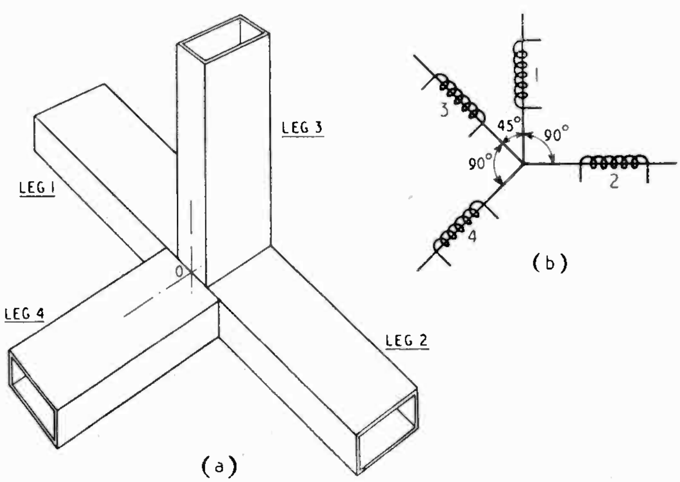

Fig. 1. *The general form of a four-leg waveguide junction is shown at (a) and its transformer equivalent at (b).*

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equally between legs 3 and 4, none entering leg 2. By symmetry a similar property holds if the source of power is in leg 2.

(d) If the pair of legs into which power flows when fed from a third leg are matched then unity standing-wave ratio will exist in this third leg.

In this paper a simple mathematical relationship between the admittances at the characteristic points is derived, embodying the above properties. An examination of this expression leads to several useful subsidiary properties.

The paper also gives details of the performance of two experimental tees, the first a simple tee made by balancing-out the mismatch in the side legs by inductive diaphragms, the second a more carefully designed magic-tee having greater bandwidth. Finally, some applications of the magic-tee are considered.

2. Mathematical Treatment

Allanson, Cooper and Cowling have shown that any lossless junction of waveguide legs can, for purposes of analysis, be replaced by a system of transmission lines of unit characteristic impedance, each line having a reactance at its input end which couples inductively with the other reactances. The input voltage of the rth line in terms of its own input current and that of each of the other lines may be written as:

$$V_{r0} = j\sum_{s=1}^{s=n} \alpha_{rs} I_{s0} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

where to allow for different sections of guide the voltage V is here the normalized voltage, averaged over the cross-section of the guide. The terms $j\alpha_{rs}$ are the mutual reactances, except where $r = s$, in which case they represent the self-reactances. It can also be shown that $\alpha_{rs} = \alpha_{sr}$.

Turning now to the case of the four-way junction, from Property (a), if power is fed down leg 3, legs 1 and 2 being equally terminated, $V_{40} = 0 = I_{40}$. The waves in 1 and 2 are in opposite phase so $V_{10} = -V_{20}$ and $I_{10} = -I_{20}$. Inserting these conditions in (1) we get:—

$$V_{10} = j[(\alpha_{11} - \alpha_{12})I_{10} + \alpha_{13}I_{30}]$$
$$V_{20} = j[(\alpha_{12} - \alpha_{22})I_{10} + \alpha_{23}I_{30}] = -V_{10}$$
$$V_{30} = j[(\alpha_{13} - \alpha_{23})I_{10} + \alpha_{33}I_{30}]$$
$$V_{40} = j[(\alpha_{14} - \alpha_{24})I_{10} + \alpha_{34}I_{30}] = 0$$

This is true for all values of I_{10} and I_{30}, hence

$$\alpha_{14} = \alpha_{24},\ \alpha_{34} = 0,\ \alpha_{11} = \alpha_{22},\ \alpha_{13} = -\alpha_{23} \quad (2)$$

The same result follows if we consider Property (b). The fact that $\alpha_{34} = 0$ means that there is no direct coupling between the E-leg and the H-leg. In the transformer to which the system is analogous the corresponding windings are at right angles.

Equations (1) may now be written:

$$\left.\begin{aligned} V_{10} &= j[\alpha_{11}I_{10} + \alpha_{12}I_{20} + \alpha_{13}I_{30} + \alpha_{14}I_{40}] \\ V_{20} &= j[\alpha_{12}I_{10} + \alpha_{11}I_{20} - \alpha_{13}I_{30} + \alpha_{14}I_{40}] \\ V_{30} &= j[\alpha_{13}(I_{10} - I_{20}) + \alpha_{33}I_{30}] \\ V_{40} &= j[\alpha_{14}(I_{10} + I_{20}) + \alpha_{44}I_{40}] \end{aligned}\right\} (3)$$

We have not yet decided what points in the waveguide legs are to represent the input points to the transmission lines (or the terminals of the perfect transformer). We shall therefore, first of all, pick arbitrarily on two symmetrical points P_1 and P_2 in legs 1 and 2 at which to measure V_{10} and V_{20}. If short-circuits are placed at these points so that $V_{10} = V_{20} = 0$, then radiation incident along leg 3 will be totally reflected, this being a particular case of Property (a).

Since $I_{10} + I_{20} = 0$

$$\left.\begin{aligned} 0 &= 2(\alpha_{11} - \alpha_{12})I_{10} + 2\alpha_{13}I_{30} \\ V_{30} &= j[2\alpha_{13}I_{10} + \alpha_{33}I_{30}] \end{aligned}\right\} (4)$$

We will choose a voltage node in leg 3 as the point P_3 at which to measure voltage. Since $V_{30} = 0$, equations (4) give:

$$2(\alpha_{13})^2 = \alpha_{33}(\alpha_{11} - \alpha_{12}) \quad \ldots \quad \ldots \quad (5)$$

By a similar process, a point P_4† in leg 4 may be found at which the voltage is zero when short-circuits are placed at P_1 and P_2, power being incident down this leg. The equations then yield the relationship:

$$2(\alpha_{14})^2 = \alpha_{44}(\alpha_{11} + \alpha_{12}) \quad \ldots \quad \ldots \quad (6)$$

It will be convenient now to write:

$$\alpha_{11} + \alpha_{12} = 2B,\ \alpha_{11} - \alpha_{12} = 2A,$$
$$\alpha_{13} = Ar,\ \alpha_{14} = Bs,\ \alpha_{33} = Ar^2,$$
$$\alpha_{44} = Bs^2.$$

The self and mutual reactances are now expressed in terms of four unknowns.

Equations (3) now become:

$$\left.\begin{aligned} V_{10} &= j[(A + B)I_{10} + (B - A)I_{20} \\ &\quad + ArI_{30} + BsI_{40}] \\ V_{20} &= j[(B - A)I_{10} + (A + B)I_{20} \\ &\quad - ArI_{30} + BsI_{40}] \\ V_{30} &= j[Ar(I_{10} - I_{20}) + Ar^2I_{30}] \\ V_{40} &= j[Bs(I_{10} + I_{20}) + Bs^2I_{40}] \end{aligned}\right\} (7)$$

† It should be appreciated that whilst P_1 and P_2 are symmetrically disposed about the junction by definition, it will be found in general that P_3 and P_4 are not at equal distances from the junction.

Setting $I_{r0} = Y_r V_{r0}$, where Y represents the normalized admittance looking towards the junction, we arrive at a set of four simultaneous equations in V_{r0} whose solution is given by the determinantal equation :

$$\begin{vmatrix} (A+B)Y_1 + j & (B-A)Y_2 & ArY_3 & BsY_4 \\ (B-A)Y_1 & (A+B)Y_2 + j & -ArY_3 & BsY_4 \\ ArY_1 & -ArY_2 & Ar^2Y_3 + j & 0 \\ BsY_1 & BsY_2 & 0 & Bs^2Y_4 + j \end{vmatrix} = 0$$

The expansion of this determinant can be much simplified by considering the determinant :

$$\begin{vmatrix} (A+B)Y_1 & (B-A)Y_2 & ArY_3 & BsY_4 \\ (B-A)Y_1 & (A+B)Y_2 & -ArY_3 & BsY_4 \\ ArY_1 & -ArY_2 & Ar^2Y_3 & 0 \\ BsY_1 & BsY_2 & 0 & Bs^2Y_4 \end{vmatrix}$$

which is the original one with the j-terms omitted and which simplifies to

$$Y_1Y_2Y_3Y_4r^2s^2 \begin{vmatrix} A+B & B-A & A & B \\ B-A & A+B & -A & B \\ A & -A & A & 0 \\ B & B & 0 & B \end{vmatrix}$$

Adding the last two columns we obtain an identity with column one. The determinant is therefore identically zero. Examination of the first minors shows that they too all vanish. These results simplify the expansion of our original determinant since they mean that third and fourth power terms in Y must all disappear. Neglecting these the determinantal equation reduces to :

$$4[1 - j\{(A+B)(Y_1+Y_2) + Ar^2Y_3 + Bs^2Y_4\} - (A+B)^2Y_1Y_2 - ABr^2s^2Y_3Y_4 - (A+B)(Y_1+Y_2)(Ar^2Y_3 + Bs^2Y_4) + (B-A)^2Y_1Y_2 + (Y_1+Y_2)(A^2r^2Y_3 + B^2s^2Y_4)] = 0$$

which on simplification gives :

$$1 - AB[4Y_1Y_2 + (Y_1+Y_2)(r^2Y_3 + s^2Y_4) + r^2s^2Y_3Y_4] - j[(A+B)(Y_1+Y_2) + Ar^2Y_3 + Bs^2Y_4] = 0 \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

This is the general relationship between admittances in the four legs as measured at points P_1, P_2, P_3 and P_4 which are related to each other in the manner described above. It should be noted that since P_1 and P_2 were chosen arbitrarily there is an infinite number of sets of points P related in this manner, each set of points giving different values for the constants A, B, r and s. If it is desired to determine appropriate values for these constants for any given size of waveguide and for a particular wavelength, a simple experimental technique may be employed. If we assume that by introducing short circuits into legs 1 and 2 we have located a set of points at which to measure impedance, and that legs 1 and 2 are correctly terminated so that ‡$Y_1 = Y_2 = -1$, then, inserting this condition in (8) we get :

$$1 - AB[4 - 2(r^2Y_3 + s^2Y_4) + r^2s^2Y_3Y_4] + j[2(A+B) - Ar^2Y_0 - Bs^2Y_4] = 0 \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

If power is fed down the H-leg, we know that no power enters the E-leg, so the coefficient of Y_4 in (9) vanishes.

That is : $AB[2s^2 - r^2s^2Y_3] - jBs = 0$

$$\text{or} \quad Y_3 = \frac{1}{r^2}\left(2 - \frac{j}{A}\right) \quad \ldots \quad \ldots \quad (10)$$

If the admittance is measured at the point P_3 it is clear that real values for A and r may be allotted. Similarly for power fed down the E-leg, we derive the relation :

$$Y_4 = \frac{1}{s^2}\left(2 - \frac{j}{B}\right) \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

from which B and s may be determined.

If the waveguide junction under the conditions necessary to obtain equations (10) and (11) could be replaced by a pure 3-way H- or E-junction respectively the admittances at certain specified points in the feeding leg could be obtained from the theoretical figures given by Cowling for any wavelength and size of guide. In fact, however, the existence of attenuated modes in the leg from which power is excluded, modify the behaviour of the junction. It is probable, however, that the theoretical data could be taken as a first approximation.

We shall carry the discussion of the determination of the constants of the junction no further, since we are primarily interested in seeing what modifications must be made to give the further symmetry properties of the magic-tee.

These modifications may be determined from the relationship (8), by finding what terminations are necessary in legs 3 and 4, power being fed down leg 1, to satisfy the

‡ We have adopted an analogy with a perfect transformer. For this analogy to hold we shall also have to adopt a convention regarding the sign of the admittance, which will be taken as positive when we are looking toward the junction and negative when we are looking away from it.

conditions that no power shall flow in leg 2 and that the generator shall see a perfect match. That is, the coefficient of Y_2 must vanish, and Y_1 must be unity.

$$1 - AB[r^2Y_3 + s^2Y_4 + r^2s^2Y_3Y_4] - j[(A + B) + Ar^2Y_3 + Bs^2Y_4] = 0$$

and $- AB[4+r^2Y_3+s^2Y_4]-j[A+B] = 0$

which reduce to:

$$\left(\frac{1}{A} - jr^2Y_3\right)\left(\frac{1}{B} - js^2Y_4\right) = -4$$

This equation is satisfied if:

$$\left.\begin{aligned} -Y_3 &= \frac{1}{r^2}\left(2 + \frac{j}{A}\right) \\ -Y_4 &= \frac{1}{s^2}\left(2 + \frac{j}{B}\right) \end{aligned}\right\} \quad \ldots \quad \ldots \quad (12)$$

Comparing equations (12) with (10) and (11), it will be seen that they represent the admittances required in legs 3 and 4 to balance out the standing waves arising in these legs when power is fed down them, legs 1 and 2 being perfectly matched. Calculation shows the correct points at which to insert matching diaphragms or rods to effect these conditions.

As may be expected equation (8) may be much simplified to represent the behaviour of the modified junction. Let Y_4 and Y_3 now represent the admittances in the E- and H-legs apart from those introduced by the modifications. Then we know for instance that if power is fed down leg 3, $Y_3 = 1$ when $Y_1 = Y_2 = -1$, and the coefficient of Y_4 vanishes; i.e.,

$$2ABs^2 - ABr^2s^2 - jBs^2 = 0$$

$$\frac{j}{A} + (r^2 - 2) = 0 \quad \therefore r^2 = 2,\ A = \infty.$$

Similarly it may be found that $S^2 = 2$, $B = \infty$ and substituting in equation (9) we get finally

$$2Y_1Y_2 + (Y_1 + Y_2)(Y_3 + Y_4) + 2Y_3Y_4 = 0 \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

This equation completely specifies the admittance properties of the modified magic-tee. The points P at which the admittances are to be measured may be found experimentally in the same way as for the unmodified junction.

It will be noted that equation (13) is symmetrical in the pair Y_1, Y_2 and the pair Y_3, Y_4 and also that the admittances of any pair are interchangeable. Properties that hold true when one leg is made the source of power will also hold when any other leg is made the source of power under similar conditions. For brevity we will consider the most useful case only and have the generator in the H-leg.

Two results follow from equation (13) which merit special attention.

(i) If we put $Y_1 = Y_2$ then (13) reduces to

$$(Y_3 + Y_1)(Y_4 + Y_1) = 0$$

so that $Y_4 = -Y_1$ independently of Y_3, as would be expected since power is excluded from leg 3 under these conditions. In particular if legs 1 and 2 are matched then $Y_4 = 1$. [Property (d)].

(ii) If we make $Y_2 = \frac{1}{Y_1}$ and match leg 3 so that $Y_3 = -1$ then (13) becomes

$$\left(2 - Y_1 - \frac{1}{Y_1}\right) - \left(2 - Y_1 - \frac{1}{Y_1}\right)Y_3 = 0$$

so that $Y_3 = 1$ independently of Y_1. Therefore, if legs 1 and 2 have equal loads but one leg is a quarter-wave longer than the other and if the E-leg is matched, then the generator sees a match.

3. Performance of a Simple Magic-Tee

We shall now leave the theoretical side, returning to it later when applications of the magic-tee are considered, and turn our attention to practical matters of construction, performance and bandwidth. A four-way junction was made from waveguide of 1 in × ½ in inside dimensions and was modified by the inclusion of matching diaphragms in the side legs. The position of these inductive diaphragms was determined, as indicated in the above section, by measuring the mismatch in the side legs when the main legs were correctly terminated, and calculating the position for the insertion of inductance to balance out this mismatch, positions as close to the junction as possible being chosen. The arrangement is sketched in Fig. 2. The wavelength at which these measurements were carried out was 3.10 cm.

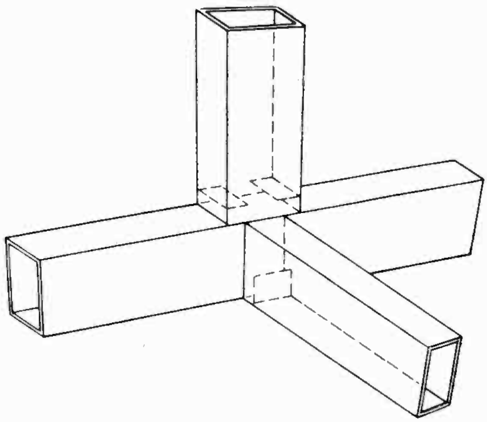
If the characteristic points of the main legs are taken as half-wavelengths from the plane of symmetry of the junction then the characteristic points of legs 3 and 4 are respectively $0.015\lambda_g$ and $0.378\lambda_g$ from the planes of the nearest side walls of the main guide. Further test results on this simple magic-tee follow.

(i) *At Design Frequency.*

Figures are given in Table I for the voltage standing-wave ratios arising from the completed magic-tee when the legs are correctly terminated, and also of the degree to which power is excluded from the leg "opposite" to the feeding member. In the worst case less than 0.2 per cent of the power "leaks" through.

TABLE I.

Feeding leg.	V.S.W. Ratio.	Attenuation to "opposite" leg
1	0.94	31 db
2	0.94	31 db
3	0.98	28 db
4	0.96	28 db



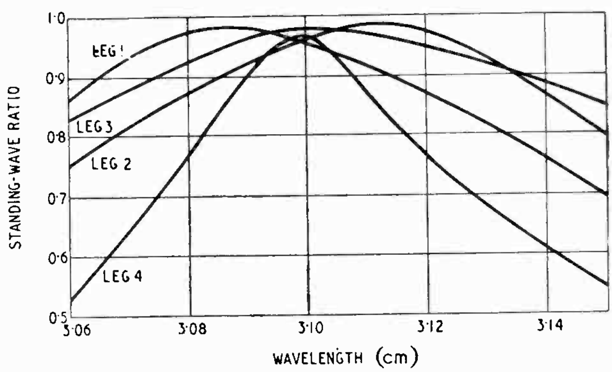


Fig. 2 (above). *Inductive diaphragms are included in the waveguide legs for matching.*

Fig. 3 (right). *Frequency sensitivity of matching of magic-tee shown in Fig. 2.*

(ii) *Over a Range of Frequency.*

Measurements have been made over the wave-band 3.06 cm to 3.20 cm of the standing-waves arising from the magic-tee with legs correctly terminated. Table II lists these results which are also plotted in Fig. 3. The last column of the table also shows the power ratio of coupling between legs 1 and 2 over the band 3.08 cm to 3.12 cm. The lack of mutual coupling between legs 3 and 4 is, of course, not sensitive to frequency.

TABLE II

Wave-length (cm)	V.S.W. Ratio. Leg 1	Leg 2	Leg 3	Leg 4	Attenuation (Legs 1–2) (db)
3.060	0.86	0.75	0.83	0.53	—
3.080	0.97	—	0.92	0.76	18
3.090	—	—	—	0.90	24.5
3.100	0.94	0.94	0.98	0.96	31
3.110	—	—	—	0.84	22.5
3.120	0.87	—	0.94	0.76	18
3.140	0.75	0.86	0.88	0.61	—
3.200	0.59	—	0.66	0.32	—

4. Magic-Tee with Increased Bandwidth

Consideration of the results for a magic-tee made with matching diaphragms in the side legs shows that the special properties (c) and (d) described in the introduction are not retained over a wide range of frequencies. Also the power leakage between legs 3 and 4 and between legs 1 and 2 is high enough to limit the application of the magic-tee as an impedance bridge for reasons which will be made apparent in the next section. Experimental effort has therefore been directed to the development of a junction which would maintain its theoretical properties to an improved degree over a wide range of frequencies.

One method of tackling the problem which suggested itself was to vary the size of the main guide relative to that of the side legs, the argument being that, in the absence of coupling between the latter the magic-tee could be regarded approximately as a combination of series and parallel T-junctions. This, of course, is not strictly the case, since for example if one were feeding the magic-tee from the E-leg the existence of attenuated modes in the H-leg would vary the behaviour to some extent from that of a series T-junction. Considering the series-tee waveguide junction it seemed probable that transmission from the branch to the main guides could be improved by lowering the equivalent impedance of the latter to roughly half that of the side leg. The equivalent impedance[3] of a waveguide in the H_{10} mode is taken to be equal to $\sqrt{\frac{\mu}{\kappa}\frac{\lambda_g}{\lambda}\frac{b}{a}}$ where λ_g is the guide wavelength, λ the free-space wavelength, and a, b are the lengths of the broad and narrow sides of the waveguide. In the series-tee the wide dimension is common to the three legs so the impedance is varied by reducing the narrow dimension of the main guide.

In the case of the parallel T-junction the impedance of the main legs should be raised relative to that of the side leg to improve transmission from the latter to the former. This can be effected by altering the wide dimension of the main guide. The experimental technique adopted with the magic-tee, therefore, was to reduce the dimensions of the main guide (and the corresponding *common* dimensions of the side legs) step by step, measuring for each case the standing-wave ratio in each of the side legs when the main waveguide branches had reflectionless terminations. These measurements were made over a range of wavelengths centred on 3.15 cm, and as a result a main waveguide size of 0.70 in × 0.28 in was chosen as giving the best results. Even with these modifications the junction still gave rise to an appreciable mismatch which remained to be balanced out by other means.

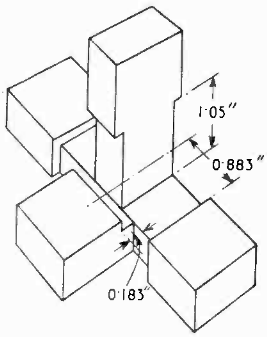


Fig. 4. *Sketch of modified magic-tee.*

For the device to be of general use it was, of course, necessary to increase or transform all the legs of the tee to the normal size of 1 in × 0.50 in. It should be pointed out however that for some applications it is possible to leave one or more of the waveguide legs in the reduced size. From the manufacturing viewpoint it was desirable to avoid the use of either tapered waveguide or of matching diaphragms. The practice has been, therefore, to make *sudden* transitions from one size to the other and to make them at such a distance from the junction that the mismatch caused by them tends to cancel the residual mismatch due to the junction itself. The final arrangement is made clear by the sketch of Fig. 4 and the experimental model is shown in Fig. 5.

The results of tests corresponding to those carried out on the magic-tee described in the previous section are shown in Tables III and IV and graphically in Fig. 6.

TABLE III
(At design frequency)

Feeding leg	V.S.W. Ratio	Attenuation to "opposite" leg
1 or 2	0.98	40 db
3	0.99	50 db
4	0.95	50 db

TABLE IV

Wave-length (cm)	V.S.W. Ratio Legs 1 or 2	Leg 3	Leg 4	Attenuation (Legs 1–2) (db)
0.375	0.78	—	0 85	—
3.100	0.82	0 77	0.89	23
3.150	0.98	0.99	0.95	40
3.200	0.74	0.69	0.89	23
3.225	—	—	0.81	—

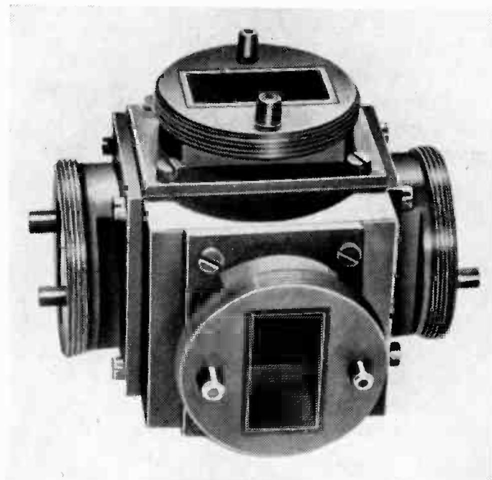


Fig. 5. *Photograph of the magic-tee junction sketched in Fig. 4.*

It will be noted, in particular, that the exclusion of power between "opposite" legs has been increased considerably, the figure of 50 db for the attenuation from leg 3 to leg 4 being near the limit of experimental accuracy.

It may be of interest here to indicate the method of manufacture adopted for producing the magic-tees. Early models were produced by a fabrication of lengths of waveguide tubing. Owing to distortions in brazing, etc., it is difficult, using this method, to maintain the high dimensional accuracy required for a junction of good performance.

A more satisfactory construction may be typified by explaining the method of production of the junction shown in Fig. 5.

Through a rectangular block of metal is broached a rectangular hole to form the E-leg. One side of the block perpendicular to the axis of this E-leg is machined so as to be flat and accurately at right angles to the E-leg and then in this face channels are milled for the main and H-legs, these channels being located from the position of the E-leg. The main and H-legs are completed by brazing on to the block a flat and substantial cover plate. The remaining faces of the block are then machined to the dimensions required for the sudden transitions to standard size of waveguide, adaptors made from this size of guide being finally fastened to the block as shown in Fig. 5.

Since doing the above work the authors have been informed by Dr. Harvey of T.R.E. that experiments at that establishment have shown that a magic-tee of much greater bandwidth can be obtained by a combination of the above method for the shunt junction (that is, the reduction of the narrow dimension of legs 1, 2 and 4) and the employment of a matching rod for the series junction. The position for the metal rod is in the plane of symmetry with the free end projecting somewhat along the E-leg in the axial direction.

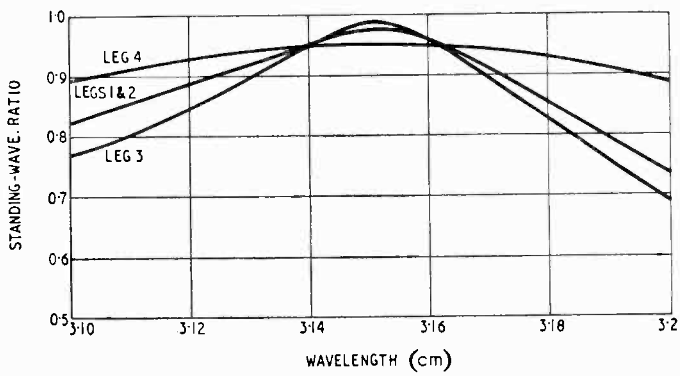


Fig. 6. *Frequency sensitivity of matching of broad-band magic-tee.*

5. Application of Magic-Tees

The possibilities of the magic-tee junction have probably not been fully explored. We shall set down here and examine in the light of the above theory some of the more obvious uses to which it has been or might be put.

(i) *The Impedance Bridge.*

If power is fed into the junction along the H-leg and a matched detector placed in the E-leg then the power in the latter is a measure of the out-of-balance of the loads in legs 1 and 2. In particular, if leg 1 is matched the power in leg 3 is a measure of the mismatch in leg 2. This fact can be used in the alignment of waveguide components, or in measuring an unknown admittance by comparison with a calibrated mismatch. The sensitivity of this microwave bridge may be determined as follows.

Equations (3) of Section 1, for the junction as modified to possess all the magic-tee properties, reduce to :

$$\left.\begin{aligned} Y_1V_{10} &= \frac{\sqrt{2}}{2}\Big[Y_3V_{30} + Y_4V_{40}\Big] \\ Y_2V_{20} &= \frac{\sqrt{2}}{2}\Big[Y_3V_{30} - Y_4V_{40}\Big] \end{aligned}\right\} \qquad (14)$$

Assuming first, that the Y-terms are all real, then from energy considerations

$$Y_1V_{10}^2 + Y_2V_{20}^2 + Y_3V_{30}^2 + Y_4V_{40}^2 = 0$$

and substituting for V_{10} and V_{20} from (14)

$$\begin{aligned} V_{40}^2\left[\frac{Y_4^2}{2Y_1} + \frac{Y_4^2}{2Y_2} + Y_4\right] & \\ + V_{30}^2\left[\frac{Y_3^2}{2Y_1} + \frac{Y_3^2}{2Y_2} + Y_3\right] & \\ + V_{30}V_{40}\left[\frac{Y_3Y_4}{Y_1} - \frac{Y_3Y_4}{Y_2}\right] &= 0 \end{aligned}$$

which on simplifying gives :

$$\frac{V_{30}}{V_{40}} = \frac{Y_4(Y_1 - Y_2)}{Y_2Y_3 + Y_1Y_3 + 2Y_1Y_2} \qquad (15)$$

If there is a matched detector in leg 3 then $Y_3 = -1$, $V_{30} = E_3$ where E_3 is the amplitude of the wave travelling down leg 3. The voltage V_{40} at the characteristic point of leg 4 is related to the incident voltage E_4 by the equation.

$$\frac{V_{40}}{E_4} = \frac{2}{Y_4 + 1}$$

Putting these conditions in equation (15) :

$$\frac{E_3}{E_4} = \frac{2Y_4}{(Y_4 + 1)}\,\frac{(Y_1 - Y_2)}{(2Y_1Y_2 - Y_1 - Y_2)}$$

but we can get Y_4 in terms of Y_1 and Y_2 when $Y_3 = -1$ from equation (13) that is

$$Y_4 = \frac{2Y_1Y_2 - Y_1 - Y_2}{2 - Y_1 - Y_2}$$

so that $\frac{E_3}{E_4} = \frac{Y_1 - Y_2}{(Y_1 - 1)(Y_2 - 1)}$ when Y_1 and Y_2 are real.

If leg 1 is also matched then

$$\frac{E_3}{E_4} = \frac{Y_2 + 1}{2(Y_2 - 1)} = \frac{\rho_2}{2} \quad \ldots \quad \ldots \quad (16)$$

where ρ_2 is the reflection coefficient in leg 2. Since one characteristic point may be arbitrarily fixed, we may choose it so that Y_2 is real for any given load, and then since Y_1 and Y_3 are always real and Y_4 is real if the other admittances are, equation (16) holds generally for any value of Y_2.

Some values for E_3/E_4 are given in Table V, of which the last row shows the difference in power level between the feeding member and the detector leg.

Let a perfect short circuit be placed in leg 1 so that $Y_1 = \infty$ then :

$$Y_4 = 1 - 2Y_2$$

If $Y_2 = 0$ we get a perfect match in the generator leg, but if Y_2 assumes a low value then Y_4 does not depart far from unity. Table VI shows the values of Y_4 to be expected for given *real* values of Y_2.

TABLE VI

$-Y_2$	0.01	0.05	0.10	0.20	0.50
Y_4	1.02	1.10	1.20	1.40	2.00
s.w.r. leg 4 ..	0.98	0.91	0.83	0.71	0.50

The procedure of measurement would be therefore to make the characteristic point in leg 2 such that Y_2 assumes a low real value and adjust the short-circuit in leg 1 until Y_1 is infinite at the characteristic point of this leg. The correct adjustment would be indicated by a minimum in the standing-wave ratio in the feeding member or a maximum in power along leg 3 and the position of the short-circuit along leg 1 would be a measure of the phase of the load on leg 2. Thus the method applies generally for the measurement of high s.w.r. of any phase.

TABLE V

Y_2	1.00	1.01	1.05	1.10	1.20	1.50	2.00
E_3/E_4	0	0.0025	0.0122	0.0238	0.0455	0.100	0.167
db_{3-4}	∞	52	38.5	32.5	27	20	15.5

These figures show that a very sensitive detector would be needed when using the tee for low-power work. The employment of a modulated signal and an amplifier would be essential for accurate matching or measurement.

In the case where leg 1 is not matched there may be a phase difference between Y_1 and Y_2 and the above theory does not strictly hold since E_3/E_4 will depend on this phase difference but the same order of sensitivity is to be expected.

(ii) *The Measurement of High Standing-Wave Ratios.*

Use may be made of the reciprocal relationship described at the end of Section 2 in the measurement of very high standing-wave ratios. In such cases the employment of a sliding-probe detector of conventional pattern can result in serious errors due to the power coupled to the detector, and more particularly due to instrument error, at low readings. By the method which the authors suggest the high s.w.r. is transformed in a determinable manner through the magic-tee to a low s.w.r. which is measurable more accurately by a sliding detector and presents a near match to the generator.

Starting with equation (13), put $Y_3 = -1$ and divide through by Y_1, then :

$$2Y_2 + (1 + Y_2/Y_1)(Y_4 - 1) - 2(Y_4/Y_1) = 0$$

(iii) *Transmission round a Difficult Corner.*

The magic-tee properties are based on the absence of power coupling between the E- and H-legs when in a symmetrical balanced condition. But the reciprocal relationship described above shows that for the case of extreme asymmetry optimum power transfer exists between these legs. At the end of Section 2 we have shown that if leg 3 is matched and we make $Y_1 = 1/Y_2$ then $Y_4 = 1$ independently of the values assumed by Y_1. In particular, if $Y_1 = 0$ and $Y_2 = \infty$, then all the power going into the junction emerges from the E-leg. This condition is obtained by putting short-circuits in the main legs at distances from the plane of symmetry differing by an odd number of quarter-wavelengths. It is possible that one of these distances could be made equal to half the wide dimension of the H-leg, so that the leg would become vanishingly small.

The value of this arrangement is that it circumvents the use of an E-bend and a 90° twist which would otherwise have to be used. If the shortest possible lengths of legs 1 and 2 are chosen the system should be of adequate bandwidth.

(iv) *Magic-Tee Phase Shifter.*

By suitable adjustment of short-circuits in the side legs the magic-tee can be made to give a pure phase shift to signals fed from leg 1 into leg 2. Suppose we have a generator in leg 1 and we make Y_3 and Y_4 purely imaginary and such that $Y_3 Y_4 = 1$. Let $Y_3 = jB$, then $Y_4 = -j/B$ and the admittance equation (13) reduces to:

$$2Y_1Y_2 + \left(Y_1 + Y_2\right)\left(jB - \frac{j}{B}\right) + 2 = 0$$

from which we may derive Y_1 in terms of Y_2 and B.

$$Y_1 = \frac{Y_2 - j\dfrac{2B}{(B^2 - 1)}}{1 - jY_2\dfrac{2B}{(B^2 - 1)}}$$

Replacing $\dfrac{2B}{(B^2 - 1)}$ by $\tan\theta$

$$Y_1 = -\left[\frac{Y_2 - j\tan\theta}{1 - jY_2\tan\theta}\right]$$

Changing the sign of Y_2 the expression takes on the more recognisable form:

$$\frac{Y_2 + j\tan\theta}{1 + jY_2\tan\theta}$$

This result is identical with that for the input admittance of a loss-free line of electrical length θ terminated in Y_2; i.e., it is the value of Y_2 when shifted in phase by θ.

To shift phase, therefore, short-circuits in legs 3 and 4 must be moved along their respective guides in such a manner that the distance of one short-circuit from the characteristic point is always a quarter-guide-wavelength different from the corresponding distance of the other short circuit.

The device, of course, will work equally well with the short-circuits in legs 1 and 2, transmission being between legs 3 and 4 as in the previous application.

(v) *Magic-Tee Frequency Discriminator.*

R. V. Pound[4] of the M.I.T. Radiation Laboratory has proposed the employment of a magic-tee as a frequency discriminator. In this device the signal whose frequency is to be controlled is coupled into leg 4 by means of a directive feed in the direction towards the junction. This and the other side leg are terminated in matched crystal detectors whose outputs are fed into a balanced amplifier. The two other legs are terminated, one in a high-Q cavity and the other in a short-circuit one-eighth wavelength further from the junction than the cavity window. Theoretically, if the cavity has negligible losses, at the resonant frequency the power coupled into the system will divide equally between the detectors and the resultant output will be zero. (This assumes that the crystals are of equal sensitivity at all power levels which assumption is the principal weakness of the idea.) However, if the frequency departs from resonance either one or the other crystal will receive more power, according to whether the frequency is high or low.

Assuming negligible loss the cavity input admittance may be represented by jB where B passes through zero on resonance. The admittance of the short-circuited leg at an equivalent distance from the junction is $j1$. Substituting these values in equation (13) leads to the following expression for Y_4.

$$Y_4 = G_4 + jB_4 = \frac{(1 - B)^2 - 2j(1 + B)^2}{(1 + B)^2 + 4} \quad \ldots \quad (17)$$

Now if E_4 is the voltage of the wave incident on the junction the energy transmitted is given by the expression

$\dfrac{4G_4E_4^2}{(1 + G_4)^2 + B_4^2}$ and since there is no loss in legs 1 and 2 this must be equal to the energy absorbed by the matched detector in leg 3, that is E_3^2 so that:

$$\frac{E_3}{E_4} = \frac{4G_4}{(1 + G_4)^2 + B_4^2} \quad \ldots \quad (18)$$

The proportion of the energy absorbed by the detector in leg 4 is of course, $1 - (E_3/E_4)^2$.

Substituting values for B in equations (17) and (18) leads to the result described above. Fig. 7 gives a plot of the resultant output expected from the amplifier. The effective bandwidth of the device is determined by the Q of the cavity, the effective range of B being from $+1$ to -1, for which values it may be easily seen that all the power goes into either the feeding leg crystal or the crystal in leg 3.

(vi) *Balanced-Crystal Mixer.*

A straightforward application for the magic-tee is in the mixer circuit of microwave radar or communications receivers. Matched crystal detectors are placed in legs 1 and 2, the signal is fed into the tee along the H-leg (say) and the local-oscillator energy along the E-leg. Due to the balanced conditions such an arrangement precludes the coupling of local-oscillator power into the signal leg. A further important advantage arises from the fact that in legs 1 and 2 energy coupled from the signal leg is in the same phase, and that coupled from the local-oscillator leg is oppositely phased at the two detectors. This means that whereas i.f. noise components due to the finite bandwidth of the local oscillator are generated in the crystals in phase, the i.f. signal components are generated in push-pull, so that an improvement in signal-to-noise ratio may be obtained by using an i.f. amplifier with push-pull input. Since the signal power divides equally between two detectors it may also be observed that double the received power is necessary to cause a crystal to burn out. Even if one crystal does fail it is still possible for the receiver to give some indication.

In a mixer designed by the authors it was found convenient to use the broad-band magic-tee described above with the modification that legs 1 and 2 were left in the reduced size (0.70 in × 0.28 in inside dimensions) it being found that crystal detectors in this size of guide had about double the bandwidth of standard-sized detectors.

6. Conclusions and Acknowledgments

It is hoped that this paper will lead to a wider understanding of the mechanism underlying the magic-tee wave-guide junction and to appreciation of its possibilities in microwave technique, some of which have been indicated above.

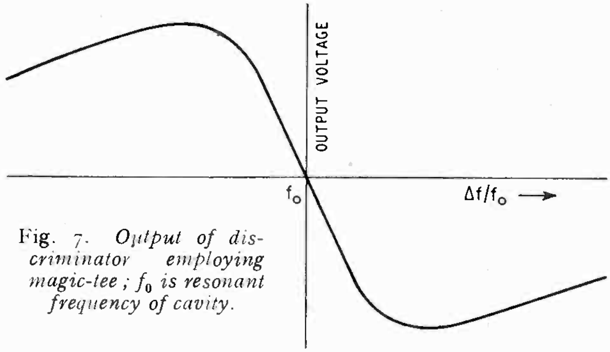


Fig. 7. *Output of discriminator employing magic-tee ; f_0 is resonant frequency of cavity.*

In conclusion, acknowledgment should be made of the assistance rendered by Dr. T. G. Cowling in the derivation of the admittance equation. Mention should also be made of the encouragement given to the authors by the technical staff of T.R.E. and helpful discussions with J. H. Briggs and Dr. A. E. Harvey are gratefully acknowledged. The authors also wish to thank Sir Arthur Fleming, C.B.E., D.Eng., Director, and B. G. Churcher, M.Sc., Manager of the Research Department at Metropolitan-Vickers Elec. Co. Ltd. for permission to publish this paper.

Addendum

Further experimental effort has been directed along the lines suggested at the end of Section 4 and has resulted in a broad-band magic-tee in which the matching in all legs is such that a voltage standing-wave ratio of better than 0.9 is obtained over a ± 2 per cent range of wavelength centred on 3.20 cm. This Tee is designed for use with waveguide of 0.9 in × 0.4 in internal dimensions.

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VALVE NOISE AND TRANSIT TIME

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(Communication from the Research Staff of the M.-O. Valve Company, Ltd. at the Research Laboratories of the General Electric Company, Wembley, England).

SUMMARY.—The treatment of noise developed in several recent papers by the authors is applied to valve noise when the transit time of the electrons is important.

Sections 2–9 express the reaction of the valve to a sinusoidal input in a form suitable to the "determinant" method of handling linear valve circuits. This part is based on the work of Benham on the plane diode, as expounded and extended to the plane triode by Llewellyn. Most of the assumptions made by these authors are adopted; but one of Llewellyn's is replaced by another.

Sections 10–14 deal with the modification of the formulae given previously rendered necessary by the finite duration of the "noise events." They involve a discussion of the charge induced by an electron on the grid as it approaches and recedes from the grid through (or with) a space charge. Alternative assumptions are proposed which lead to different numerical constants in the practical formulae.

Section 15. The formulae of Bakker and of Ferris and North for "induced grid noise" agree with our formulae only if the least plausible of these alternatives is adopted.

Section 16. It is doubtful whether Benham's and Llewellyn's assumptions are applicable to any valve of practical importance. In particular the effect of the electron current on the input capacitance appears to be definitely greater than that predicted by the theory. The position is therefore unsatisfactory; further theoretical and experimental work is required.

1. Introduction

THROUGHOUT this paper knowledge is assumed of three previous papers, denoted by A, B, C[1].

It was assumed explicitly in C that there were no "transit time effects; i.e., that the duration τ of a noise event (in particular the passage of an electron between two electrodes) was so small that $f\tau$ can be regarded as zero, where f is any frequency to which the system responds appreciably. In this paper we examine some of the effects of abandoning this assumption.

In A, Sect. 8, it was pointed out that there are two related transit time effects. One is on $\phi(j\omega)$, representing the response of the system to a sinusoidal input; the other is on $s(t)$, representing its response to a noise event. They will be taken in that order. In dealing with both problems assumptions will be made explicitly that are clearly not indubitable; the question whether they are legitimate in any practically important circumstances will be left to the final section; in all earlier sections we shall simply develop the consequences of these assumptions without, we hope, introducing any that are not stated explicitly.

MS accepted by the Editor, January, 1947.

† Formerly of the Staff of the G.E.C. Research Laboratories.

2. Sinusoidal Input

The effects of a sinusoidal input are introduced into C through the formulae of B. The question is, therefore, how these formulae are to be modified to take account of transit time. In B it is assumed that the terms g and α, which determine the electron currents between the electrodes, are all real and independent of frequency. That implies that these currents are in phase with the voltages between the electrodes, which is not true when transit time is appreciable. All the other assumptions made in B remain true; and the differences of phase between voltage and current in the external circuits are represented by complex and frequency-dependent admittances. It appears reasonable, therefore, to assume that the effect of transit time on the formulae of B can be represented *wholly* by taking the g and α terms in equation (1) of that paper to be complex and frequency-dependent.

3. Initial Electron Velocity

When this assumption is made, the equation that must replace (1) of B can be derived, for a plane negative-grid triode, from the results, based mainly on the work of Benham[2], set forth by Llewellyn in his tract on "Electron Inertia Effects" (Camb. Univ. Press, 1943). But it must be remembered that, as Benham recognized, the work is based on a logical fallacy. The method of calculation adopted necessarily assumes that all electrons leaving the cathode at any instant have the same velocity and acceleration, and therefore have the same velocity and acceleration at any point in the interelectrode space. But, if this were so, there would be nothing to distinguish them, and all or none must arrive at any other electrode. On the other hand, the assumption that there is a space-charge limited current, dependent on the voltage between the electrodes, means that some, but not all, of the electrons arrive at a second electrode. The only justification for adopting both of these mutually contradictory assumptions is that results concerning the steady state, derived by their use, agree closely with experiment. This is, of course, no justification for their use in more general circumstances; but since there appears to be no method of progressing at all without their use, we shall adopt them, recognizing that all our results are precarious.

Little further danger is incurred if we assume that the initial velocity of the electrons is zero (Llewellyn's "complete space charge"). Everyone finds the assumption necessary if formulae applicable to experiment are to be obtained; its danger is least if the empirical constants used are derived from the formulae, so that their values involve the assumption. It is, of course, most plausible when the current drawn from the cathode is only a small fraction of the emission.

4. Diode

For a plane parallel diode with complete space charge, Llewellyn's equation (4.15), with the second and third terms on the right omitted, gives the relation between V, the amplitude of a *small* voltage of frequency $\omega/2\pi$ imposed on the steady voltage between electrodes, and I_0, the amplitude of the resulting small alternating current imposed on the steady current I_0. With some change of notation‡ it can be written in practical units

‡ We have not hesitated to change Llewellyn's notation, because he himself has done so in later papers. It should be noted that our formulae relate to voltages and currents (not current densities), not to his W and J. Some of the trigonometrical functions $A(\theta)$, etc., are those which Benham and later writers have denoted by other symbols. It is highly desirable that the notation should be standardized; but since there is no agreement at present, we have not followed any previous writer.

where

$$V = I_t \,.\, \mathbf{A} = I_t \,.\, a_0 \,.\, A(\theta) \quad \ldots \quad \ldots \quad (4.1)$$

$$a_0 = T/2C = [I_0 T^4/S^2] \,.\, 1.87 \times 10^{40} \quad (4.2)$$

$$A(\theta) = \frac{12}{\theta^4}\left[\left(2 - 2\cos\theta - \theta\sin\theta\right) + j\left(2\sin\theta - \theta\cos\theta - \theta - \frac{\theta^3}{6}\right)\right] \quad (4.3)$$

$$= 1 - \frac{\theta^2}{15} + \frac{\theta^4}{560} \ldots + j\left(-\frac{3\theta}{10} + \frac{\theta^3}{84} \ldots\right) \ldots \quad \ldots \quad (4.4)$$

$$\theta = \omega T \quad \ldots \quad \ldots \quad \ldots \quad (4.5)$$

and

T = transit time of the electrons in the steady state,

S = area of the electrodes,

x = distance between the electrodes,

$C = S/(4\pi x \times 9 \times 10^{11})$ .. .. (4.6)
= capacitance between the electrodes in farads.

When $\theta \to \infty$, $I_t \to j\omega CV$. Now $T \to \infty$ implies $I_0 \to 0$, and $\omega \to \infty$ means that the electrons cannot move owing to their inertia. Hence $\theta \to \infty$ implies that the electrons have no effect; and it is right then that the current should reduce to that through the interelectrode capacitance in the absence of electrons. In B we regarded this capacitance current as belonging to the external circuit; it is convenient to maintain this convention (although there is now less need for it) in order to avoid needless changes in the formulae. Accordingly we shall write

$$I = I_t - j\omega CV \quad \ldots \quad \ldots \quad \ldots \quad (4.7)$$

and call I the electron current. We have then

$$I = V(1/\mathbf{A} - j\omega C) = V/\mathbf{A}' \text{ (say)} \quad \ldots \quad (4.8)$$

5. Triode

All this applies to a diode. We shall follow Llewellyn and many others in assuming that a negative-grid triode is equivalent to a pair of diodes separated by a continuous plane, occupying roughly the position of the grid wires and completely permeable to electrons. This plane (G-plane) acts as anode for the cathode of the triode and as cathode for its anode; the electrons, in passing through the G-plane from one diode to the other, suffer no change of velocity, but a change of acceleration corresponding to the difference of the fields on the two sides of it.

Some assumption has to be made about V_G, the a.c. component of the voltage between the G-plane and the cathode of the triode. Here we shall not follow Llewellyn. He derives V_G from the assumption that the impedance between the G-plane and the grid wires is a pure capacitance, on the ground that the space between them is free from electrons. This statement appears contrary to the existence of an "induced grid current" (see Section 10) arising from the passage of electrons through the grid. We shall make the simpler assumption that

$$V_G = \mu_2 V_{21} + \mu_3 V_{31} \quad \ldots \quad \ldots \quad (5.1)$$

where V_{21}, V_{31} are the a.c. components of the voltages of grid to cathode and anode respectively, and μ_2, μ_3 are real constants independent of frequency. This must be true for sufficiently low frequencies; the question is only over what range of frequency it is true in given circumstances.

6. Cathode-Grid Section of Triode

We now study the diode consisting of cathode and G-plane, using the suffix c to denote quantities characteristic of it. If I_c is the electron current flowing from the cathode, we have from (4.8)

$$I_c = \mu_2 V_{21}/\mathbf{A}' + \mu_3 V_{31}/\mathbf{A}' \quad \ldots \quad \ldots \quad (6.1)$$

If this is compared with equation (1) of B, it appears that $\mu_2/\mathbf{A}'_c$, $\mu_3/\mathbf{A}'_c$ are the complex and frequency-dependent quantities y_2, y_3 which (according to Section 2) are to replace the mutual and anode conductances; these conductances will be written g_{20}, g_{30}. Further since no lower limit of frequency is imposed on (5.1), y_2, y_3 must become g_{20}, g_{30} when f and θ are zero. But, when $\theta = 0$, $1/\mathbf{A}'_c = 1/\mathbf{A}_c = 2C_c/T_c$. Hence we have

$$(g_{20}, g_{30}) = (\mu_2, \mu_3) \,.\, 2C_c/T_c \quad \ldots \quad (6.1)$$

$$(y_2, y_3) = (g_{20}, g_{30}) \cdot g(\theta) \quad \ldots \quad \ldots \quad (6.2)$$

where

$$g(\theta) = \frac{a_0}{\mathbf{A}'_c} = \frac{1}{A(\theta)} - j\omega C_c \cdot \frac{T_c}{2\omega C_c} = \frac{1}{A(\theta)} - j\frac{\theta}{2} \quad \ldots \quad \ldots \quad \ldots \quad (6.3)$$

$$= 1 - \frac{7\theta^2}{300} \ldots + j\left(-\frac{\theta}{5} + \frac{23\theta^3}{21000} \ldots\right) \quad (6.4)$$

y_2, y_3 would therefore be known if T_c, and therefore θ, were known. T_c cannot be measured directly; if it is to be calculated, the calculation should, according to Section 2, depend as much as possible on the equations already derived. From (4.2) we have

$$T_c{}^3 = 2.67 \times 10^{-41} \,.\, S^2/I_0 C_c \quad \ldots \quad (6.5)$$

while $g_{20} = \mu_2/a_0$ gives

$$T_c{}^4 = 5.34 \times 10^{-41} \cdot \mu_2 S^2/I_0 g_{20} \quad \ldots \quad (6.6)$$

S, I_0, g_{20} are measurable. C_c, μ_2 are not; but some estimate of their values can be made. For C_c should not differ greatly from the cathode-grid capacitance measured with no emission from the cathode; and in a close-mesh grid μ_2 must be nearly 1. If, when these approximate values are used, the values of T_c deduced from (6.5) and (6.6) agree well, we are justified in using either of them to determine $g(\theta)$. If they diverge widely, the only legitimate conclusion is that the assumptions of Section 3 are inadequate, and that the whole theory must be recast.

7. Grid-Anode Section of Triode

We now turn to the α terms, and for this purpose study the diode whose cathode is the G-plane and whose anode is the anode of the triode; characteristics of this diode will be represented by suffix a. The voltage across this diode is

$$V_a = V_{31} - V_G = (1 - \mu_3)V_{31} - \mu_2 V_{21} \quad \ldots \quad \ldots \quad (7.1)$$

Llewellyn's equations (6.3), when the initial velocities from the cathode of the triode are zero, can be written

$$V_a = I_{ta}\mathbf{A}_a + I_{tc}(\mathbf{G}_c\mathbf{B}_a + \mathbf{D}_c\mathbf{C}_a) \quad \ldots \quad (7.2)$$

or

$$I_{ta} = V_a/\mathbf{A}_a - I_{tc}(\mathbf{G}_c\mathbf{B}_a + \mathbf{D}_c\mathbf{C}_a)/\mathbf{A}_a \quad (7.3)$$

where I_{tc}, I_{ta} are the total currents, including the capacitance current, across the two diodes, and $\mathbf{A}_a$, $\mathbf{B}_a$, $\mathbf{C}_a$, $\mathbf{D}_c$, $\mathbf{G}_c$ are functions of the transit angles that it will be more convenient to give later. I_a, the electron current through the second diode, is $I_{ta} - j\omega C_a V_a$. Hence, if we write,

$$1/\mathbf{A}'_a = 1/\mathbf{A}_a - j\omega C_a \quad \ldots \quad \ldots \quad (7.4)$$

and remember that $I_{tc} = I_c \,.\, \mathbf{A}'_c/\mathbf{A}_c$, we have

$$I_a = \frac{V_a}{\mathbf{A}'_a} - I_c \cdot \frac{\mathbf{A}'_c}{\mathbf{A}_c} \cdot \frac{\mathbf{G}_c\mathbf{B}_a + \mathbf{D}_c\mathbf{C}_a}{\mathbf{A}_a} \quad (7.5)$$

Now, by definition, $\alpha_3 = I_a/I_c$. Hence, substituting for I_c from (6.1),

$$\alpha_3 = -\frac{\mathbf{G}_c\mathbf{B}_a + \mathbf{D}_c\mathbf{C}_a}{\mathbf{A}_a} \cdot \frac{\mathbf{A}'_c}{\mathbf{A}_c} - \frac{(\mu_3 - 1)V_{31} + \mu_2 V_{21}}{\mu_3 V_{31} + \mu_2 V_{21}} \cdot \frac{\mathbf{A}'_c}{\mathbf{A}'_a} \quad \ldots \quad (7.6)$$

α_3 thus depends on the a.c. components of the electrode potentials, whereas it was assumed in B that it was independent of them. However α_3 would become independent of them if V_{21}/V_{31} were constant; and this would be true if the components arose entirely from a signal generator always applied between the same pair of terminals. In any problem to which this paper is relevant, this will be true for the regular components; for the triode will be part of either a common grid or a common-cathode amplifier. In all strictness it should be recognized that V_{21}, V_{31} have irregular components arising from the noise events; but we are compelled to assume that these cancel out, and can be ignored, by our fundamental assumption in A that the effect of any noise event is the same whatever other noise events have preceded it. Accordingly there is nothing in (7.6) to suggest that the formulae of B may not still be applied to any relevant problem, so long as it is remembered that α_3 is complex and frequency-dependent.

The current flowing to the G-plane is the difference between that flowing from the cathode and that flowing to the anode of the triode. This is true whether electrons are or are not present. Accordingly, if I_G is the electron current to the G-plane,

$$I_c = I_G + I_a \qquad (7.7)$$

and, since α_2 is defined as I_G/I_c,

$$\alpha_2 = 1 + \frac{\mathbf{G}_c\mathbf{B}_a + \mathbf{D}_c\mathbf{C}_a}{\mathbf{A}_a} \cdot \frac{\mathbf{A}'_c}{\mathbf{A}_c} + \frac{(\mu_3 - 1)V_{31} + \mu_2 V_{21}}{\mu_3 V_{31} + \mu_2 V_{21}} \cdot \frac{\mathbf{A}'_c}{\mathbf{A}'_a} \qquad (7.8)$$

8. Determination of Coefficients

It is now time to give the functions $\mathbf{A}_a$, $\mathbf{B}_a$, $\mathbf{C}_a$, $\mathbf{G}_c$, $\mathbf{D}_c$. Let the distance between the G-plane and the cathode be x_c, that between the anode and the G-plane x_a. Let T_a be the time for the electrons to traverse x_a in the steady state. Let

$$\phi = \omega T_a \qquad (8.1)$$

$$y = x_a/x_c \qquad (8.2)$$

$$h = T_a/T_c = \phi/\theta \qquad (8.3)$$

If g is the ratio of the acceleration of the electrons when leaving the G-plane on the anode side in the steady state to their acceleration when entering it on the cathode side, then elementary considerations show that

$$gh^2 + h = (y - h^3)/3 \qquad (8.4)$$

The functions can now be stated in terms of g, h, and the transit angles ϕ, θ:—

$$\mathbf{A}_a = a_0\left[-j\frac{6(gh^2 + h)}{\theta} + h^4 A(\phi)\right] \qquad (8.5)$$

$$\mathbf{B}_a = a_0 \cdot j\frac{6}{\theta}\left[gh^2 X_1(\phi) + hX_0(\phi)\right] \qquad (8.6)$$

$$\mathbf{C}_a = a_0 - \frac{6h^2}{T_c^2} X_1(\phi) \qquad (8.7)$$

$$\mathbf{D}_c = j\frac{T_c^2}{\theta}\left[X_1(\theta) - X_0(\theta)\right] \qquad (8.8)$$

$$\mathbf{G}_c = X_1(\theta) \qquad (8.9)$$

where $A(u)$ is the function already defined in (4.3) and

$$X_0(u) = \frac{1}{u}\left[\sin u + j(\cos u - 1)\right] = 1 - \frac{u^2}{6} + \frac{u^4}{120} \ldots + j\left(-\frac{u}{2} + \frac{u^3}{24} \ldots\right) \qquad (8.10)$$

$$X_1(u) = \frac{2}{u^2}\left[u \sin u + \cos u - 1 + j(u \cos u - \sin u)\right] = 1 - \frac{u^2}{4} + \frac{u^4}{72} \ldots + j\left(-\frac{2u}{3} + \frac{u^3}{15} \ldots\right) \qquad (8.11)$$

When $\omega = 0$ and $\phi = \theta = 0$, dashed and undashed quantities are equal, and all the trigonometrical functions are 1. Then

$$\mathbf{A}'_c/\mathbf{A}_c = 0; \ \mathbf{G}_c\mathbf{B}_a/\mathbf{A}_a = -1; \ \mathbf{D}_c\mathbf{C}_a/\mathbf{A}_a = 0 \qquad (8.12)$$

so that α_2 has its low-frequency value 0, as it should.

It does not appear possible to determine g, h, from low-frequency measurements in a manner analogous to that by which T_c was determined in Section 6. Llewellyn, at the end of his Chapter III, determines them from steady-current measurements. But his method requires an assumption about the voltage of the G-plane when the current is steady that is not necessarily consistent with that concerning the a.c. component of that voltage when $\omega \to 0$. Moreover it requires x_a, x_c to be known; but neither of these, nor even their sum, can be determined by measurement of the inter-electrode distances in the valve; for the G-plane need not coincide with the centres of the grid wires, nor the effective cathode (in view of the assumption of zero initial velocities) with the actual cathode. Nevertheless his method, with approximate values of the x terms, almost certainly gives values of the right order. From these it appears that, in all normal valves, h is a small fraction generally less than 0.2. It becomes interesting therefore to examine the limits of the formulae when $h \to 0$. We then have

$$\mathbf{B}_a = -\mathbf{A}_a\,;\ \mathbf{D}_c\mathbf{C}_a = 1/\mathbf{A}'_a = 0 \quad \ldots \quad (8.13)$$

Consequently

$$\alpha_2 = 1 - \mathbf{G}_c \,.\, \mathbf{A}'_c/\mathbf{A}_c$$

$$= 1 - \mathbf{G}_c \Big/ \left[1 - j\frac{\theta}{2} \,.\, A(\theta)\right] \quad \ldots \quad (8.14)$$

$$= \frac{\theta^2}{60} \ldots + j\left(\frac{\theta}{6} - \frac{\theta^3}{60} \ldots\right) \quad \ldots \quad (8.15)$$

9. Input Admittance

According to B, Section 8, the input admittance to the cathode-grid of a common-cathode triode is

$$Y_{22i} = \frac{\Delta}{\Delta_{22}} = \alpha_2 y_2 + Y_{12} + Y_{23} - \frac{(\alpha_3 y_2 - Y_{23})(\alpha_2 y_3 - Y_{23})}{\alpha_3 y_3 + Y_{13} + Y_{23}} \quad \ldots \quad (9.1)$$

If Y_{13} is so large that the last term is negligible, the part of the admittance associated with the electron current is, from (6.2), (6.4), (8.15),

$$\alpha_2 y_2 = \mu_2 \,.\, \frac{2C_c}{T_c}\left(\frac{\theta^2}{60} + \ldots + j\frac{\theta}{6} \ldots\right) \left(1 - \frac{7\theta^2}{300} \ldots - j\frac{\theta}{5} \ldots\right) \quad \ldots \quad (9.2)$$

$$= \mu_2 \,.\, \frac{2C_c}{T_c}\left(j\frac{\theta}{6} \ldots + \frac{\theta^2}{20} \ldots\right) \quad \ldots \quad (9.3)$$

The term in θ is $j\omega\mu_2 C_c/3$ and can, therefore, be regarded as representing the presence of an additional cathode-grid capacitance C' due to the electron current, where

$$C' = \mu_2 C_c/3 \quad \ldots \quad \ldots \quad \ldots \quad (9.4)$$

In a diode, in which the anode receives the full electron current, $\alpha_2 = \mu_2 = 1$. Consequently the term of $\alpha_2 y_2$ in θ is $-j\omega.2C_c/5$, and

$$C' = -2C_c/5 \quad \ldots \quad \ldots \quad \ldots \quad (9.5)$$

This effective increase of input capacitance in the triode and this effective decrease in the diode are in accordance with Llewellyn's equations (6.50), (5.24)—due allowance being made for the rather different approximations —and are *qualitatively* in accordance with the facts. They represent the first-order effects of transit time; the effects that are usually called "transit-time effects" are those of the second and higher orders.

When in B it was assumed that the α and g terms were wholly real, that the electron currents were in phase with the voltages producing them, and that any out of phase currents belonged to the external circuit, it was assumed in effect that the first order effects of transit time were negligible. These first-order effects, but not those of higher order, can be taken into account by using for the interelectrode capacitances those determined with the electron current flowing instead of those determined in its absence.

10. Induced Grid Noise

We now turn to the second part of our inquiry, which concerns what other writers have termed "induced grid noise," and have explained in some such terms as these. As an electron, carrying charge Q, approaches the grid from the cathode, it induces an increasing charge on the grid until, when it is in the grid plane, it induces practically its full charge. As it passes on towards the anode, the charge that it induces on the grid decreases, while that which it induces on the anode (small until it has passed the grid) increases. When it is just outside the anode it induces its full charge on the anode and none on the grid; no further change occurs when its falls into the anode. If ηQ with suffix c or a is the charge induced at any instant on grid or anode, the current flowing to the electrode will be

$$\xi(t) = Q \,.\, d\eta/dt \quad \ldots \quad \ldots \quad \ldots \quad (10.1)$$

The η and ξ terms will, therefore, be functions of time, the time since the electron left the cathode, and roughly of the forms shown in the figure.

The passages of electrons through the grid to the anode are the noise events. In strictness they vary with the a.c. components of the electrode voltages, and with the exact path of the electron through the grid. We shall assume that it is sufficiently accurate to suppose that only the steady voltages need be considered, and—this is more precarious—that, when the grid is replaced by the G-plane, so that all the electron paths are similar, the effect is the same as that of taking the mean over all paths through the actual grid.

In order to deal with such noise events by our standard procedure, we need, in a slightly modified form, a proposition stated without proof in A, Section 8. Accordingly we shall start by proving it.§

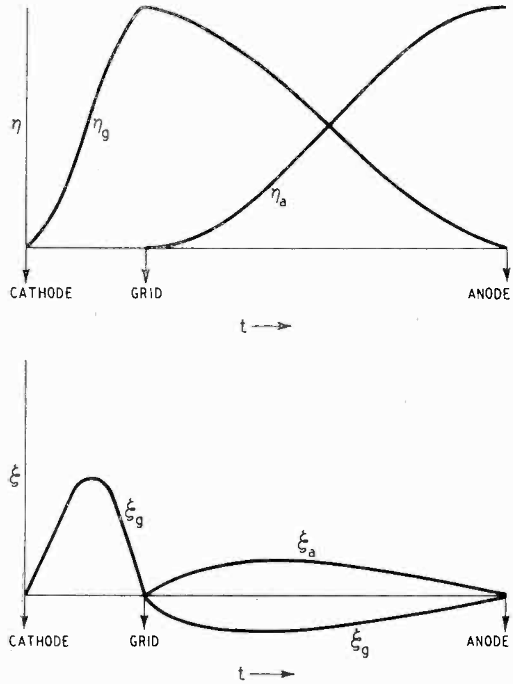
Let $s_0(t)$ be the effect at the output of the instantaneous transference, at $t = 0$, of a charge Q across the input. Let $s_T(t)$ be the

§ We owe this proof to Mr. D. O. North of R.C.A.

effect of the passage across the input of a current $i = \xi(t)$, which is zero except when $0 < t < T$. The charge conveyed by this current during dt_1 at t_1 is $\xi(t_1)dt_1$; the interval between the passage of this charge and the observation of the output at t is $t - t_1$. Consequently the contribution of this charge to the value of $s_T(t)$ at t is $\xi(t_1)dt_1/Q \cdot s_0(t - t_1)$. Hence

$$s_T(t) = 1/Q \cdot \int_0^t \xi(t_1) \cdot s_0(t - t_1)dt_1 \quad (10.2)$$

But $\xi(t_1) = 0$ when $t_1 < 0$; and $s(t - t_1) = 0$ when $t - t_1 < 0$, and $t_1 > t$. The integrand in (10.2) is zero except within the limits 0 to t, and the integral is not changed if these limits are replaced by $-\infty$ to $+\infty$.



Variation of induced charge and induced current as a charge passes through the grid.

Now it is a known property of Fourier transforms that, if $f(t)$, $F(j\omega)$ are one pair of transforms and $g(t)$, $G(j\omega)$ another pair, then

$$f(t)g(t - u)du$$

and

$$\sqrt{2\pi}F(j\omega)G(j\omega) \quad \ldots \quad \ldots \quad \ldots \quad (10.3)$$

are transforms. Hence, if $S_0(j\omega)$ is the transform of $s_0(t)$ and $\Xi(j\omega)$ of $\xi(t)$, the transform of $s_T(t)$ is

$$S_T(j\omega) = 1/Q \cdot \sqrt{2\pi}\, \Xi(j\omega)S_0(j\omega) \quad \ldots \quad (10.4)$$

or, by A (4.8),

$$= \Xi(j\omega)\phi(j\omega) \quad \ldots \quad \ldots \quad \ldots \quad (10.5)$$

where, if the effect at the output is a voltage, $\phi(j\omega)$ is the transfer impedance from input to output.

(10.4) means that the current $\xi(t)$, transferring a total charge $\int_0^T \xi(t)dt$ during the period T, produces the same effect on a system responding only to the frequency $\omega/2\pi$ as a particle passing instantaneously would produce if it carried a charge $\sqrt{2\pi}\, \Xi(j\omega)$.

11. The R.M.S. Noise

However we cannot use the proposition in quite this simple form; for the currents $\xi_c(t)$, $\xi_a(t)$, which flow simultaneously and inseparably, act at different inputs. If $s_{c0}(t)$, $s_{a0}(t)$ are the effects at the same output of a charge Q passing instantaneously from the cathode to the grid and the anode respectively, we have for their combined effect, in place of (10.2),

$$s_T(t) = 1/Q \cdot \int_0^t \left\{ \xi(t_1)s_{c0}(t - t_1) + \xi_a(t_1)s_{a0}(t - t_1) \right\} dt_1 \quad \ldots \quad \ldots \quad (11.1)$$

and, in place of (10.4)

$$S_T(j\omega) = \sqrt{2\pi}/Q \left\{ \Xi_c(j\omega)S_{c0}(j\omega) + \Xi_a(j\omega)S_{a0}(j\omega) \right\} \quad \ldots \quad \ldots \quad (11.2)$$

By the fundamental equation C (2.2)

$$\overline{(v - \bar{v})^2} = 2\lambda \cdot \frac{2\pi}{Q^2} \int_0^\infty |\Xi_c(j\omega) S_{c0}(j\omega) + \Xi_a(j\omega)S_{a0}(j\omega)|^2 d\omega \quad (11.3)$$

But λ, the mean frequency of the events, is still I_0/ϵ; and the transfer impedances $\phi(j\omega)$ of A turn out to be the ρ terms of C. Hence if the notation of C for the electrodes is used, we have in place of C (8.3)

$$v_\sigma^2 = \frac{2I_0}{\epsilon} \int_0^\infty |\Xi_c(j\omega)\rho_{13,12} + \Xi_a(j\omega)\rho_{13,13}|^2 d\omega \quad \ldots \quad \ldots \quad (11.4)$$

Let us write $H(j\omega)$ for the Fourier transform of $d\eta/dt$, and $\Gamma_{T\epsilon}$ for the Q of Section 10, so that (10.1) becomes

$$\Xi(j\omega) = \Gamma_{T\epsilon}H(j\omega) \quad \ldots \quad \ldots \quad (11.5)$$

Then (11.4) becomes

$$v_\sigma^2 = 2I_0\epsilon\Gamma_T^2 \int_0^\infty |H_c(j\omega)\rho_{13,12} + H_a(j\omega)\rho_{13,13}|^2 d\omega \quad \ldots \quad \ldots \quad (11.6)$$

It is important to observe that (11.6) cannot be put into the form appropriate to two independent emissions of electrons, one of which can effect one input only and the other the other input only; for in that form the integrand breaks into two terms, one proportional to $|\rho_{13, 12}|^2$ and the other to $|\rho_{13, 13}|^2$. Nor can it be reduced to the form C(8.11), which is appropriate when each of a set of randomly emitted electrons affects both inputs in one of two alternative ways, the choice between which is random. In our case, each electron of a random set affects both inputs in ways that are the same for all members of the set.

12. The Function η

If the argument of Section 10 is correct, Γ_T must be the same as North's Γ. For both have the same physical meaning, namely that $\Gamma\epsilon$ is the total charge received by the anode as a consequence of a noise event; i.e., the emission of an "extra" electron from the cathode. Since it is assumed, in estimating this charge, that the state is steady and that no a.c. voltage is applied, neither Γ_1 nor Γ can vary with ω. But they vary with the transit time T, because they are determined by factors (e.g., the distance and steady voltage between the electrodes and the current flowing between them) which also determine T.

But it is not certain that the argument is correct, and that we may assume that a single charge moves through the grid from cathode to anode. For, according to A, Section 7, Γ is less than 1 because the real charge on the electron is associated with a "deficiency charge." Can we assume that the deficiency charge is always coincident with the real charge, and that it remains constant during the motion? In spite of these doubts we shall make the assumption.

If then z is the distance from the cathode of an extra electron travelling across a diode, we can write

$$d\eta/dt = d\eta/dz \,.\, dz/dt \quad \ldots \quad (12.1),$$

and identify dz/dt with the velocity of one of the main body of electrons, carrying the current I_0, at the same distance. If $t = 0$ at the instant when the electron leaves the cathode, then Llewellyn's equations (3.1), (3.3) shows that, when there is complete space charge,

$$dz/dt = 3xt^2/T^3 \quad \ldots \quad (12.2),$$

where x is, as before, the separation of the plates of the diode and T is the transit time across it. If, on the other hand, there is no space charge

$$dz/dt = v + \gamma t \quad \ldots \quad (12.3),$$

where v is the velocity with which the electron starts from the cathode and γ is the acceleration to which it is subject;

$$x = vT + \tfrac{1}{2}\gamma T^2 \quad \ldots \quad (12.4),$$

Concerning η the simplest assumption is that the space charge does not affect the charges that the extra electron induces on the electrodes when it is in any given position. We shall consider this assumption first, and examine an alternative later. Then it is a known result (see e.g., Shockley[3]) that, if η_A, η_C refer respectively to the charges induced on the anode and cathode of the diode,

$$\eta_A = z/x\,; \quad \eta_C = 1 - z/x \quad \ldots \quad (12.5)$$

$$d\eta_A/dz = -\,d\eta_C/dz = 1/x \quad \ldots \quad (12.6)$$

13. Determination of H_c and H_a

We assume that, in the pair of diodes equivalent to the triode, the cathode-grid diode has complete space charge and the grid-anode diode no space charge. Then

$$d\eta_c/dt = 3t^2/T_c^3 \quad (0<t<T_c) \quad \ldots \quad (13.1)$$

$$= -\frac{v_a + \gamma(t - T_c)}{x_a} \quad (T_c<t<T_c + T_a) \quad \ldots \quad (13.2)$$

$$d\eta_a/dt = +\frac{v_a + \gamma(t - T_c)}{x_a} \quad \ldots \quad (13.3)$$

$$\sqrt{2\pi}\, H_c(j\omega) = \int_{-\infty}^{+\infty} d\eta_c/dt \,.\, e^{-j\omega t}\, dt \quad (13.4)$$

$$= \int_0^{T_c} 3t^2/T_c^3 \,.\, e^{-j\omega t}\, dt - \int_{T_c}^{T_a+T_c} \frac{v_a + \gamma(t-T_c)}{x_a}\, dt \quad (13.5)$$

or, writing $t_1 = t - T_c$

$$= \int_0^{T_c} 3t^2/T_c^3 \,.\, e^{-j\omega t}\, dt - e^{-j\omega T_c} \int_0^{T_c} \frac{v_a + \gamma(t-T_c)}{x_a} \,.\, e^{-j\omega t_1}\, dt_1 \quad (13.6)$$

The integrals involved here all belong to the set

$$X_n(u) = \int_0^T \frac{n+1}{T^{n+1}}\, t^n \,.\, e^{-j\omega t}\, dt \quad \ldots \quad (13.7)$$

where $u = \omega T$. The members $X_0(u)$, $X_1(u)$

have already been given in (8.10), (8.11); the next two members are

$$X_2(u) = \frac{3}{u^3}\Big[u^2 \sin u + 2u\cos u - 2\sin u + j(u^2\cos u - 2u\sin u - 2\cos u + 2)\Big]$$

$$= 1 - \frac{3u^2}{10} + \frac{u^4}{56}\ldots + j\left(-\frac{3u}{4} + \frac{u^3}{12}\ldots\right) \quad \ldots \ldots (13.8)$$

$$X_3(u) = \frac{4}{u^4}\Big[u^3\sin u + 3u^2\cos u - 6u\sin u - 6\cos u + 6 + j\,(u^3\cos u - 3u^2\sin u - 6u\cos u + 6\sin u)\Big]$$

$$= 1 - \frac{u^2}{3} + \frac{u^4}{48}\ldots + j\left(-\frac{4u}{5} - \frac{2u^3}{21}\ldots\right) \quad \ldots \ldots (13.9)$$

(13.6) now becomes

$$\sqrt{2\pi}\, H_c(j\omega) = X_2(\theta) - (\cos\theta - j\sin\theta)\,.\,1/x_a\,.\,\{v_a T_a X_0(\phi) + \gamma T_a^2/2.X_1(\phi)\} \quad \ldots (13.10)$$

If we proceed to the limit $h = 0$ (Section 8), when $\phi = 0, X_0(\phi) = X_1(\phi) = 1$, (this limit is consistent with the assumption that there is no space charge in the grid-anode space), (13.10) becomes, in virtue of (12.4)

$$\sqrt{2\pi}\, H_c(j\omega) = X_2(\theta) - (\cos\theta - j\sin\theta) = \theta^2/5\ldots + j\theta/4 \quad \ldots \ldots (13.11)$$

so that, ignoring all but the lowest power of θ,

$$2\pi|H_c(j\omega)|^2 \approx \theta^2/16 \quad \ldots \ldots (13.12)$$

Obviously

$$\sqrt{2\pi} H_a(j\omega) = \cos\theta - j\sin\theta\,;$$

$$2\pi|H_a(j\omega)|^2 = 1 \quad \ldots \ldots (13.13)$$

14. Space Charge

But the assumption on which Section 13 is based, namely that the presence of the space charge does not affect the charge induced by the extra electron on the electrodes when it is at a given z, may be false. For the presence of the extra electron will distort slightly the distribution of the main electrons forming the space charge, and may therefore, affect the charge that they induce on the electrodes. It is very difficult, perhaps impossible, to calculate directly the magnitude of this effect; accordingly we shall deduce it from an assumption that is admittedly very speculative.

Suppose that a capacitor with plates C, A is charged to a voltage V_0 by the transference of a charge Q from C to A. Then a point P between the plates will assume some voltage V relative to C. It can be proved that, if a charge Q is now transferred from C to P, the charge induced on A is $V/V_0.Q$; in other words, for such an arrangement $\eta_A = V/V_0$. For a parallel plate capacitor with uniform dielectric, we thus get $\eta_A = z/x$, agreeing with (12.5). But the proposition is true even if the dielectric is not uniform, so that the field is not uniform even if the plates are parallel planes. Let us assume then that a parallel diode is equivalent for the purpose of this proposition to a parallel-plate capacitor filled with a non-uniform dielectric, such that the distribution of field and voltage between the plates is that actually obtaining in the space-charge diode. This assumption, combined with the previous assumption that the extra electron moves with the space-charge electrons, leads to $\eta_c = V(t)/V_G$, where $V(t)$ is the voltage relative to C of the point that the electron occupies at time t after leaving C. $d\eta_c/dt$ can be derived by simple differentiation with respect to t.

But from Llewellyn's equation (3.5), when there is complete space charge.

$$V(t) \propto I_0 t^4 \quad \ldots \ldots \ldots (14.1).$$

Hence (13.1) is to be replaced by

$$d\eta_c/dt = 4t^3/T_c^4 \quad \ldots \ldots \ldots (14.2);$$

X_2 is to be replaced in (13.10), (13.11) by X_3; and (13.12) by

$$2\pi|H_c(j\omega)|^2 \approx \theta^2/25 \quad \ldots \ldots (14.3).$$

We do not pretend that this argument proves anything more than that the numerical coefficient in (13.12) may be too large; and that experimental proof that it is actually somewhat smaller would not force us to abandon any of our main conceptions or to revise very greatly any of our assumptions. In view of the succeeding section, it is important to ask whether some other minor modification would lead to a numerical coefficient greater than that of (13.12).

If X_1 could be substituted for X_2, we should get in place of (13.12).

$$2\pi|H_c(j\omega)|^2 \approx \theta^2/9 \quad \ldots \ldots \ldots (14.4)$$

That substitution would result from an assumption that the introduction of the factor Γ accounts *completely* for the effect of space charge on the extra electron; and that, having introduced it, we may suppose that the extra electron, not only induces as if it were in a vacuum, but also moves as it it were in a vacuum. But, if that were so, the extra electron would be out of step with the main body of electrons that produces the deficiency charge associated with it. If its e/m were the same as that of ordinary electrons, its transit time would be less and it would gain continually on the space-charge electrons; if the transit time were artificially made the same by reducing e/m, it would be in front of the space-charge near the cathode and behind it at the grid. Such a supposition would strike at the root of our physical conception of Γ, and would require a profound alteration of the theory presented here. Without such an alteration we have not found any plausible way of arriving at a numerical coefficient greater than the 1/16 of (13.12).

15. Comparison of Formulae

A formula for the "induced grid noise" has been given by Bakker[5], and also by Ferris and North[6]; we must inquire whether this formula agrees with the result of the preceding sections.

The formula applies when the noise is measured in the grid circuit, the anode circuit being of zero impedance; i.e., when, in the equations of Section 11, $\rho_{13',12}$ is replaced by $\rho_{12,12}$ and $\rho_{13,13} = 0$. Further it is assumed that the amplifier is sensitive only to the narrow band $\Delta\omega$ about ω_0. (11.5) in these circumstances becomes

$$v_\sigma{}^2 = 2I_0\epsilon\Gamma^2|H_c(j\omega_0)\rho_{12,12}|^2 \quad \ldots (15.1)$$

Again, in the application of the formula, it is assumed that, in accordance with usual practice, the noise is expressed in terms of I_d, the steady current flowing in a saturated diode that produces the same noise. For such a diode, $2\pi|H_c(j\omega_0)|^2$ and Γ^2 are both 1; so the noise voltage it produces is given by

$$v'^2 = 2I_0\epsilon .\ 1/2\pi .|\rho_{12,12}|^2 \quad \ldots \quad \ldots (15.2)$$

and $v'^2 = v_\sigma{}^2$ implies

$$I_d = I_0\epsilon\Gamma^2.|2\pi H_c(j\omega_0)|^2 \quad \ldots \quad \ldots (15.3)$$

Now, according to North[4], a good approximation to Γ^2 is given by

$$\Gamma^2 = 6\,(1 - \pi/4).k\theta_c/\epsilon.g_{20}/I_0 \quad \ldots (15.4),$$

where θ_c is the cathode temperature. Hence, to this approximation, since $6(1 - \pi/4) = 1.288$,

$$I_d = 2\pi|H_c(j\omega_0)|^2 g_{20}.0.322 \times 4k\theta_c/\epsilon \ (15.5).$$

On the other hand, by combining Bakker's equations (2), (19), we get (in our notation)

$$I_d = \theta^2/9.g_{20}.0.322 \times 4k\theta_c/\epsilon \quad \ldots (15.6)$$

(15.5), (15.6) agree only if the unplausible (14.4) is true, not if the more plausible (13.12) or (14.3) is true.

On the other hand, our results agree with a proposition used by Bakker in comparing his formula with experiment, namely that the part of the input conductance that depends on the electron current is $g_{20}.\theta^2/20$. For the input conductance is the real part of the input admittance $Y_{22}i$; (9.3), (6.1) lead to Bakker's proposition, if all but the lowest power of θ are ignored. Again we agree with Bakker's formula (33) for the octet that he discusses**, so long as it is assumed (not very plausibly) that there is no space charge between the grids 3 and 4. The formula is obtained easily by the method of Sections 12, 13, using the equations for no space charge throughout.

16. Conclusions

Bakker finds that his formula (15.6) agrees well with his measurements, the observed values being slightly greater than those calculated. If he had used a formula obtained by substituting from (13.12) or (14.3) in 15.5), the discrepancy would have been much greater and probably well outside the range of experimental error. The agreement that he finds is therefore disconcerting, and suggests doubts whether a theory based on the Benham-Llewellyn approximations is applicable to his experiments.††

It would not be surprising if it were not; for the electrodes in the valve he used (EF50) diverge widely from plane parallelism; the cathode is a circular cylinder, and the grid an elliptical cylinder surrounding it.

There is some slight evidence that the theory is not applicable. One test of the assumptions is the validity of the formulae for the first-order effects discussed in Section 9. It requires a knowledge of the cathode-grid capacitance C_c. The construction of the EF50 makes it difficult to estimate C_c in this valve; Bakker and his associates, though they discuss this valve fully, give no value for it. However it is worth recording that the best estimate that we can make of C_c combined with measurements of the effect of the electron current on the input capacitance, suggests that C'/C_c is appre-

** Subject to the substitution in that formula of $(s^6 + s^3/2)$ for $(s^6 + s^3/2)$ as the factor of $\sqrt{s^2 - 1}$.

†† The arguments by which Bakker derives (15.6) are so condensed and employ conceptions so different from those that we regard as significant that we are unable to determine how, if at all, his physical assumptions differ from ours.

ciably greater than that given by (9.4).

A much more satisfactory test of the theory would be provided by experiments on valves, such as the CV273, which approximate much more closely to plane parallel triodes. One of us (E.G.J.) with B. L. Humphreys had measured the input capacitance and the mutual conductance of such valves in various conditions. The results of their work will be reported fully in due course. Here it suffices to say that, though the two estimates of T_c, based on (6.5) and (6.6), agree sufficiently, C'/C_c appears again to be definitely greater than the value given by (9.4). Since (9.4) seems to involve no doubtful assumption other than those of Benham-Llewellyn, it must remain very doubtful at present whether these assumptions are sufficiently true of any valve of practical interest.

In conclusion, the authors desire to tender their acknowledgments to the M.O. Valve Co. Ltd., on whose behalf the work described in this publication was carried out.

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PHYSICAL SOCIETY'S EXHIBITION

Scientific Instruments and Apparatus

HELD at Imperial College, London, from 6th to 9th April, the thirty-second Physical Society's Exhibition was noteworthy for a successful attempt by the organizers to check the tendency shown in recent years for it to develop into a radio exhibition. It has been done by stressing the scientific side rather than the engineering, and it has resulted in the exclusion of much apparatus of the "test-gear" type.

The interests of the engineer and the physicist are now so interwoven, however, that in its new form the exhibition was still of outstanding interest to engineers, and the great number of electronic devices and applications outside the field of communications was perhaps the most striking feature of the exhibition as a whole. It was divided into the customary Trade and Research sections, but the distinction was less clearly drawn than usual.

Meters

Although valveless instruments continue to be shown by many firms, their design is by now fairly stable, and major advances are hardly to be expected. Electronic meters, on the other hand, seemed to be more numerous and varied than ever. This was particularly so with regard to valve voltmeters; here progress was evident in extended ranges of voltage at both extremes, greater frequency range, greater stability and absence of zero drift, and versatility. In most types the indicator is connected between the cathodes of a pair of cathode followers. Whereas a few hundred volts was until recently the top limit of measurement, the Rediffusion Model M.36 reads up to no less than 15 kV. Three alternative heads are provided; a probe for the lowest voltages, and capacitance potential dividers for the higher ranges. It is intended primarily for work on r.f. heaters. The potential divider in the Mullard instrument, scaled in a number of ranges from 3 V to 10 kV, is resistive and presents a load of 100 MΩ on the top range. The same meter has six decade resistance ranges, from 20 Ω to 2 MΩ mid-scale reading.

Mullard valve volt-ohmmeter reading up to 10 kV.

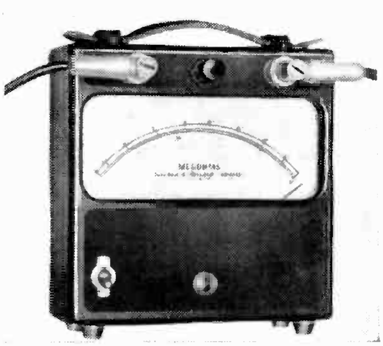
The Advance valve voltmeter is unusual in being driven by raw a.c., with a saving of several components. Like a number of others, such as those by Electronic Instruments, Sifam and J. Langham Thompson, it measures zero-frequency voltages.

The miniature high-gain z.f. amplifier incorporated in the Elliott model enables a full-scale sensitivity of 1 mV to be provided, together with resistance ranges from $10^{-4}\Omega$ to $10^{12}\Omega$. The Kodak instrument has four ranges, from 20 mV to 10 V full-scale,

"Micovac" electronic test meter, with probe, for use up to 200 Mc/s (Electronic Instruments).

all with the notably high input impedance of 200 MΩ. An exceptional feature of the Furzehill Type 378B, with similar ranges plus a fifth reading to 100 V is that a germanium rectifier takes the place of the usual diode. The frequency is limited to 250 kc/s in the model first shown last year, but a more recent model is rated up to 1 Mc/s. Electronic Instruments Model 26, with the conventional diode, is useful from 20 c/s to 300 Mc/s, and is now scaled for a.c. and d.c. up to 300 V.

Among the more versatile instruments known as electronic test meters, a new example is the "Micovac," which is notable for being battery-driven and therefore self-contained. The d.c., a.c. and r.f. ranges (up to 200 Mc/s) are very numerous.



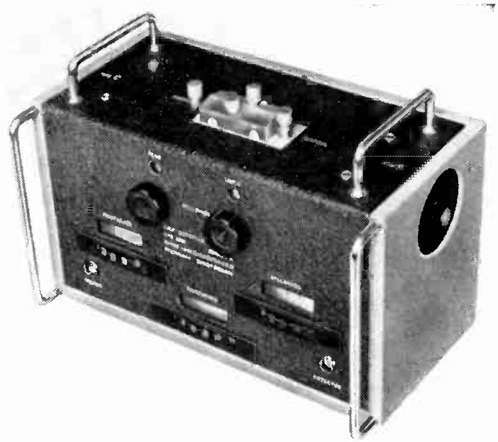
Sangamo-Weston S.118 megohmmeter.

The tendency for ohmmeters, too, is to enlist electronic aid in order to cover a very wide range of resistance. In the Sangamo Weston S.118, $10^{1}\Omega$ to $10^{11}\Omega$ is covered in one range, which is very closely logarithmic. The resistance to be measured is connected between grid and anode of a triode, the indicator being in the cathode circuit. A guard circuit is provided to exclude the effects of irrelevant leakage. This instrument imposes a low testing voltage, $\ngtr 50$ V (which is also readable on the scale), but the Avo and Taylor insulation testers, with similar ranges, apply 500 V. The top reading in the Electronic Instruments megohmmeter is as high as 2×10^{7} MΩ, and the ranges cover seven decades.

In the category of electronic meters may perhaps also be included the Sullivan logarithmic bridge detector. The log law is obtained by a metal-rectifier shunt, and the input range is 6 μV to 75 V at 40 c/s to 20 kc/s.

Standards and Bridges

Very few new standards of *L*, *C* and *R* were to be seen, as the tendency is towards specialized self-contained bridges, but those which were exhibited reflected the attention devoted to very high frequencies. An example was the Labgear substandard precision variable capacitor, with micrometer adjustment and coaxial connection, range 5–25 pF. Temperature errors in standards, resonant circuits, etc., can be counteracted by an improved version of the Sullivan-Griffiths temperature compensator which was first described



Direct-reading bridge, Type 801, for frequency range 1 to 100 Mc/s (Wayne Kerr).

in *Wireless Engineer* in the April 1942 issue. It can be adjusted to give either positive or negative compensation, and is cyclic within 1 in 10^{6} per °C. The mean capacitance is 2 pF.

Bridges tend not only to be more specialized and self-contained but to be direct-reading and to work at higher frequencies. Thus a number of r.f. bridges were shown, including those by Wayne Kerr and Pye in the Trade section and Cossor in the Research section. The Wayne Kerr Model B.801, manufactured under licence from the B.B.C., employs the C.G. Mayo tapped-transformer ratio arms, and measures the parallel susceptance and conductance of the unknown over a frequency range of 1 to 100 Mc/s.

An example of the substitution resonance method of r.f. measurement was shown in the form of a unit by the makers of Avometers, and is one of a series that the firm is producing in this class.

The G.E.C. Research Laboratories demonstrated a semi-bridge comparison version of the resonance method of impedance measurement for frequencies up to 500 Mc/s. The r.f. potentials across two similar tuned circuits, one shunted by a standard resistor and the other by the unknown, are equalized by adjustment of exponential calibrated capacitors in series with the signal source; the wide range of 1 to 10^5 Ω is measurable.

For centimetre waves, waveguide benches are usual for impedance and frequency measurement, and some examples were shown in the Research section and by Plessey.

Frequency standards were less prominent than usual. The Plessey TD.2483 is a heterodyne wavemeter in which ambiguity due to harmonics is avoided by using a pure waveform of uniform amplitude, harmonics being not more than 2 per cent and therefore readily distinguishable.

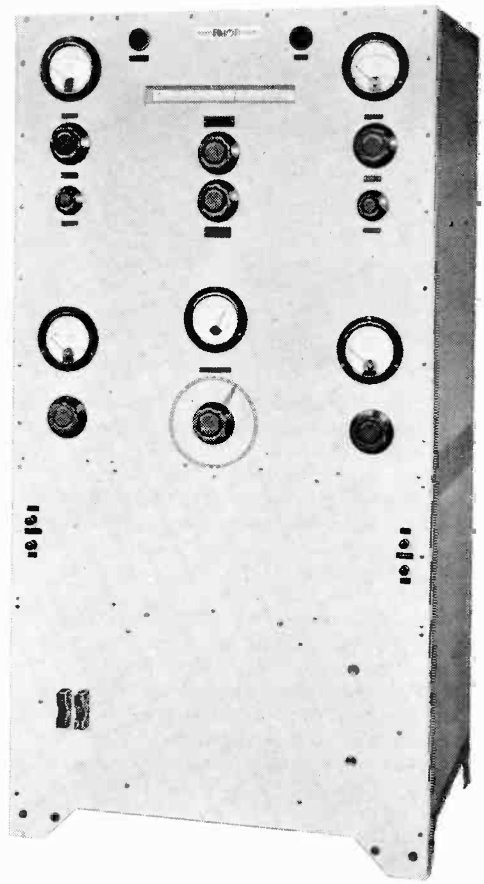
Oscillators and Power Supplies

The range of frequency covered by the Dawe RC oscillators has now been extended to 0.1 c/s–5 Mc/s. Sullivan have entered the RC-oscillator field with both continuous and step variation, from 40 c/s to 100 or 125 kc/s. An RC oscillator is also used for the drive in a high-stability power-frequency generator by Elliott, mainly for precision meter testing. The frequency range is 40–2,500 c/s, and voltages up to 480 V and currents up to 50 A are available, constant to within a few parts in 10^4 over 5-minute periods.

Special power sources for energizing bridges were shown by many firms and are usually characterized by fixed frequency and low harmonic content. An example is the Sullivan Fixed-Frequency Oscillator.

Considerable interest is now shown in stabilized power supplies, whether incorporated in equipment or available as separate units for general use. Among the latter were those by Metrovick and Furzehill; the former providing up to 0.25 A at 650 $\pm$ 10 V, constant within $\pm$ 0.01 per cent for a whole day and $\pm$ 0.001 per cent over short periods. The most popular method employs a series valve controlled by an amplifier which compares the output with a reference voltage. Among the gas-filled regulator tubes used as the voltage standard, the Mullard 85A1, with 85.5-V drop, is claimed to be free from sudden shifts of glow, and hence of voltage, with varying current. The Mazda 12E1 tetrode has been specially designed as a series valve for voltage-stabilized units, and is rated for an anode dissipation of 35 W, the maximum conditions being anode voltage 700, or cathode current 300 mA.

An outstanding development which is basically a stabilized power supply, but of such refinement that it is also a precise standard of voltage, is made by Tinsley. By substituting a standard cell for the gas-discharge reference tube, and a galvanometer and photo-cell for the amplifying control valve, with an accurately calibrated potential divider for comparing the output voltage with the standard cell, the selected output voltage is not only exceptionally stable but is known to 1 in 10^4, enabling measurements and calibrations to this order of accuracy to be made much more easily than with conventional potentiometry. The same firm also exhibited a source of highly stabilized alternating current.



Elliott high-stability 40-2,500 c/s current and voltage test generator.

An interesting power-supply device with quite a different application is the Westinghouse "Westeht" voltage multiplier, enabling the 6 kV or so required for a television tube to be obtained from the ordinary 350–0–350 V receiver transformer. The new type of metal rectifier incorporated is also suitable for the 25 kV, or even 50 kV, needed for projection tubes.

Cathode-Ray Tube Equipment

Oscilloscopes, as such, were almost entirely absent this year, though they were very much in evidence for demonstrations. The Southern Instruments Type ME15 is an example of equipment produced to meet special requirements. It is arranged for taking photographic records, either single pictures of the screen as seen by eye, or continuous strips of waveforms.

Furzehill type 1931 *stabilized power unit, with provision for metering the several positive and negative supplies.*

itself a sufficient criterion; and the equipment now being developed is for showing the shape of the reverberation decay curve, with the intensity and time delay of any distinctive echoes, when the room is pulse-excited at any desired audio frequency.

Apparatus for the rapid tracing of a.f. characteristics was demonstrated by Rediffusion. A feature of this equipment is that no synchronization is needed between the variable-frequency oscillator and the c.r. curve tracer, which can be at widely separate points. The Y-deflection is, as usual, proportional to signal output; the X-deflection is developed from the output signal by a frequency discriminator, and is therefore independent of the speed at which the oscillator frequency control is driven. After the curve has been traced on the long after-glow screen, vertical and horizontal graph lines are superimposed by the beam, to provide quantitative scales of c/s and db. The range and spacing of the db-scale can be adjusted as required, and extra frequency lines can be inserted to mark special features of the curve.

An interesting demonstration in the Research section by D. M. MacKay of Kings College, London, was a method of displaying three-dimensional data on the c.r. tube. Unlike the methods shown at a meeting of the I.E.E. on 24th March last, which used magslips for producing the perspective picture of a solid figure, Mr. MacKay's employs potentiometers. The figure can be rotated by the controls about any of the three axes, say for examining the relationships between two of the variables, or for obtaining a perspective impression of all three. The example demonstrated was a display of the I_a–V_a–V_s characteristic surface of a tetrode showing the development of a secondary-emission "valley" at certain adjustments. A system of this kind would be of immense value for instructional purposes.

The only evidence of radar in its original rôle was a model of the Sperry equipment now being installed for the Mersey Docks and Harbour Board, but the same principle could be seen on the G.P.O. Research stand applied to the location of line faults. Any discontinuity in impedance, such as would be caused by a complete or partial open or short circuit, gives rise to an echo, from which the distance of the fault can be read off in the usual way from a c.r. tube display.

Another Research exhibit, by Standard Telephones and Cables, demonstrated the acoustic analogue of radar applied to the investigation of the characteristics of auditoria and studios. Experience has shown that the reverberation time is not in

Another Research demonstration by Mr. MacKay, using c.r. tubes in two roles, showed how derivatives of mathematical functions known only in graphical form can be generated at high speed for use in electronic analogue computors. A mask cut to the shape of the known curve is placed on the screen of one c.r. tube, which has an X-sweep. The Y-deflection is controlled by a photo-cell in such a way that the beam follows the outline. The deflection voltage can then be integrated, etc., as required.

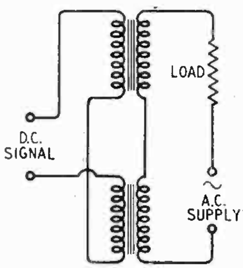
Quartz-crystal oscillator for high-quality f.m. transmission.

Miscellaneous Apparatus

Magnification curves of ferrous materials have often been taken on c.r. oscillographs, but Metrovick give reasons for developing a servo-driven recorder for tracing the curve out comparatively slowly on graph paper, and the apparatus was demonstrated on their Research stand. No photography is required, more accurate curves are traced, the effects of eddy currents are avoided and the initial magnetization curve can be plotted.

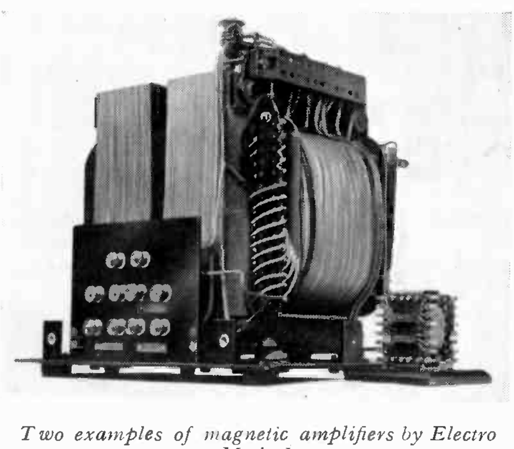
Azimuth-radiation characteristics of centimetre-wave aerials can also be traced either by c.r. apparatus or by a mechanical recorder. In the Marconi equipment the former is used for initial adjustment of the radiator, and the latter for making a permanent record. Servo recording equipment was demonstrated also on the Cossor Research stand.

Another Marconi exhibit is to play an important part in the 25 kW f.m. transmitter now being made for the B.B.C. Difficulty has been experienced in frequency-modulating a quartz-controlled master oscillator linearly over the standard 75-kc/s deviation for high-quality broadcasting. Unwanted modes of oscillation are eliminated in the Marconi system by a device analogous to a waveguide. A square frame mask is used on the quartz plate, and the unwanted frequencies are absorbed in a load while the wanted ones are reflected into the central area. The demonstration showed almost perfect linearity over a deviation of ± 1 in 1000.



Principle of d.c. magnetic amplifier.

The magnetic amplifier has until recently been little known in this country, but interest has been stimulated by the release of information from Germany. The principle was demonstrated on the Elliott Research stand. Basically it is the control of an alternating current by the direct current to be amplified; the latter reduces the permeability of a magnetic core and thereby reduces the reactance of an a.c. coil wound on it. To prevent the d.c. closed circuit from loading the a.c. circuit, two reactors are connected as shown in the diagram.



Two examples of magnetic amplifiers by Electro Methods.

Obviously the performance depends largely on the core material. Mumetal is often used, but another Telcon alloy, H.C.R. metal, which has an almost rectangular hysteresis loop, is offered specially for this application. Complete magnetic amplifiers were shown by several exhibitors, including Electro Methods, Everett Edgcumbe, and Ferranti. Whereas valve amplifiers are particularly suitable for high-impedance systems and high frequencies, magnetic amplifiers are applicable principally to low-impedance and d.c. systems. Power gains of 10^4 or more are obtainable in one stage of magnetic amplification, and stages can be cascaded.

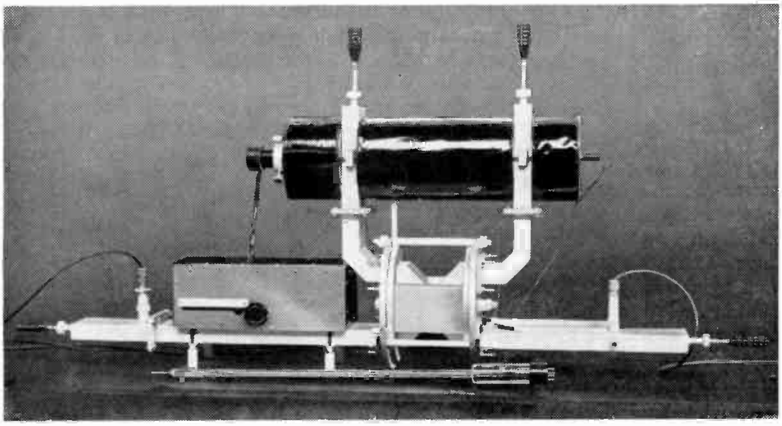
R.f. heaters were shown by Radio Heaters Ltd., who had two new developments. One is a rapid moisture tester, in which a sample can be weighed continuously while being rapidly dried by r.f. energy. Tests formerly taking many hours can thus be performed in a few minutes. The other is a continuous r.f. oven, for materials, such as p.v.c. chips, where the time factor in a heating process is important. Although complete heaters were not conspicuous, many firms exhibited valves, components and measuring equipment particularly suitable for this application.

Somewhat remote from practical technology of this kind was a National Physical Laboratory exhibit of apparatus for measuring the velocity of propagation of electromagnetic waves. So far from requiring a vast base on which to measure this greatest of all velocities, the apparatus consists of a cylindrical resonant cavity only 8-cm long. A signal at about 3,300 Mc/s is introduced, and its frequency exactly adjusted until resonance is shown on an indicator. The velocity is deduced from the frequency and the dimensions of the cavity, which are measured with the utmost precision, and correction is made for finite conductivity, as described in *Nature*, 1947, Vol. 159, p.611. The velocity in a vacuum is given as 299,793 $\pm$ 9 km/s.

E 1872 transmitting valve; anode dissipation up to 150 kW, frequency up to 30 Mc/s (Marconi-Osram).

Valves

Not very many new types were shown, and those for use in voltage stabilizers have already been mentioned. Sub-miniature amplifying valves for hearing aids were on several stands. At the other extreme was the Marconi-Osram E1872 with an anode dissipation of 150 kW at 30 Mc/s, or at a reduced rating up to 50 Mc/s The E1756 is a coaxial-line triode rated at 5 kW up to 100 Mc/s, and the E1862 an earthed-grid triode, 1 kW up to 600 Mc/s. The Ferranti xenon-filled high-voltage rectifiers are claimed to be superior to mercury-vapour types, as the peak-inverse voltage is practically unaffected by ambient temperatures much above those for which mercury rectifiers are usually rated. The Mazda 27M1 is a nine-stage photo-electric multiplier with a sensitivity of 10 A per lumen.



Standard Telephones demonstration of travelling-wave amplifier having 1,000-Mc/s bandwidth at 10,000 Mc/s. The tube is shown lying in front of the apparatus.

Equipment for giving direct readings of equivalent noise resistance of amplifier and frequency-changer valves was demonstrated by Mullard. The noise is amplified over a known band around 100 kc/s, but a 1-kc/s pilot signal is applied through the chain for operating an a.g.c. system which stabilizes the gain of the amplifier.

The travelling-wave tube, for amplifying centimetre waves, has not reached a commercial stage, but there were G.E.C. and Standard Telephones Research exhibits. The latter demonstrated a gain of 20 db (up to 30 db has been obtained) at 10,000 Mc/s. The noise factor is too high for pre-detector amplification, but at relatively high levels the gain provided is useful owing to the bandwidth, which is of the order of 1,000 Mc/s.

Components

A special requirement for capacitors with very low leakage and power factor, for computors, etc., was the origin of some examples shown by T.C.C. in which plastic film is used instead of paper. The leakage resistance is said to be $10^{12}\Omega$ for 1 μF, and the power factor 0.0002. Dielectric hysteresis is also very low. An E.R.A. Research exhibit demonstrated investigations into imperfections in the silver coating in mica and ceramic capacitors, which give rise to spontaneous fluctuations of capacitance.

The well-known Berco Regavolts are now obtainable with motor drive for automatic or remote control. On a different scale were the precision potentiometers shown by Ferranti. These are wound on a toroidal former to an accuracy of 0.25%, and, by the use of precious metals, contact resistance and operating torque are both made very small. For applications where a precision a.c. power potentiometer is needed, the Muirhead Ipot is appropriate. It is similar in principle to the Variac, but gives a closely linear relationship between potential and angle to within $\pm$ 1 in 2,000 of the maximum.

A number of miniature moving-coil relays were shown, notably by Electro Methods and Sangamo Weston. The latter claim to be able to provide operation on as little as 50 μA.

The "Sentercel" selenium rectifiers of Standard Telephones have been developed in the direction of higher working frequencies, the limit now being 0.5–5 Mc/s, according to the nature of the duty.

Special types of coaxial cables included low-capacitance types by Transradio, with 4 to 7 pF per foot, and Telcon non-microphonic cables for excluding interference caused by flexing or impact.

Materials

Telcon easily-saturable alloys for magnetic amplifiers have already been mentioned. The same firm supplies Permendur, claimed to have the highest saturation induction of any known material. It is applied, for example, in obtaining the exceptionally high field excitation of Voigt loudspeakers. Core materials with a wide range of properties were shown by Transformer Steels Ltd. The Ferroxcube magnetic materials, of which the ceramic manganese zinc ferrite is an example, have the peculiar property of being non-conductors and hence substantially free from eddy-current losses. Some applications demonstrated by Mullard included i.f. transformer cores, providing a higher Q than dust cores for a given size of transformer, and a concentrator for an induction process heater.

Technical advances in the construction of valves and other electronic applicances, and in climate-resisting sealed connections, have been due largely to the development of glasses with suitable temperature characteristics. "Soldering" with glass was illustrated in a research exhibit of B.T-H., and glass-to-metal seals on the Telcon stand.

BOOK REVIEW

Theory and Application of Microwaves

By A. B. BRONWELL and R. E. BEAM. Pp. 470 + vii, with 253 illustrations. McGraw-Hill Publishing Co. Ltd., Aldwych House, London, W.C.2. Price 36s. (in U.K.).

There is an unfortunate tendency for writers of books on new applications of electromagnetic theory to devote far too much space to needless recapitulation of standard and basic work, which is already dealt with admirably in several widely available texts. While there is not a great deal to quarrel with in the actual treatment of the matters included in this book, it does contain much that is superfluous.

The first four chapters deal with a historical introduction, the fundamental laws governing the behaviour of charged particles in electric fields; current, power and energy relationships, and the physical basis of equivalent circuits. Of these, only their introduction—some four pages—need have been included. Chapters 5, 6 and 7, on negative-grid triode oscillators and amplifiers, transit-time oscillators (including the klystron) and magnetron oscillators, are appropriate to the subject of microwaves. On the whole these chapters are quite well done and give a clear physical presentation of the phenomena involved, but it is surprising to find not even a brief reference to the pioneer work of Barkhauzen and Kürz in the section on positive-grid oscillators in Chapter 6.

We then come to Chapters 8, 9 and 10, which cover transmission-line equations, the graphical solution of problems associated with transmission lines, and matching systems, and many readers will again feel that they are covering very well-worn ground. In Chapter 10 there are some references to impedance and power measurement at microwave frequencies, and this is one place where the authors could profitably have expanded the treatment.

Chapter 11, on transmitting and receiving systems, contains a quite inadequate section on propagation characteristics. The range of usefulness of microwaves for communication, or other purposes, by radiation through the atmosphere is fundamentally dependent upon the manner of propagation of these waves. The matter is here dismissed, however, in two or three short paragraphs. The authors, incidentally, say that the troposphere extends from the earth's surface up to the ionosphere, and would thus appear to be unaware of the existence of the stratosphere. This section also contains an incorrect treatment of behaviour of a half-wavelength dipole: it is not always necessary to dispense with rigour—as has occurred here—in order to give a clear exposition of a particular problem. Thus, it may be shown that equivalent absorbing area of an ideal short doublet antenna is $3\lambda^2/8\pi$; i.e., the doublet transfers to a matched load the energy crossing such an area. It is, therefore, incorrect to state that this is also the area appropriate to a half-wavelength dipole which, it is well known, has a power gain of about 1.09 relative to a doublet. The rest of this chapter is concerned with various aspects of transmitting and receiving systems, and includes such topics as amplitude, phase and frequency modulation, and also a short section on the automatic frequency control of microwave oscillators. A few matters relating to pulsed systems, particularly radar systems, are discussed in Chapter 12, which is quite short, being only eight pages in length.

Chapters 13, 14 and 15 treat in some detail Maxwell's Equations, the propagation and reflection of plane waves, and the solution of electromagnetic field problems. Chapter 16 begins with a physical presentation of the transmission of waves by successive reflections at the walls of a guide; then there follows a treatment on the basis of the formal solution of Maxwell's Equations, with the appropriate boundary conditions, leading to a description of the field structure of the various possible modes of transmission in rectangular- and circular-section waveguides. Attenuation arising from losses both in the dielectric medium and the walls of the guide is discussed.

The last five chapters, 17 to 21, deal with the applications of waveguides and resonators, linear antennae and arrays, the impedance of antennae and finally other radiating systems such as the biconical antenna and horn radiators. There is a bad error on page 427 of Chapter 20 where the field of a linear antenna is derived. Fig. 2 on this page shows what is supposed to be the current distribution of a centre-fed linear aerial: the distribution given is incorrect, for it must be symmetrical about the centre point, and not asymmetrical as shown. As described here, such an aerial would have no radiation in the equatorial plane whereas in fact there will be appreciable radiation: for an aerial about one wavelength long, such as that illustrated, the radiation is a maximum in the equatorial plane. There is radiation in this plane for an aerial an odd number of wavelengths long, and no radiation for aerials an even number of wavelengths long. One can also quickly see the inconsistency if one considers the state of affairs in, say, a parallel-wire transmission line feeding such an antenna, for, unless the current distribution in the antenna is symmetrical about its centre, it will be found that the currents in the two wires of the transmission line are in phase, instead of being out of phase, as is well known to be the case. The current distribution shown is that appropriate to an end-fed antenna.

There are three appendices, one on systems of units, one giving the electrical properties of various selected materials, and another containing a useful summary of some of the commoner formulae encountered in vector analysis.

The book is excellently produced and liberally illustrated. The authors have justified their claim that, "throughout the engineering point of view has been stressed and that, wherever possible the analytical results have been expressed in a form convenient for engineering use." A good feature is the number of illustrative examples included in the text: there is also a wide selection of problems suitable for students. This is not an elementary text and it is, therefore, a pity that a better balance was not maintained in the selection of the subject matter, for the class of reader to whom it would appear to be addressed will almost certainly have at his disposal other books containing much of the well-established basic material. J. A. S.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Producing High-Frequency Pulses

SIR,—The conventional method of producing sharp pulses by differentiating a rectangular wave is quite satisfactory at frequencies below 100 kc/s. However, at higher frequencies, particularly when a continuous frequency spectrum is required, considerable difficulty is experienced in designing the various amplifier and clipping stages. These stages must necessarily be resistance-capacitance coupled and designed on video-amplifier lines.

Several alternative methods of producing pulses have been mentioned in the literature, one of which uses the class C amplifier[1]. This method, although apparently not utilized, may be conveniently adapted to the production of very high-frequency pulses.

The class C amplifier is used in a method described below and enables sharp pulses having a recurrence frequency of several megacycles per second to be obtained. There are two main requirements of a pulse; first, the ratio of pulse duration to pulse period must be small and secondly, the amplitude must be sufficiently large. In many applications, particularly in trigger circuits, the shape of the pulse is immaterial and need not be exponential as is that produced by differentiating a rectangular wave, but may be a portion of a sinusoidal wave.

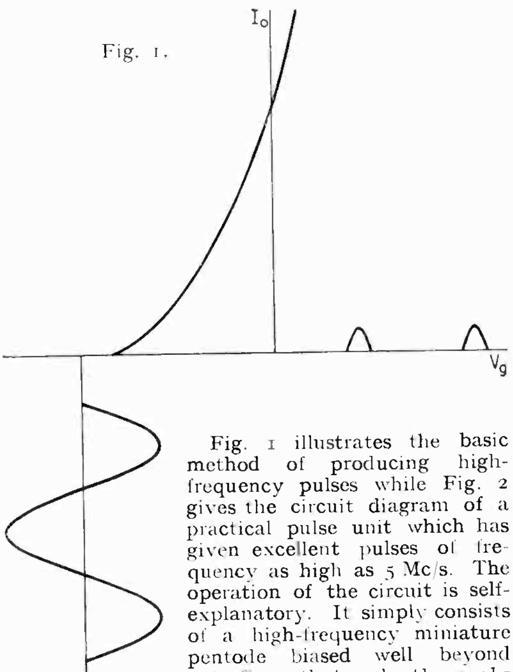


Fig. 1.

Fig. 1 illustrates the basic method of producing high-frequency pulses while Fig. 2 gives the circuit diagram of a practical pulse unit which has given excellent pulses of frequency as high as 5 Mc/s. The operation of the circuit is self-explanatory. It simply consists of a high-frequency miniature pentode biased well beyond cut-off so that only the peaks of the input wave appear as plate current excursions. For frequencies lower than 5 Mc/s, Z_L may be a 2,000-Ω resistor but for higher frequencies a compensated load is essential. The pulse duration and amplitude may be varied by means of R_1 and R_2, their values depending on the amplitude of the input wave.

[1] H. T. Reich. "Theory and Applications of Electron Tubes," p. 359. McGraw-Hill Book Co. Inc., New York. 1944.

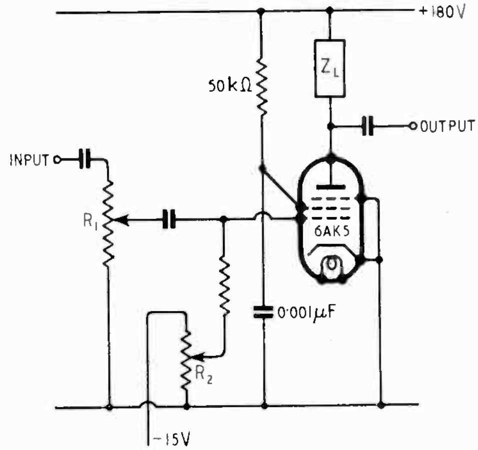


Fig. 2.

Pulses with a recurrence frequency of 5 Mc/s obtained from this unit have been used to operate a counter stage which was designed for high-frequency operation. There is no reason why pulses of higher frequency could not be obtained if the load impedance Z_L is designed accordingly.

R. D. CARMAN.

Munitions Supply Laboratories,
Melbourne, Aust.

Standard Terms and Abbreviations

SIR,—I was interested to read your March Editorial and Mr. Scroggie's letter in the April issue, because for several years now I have been campaigning within my own company for standardization of abbreviations and nomenclature.

I must agree that it is reasonable for "ampere" to be shortened to "amp" both for reasons of common usage and euphony and, as you say, it has a precedent in "farad" and "volt." Certainly, too, "k" ought to be elevated to "K" and *Electronic Engineering* has already adopted this. If your journal and your sister publications *Wireless World* and *Electrical Review* followed suit, I am sure that official récognition would not be far behind.

On a par with the American practice of stating power in watts is that of giving high frequencies in kilocycles (or should I say kilocycles per second to avoid being hoist on my own petard ?) instead of in megacycles; i.e., 21,000 Kc/s band instead of 21 Mc/s. I have not yet seen a reference to "a 10,000,000 Kc/s radar" but, no doubt, that will come.

This question of the "per second" is one which has exercised me considerably. Everybody in the business says "megacycles," "kilocycles," "cycles," omitting the per second and I am not sure that this practice is at all bad. It is certainly common enough in American periodicals to see kc and Mc without the /s. The German unit "hertz" has much to commend it, but I feel that further attempts to introduce it in the English-speaking countries would again meet with failure.

The practice of using capitals for abbreviations—A.C., H.T., U.H.F., etc., seems to be utterly illogical and indefensible and I have for long insisted that the rule must be "what would you use if the words were written in full?" This immediately gives lower case letters, a.c., h.t., u.h.f., etc., which, incidentally are much more pleasant to the eye than a spatter of capitals across the page. The custom of using small capitals adopted by some journals seems a weak compromise which gets nowhere. I cannot agree with the American form, a-c with a hyphen, even when the abbreviation is used adjectivally, because surely the hyphen is not normally used when the words are written in full.

The phrase "a.c. current" would be absurd if one saw it written in full, although perhaps not so absurd as "a.c. voltage," but these terms are so time-honoured that no attempt to alter them would succeed.* Another point which often occurs is the use of "a" or "an" before certain abbreviations using vowels and vowel-sounding consonants. Examples are "a f.m. transmitter" or "an f.m. transmitter." In speech one certainly says "an eff-emm transmitter" but in writing it is surely correct to put "a f.m. transmitter" because it is an abbreviation. Others commonly seen are "an h.t. supply unit," "an r.f. amplifier" and so on. Fowler seems to offer no guide here.

And again, is it db or dB? I have adopted db, although I must agree with colleagues who say that as units derived from personal names such as watts, farads, volts, etc., are given a capital, it should really be "B" for bel and thus "dB" for decibel. You use db in your columns and so do many other people, but I have a sneaking feeling that dB is more correct and logical.

One last point, Mr. Scroggie mentions the American usage of "m" for both mega- and milli-; there is also continued British usage of "m" for micro- when referring to capacitances. This is a technical solecism often committed by condenser manufacturers. Another one is the use of ω for Ω even by people who should know better. But similar examples can be quoted *ad nauseam* and I have a strong suspicion that they will continue while so many engineers persist in a *laisser-faire* policy towards corrections and logic in speech and writing. Messrs. A. Reyrolle are to be congratulated on their move to introduce standardization within their own ranks and it is to be hoped that they will show the path to all of us who have occasion to write on technical matters.

E. D. HART.

St. Albans, Herts. Marconi Instruments Ltd.

*We cannot feel that "a.c. current" and "a.c. voltage" are permissible expressions. To allow them is only to open the door to "i.f. frequency" and a host of others. They can all, quite easily, be avoided.

It seems likely that they have arisen through a feeling, which we share, that it is desirable to use abbreviations only adjectivally. One clearly cannot write "The a. current is . . ." and it is unpleasant to see "The a.c. is . . ." however accustomed one is to it in speaking. Hence, we get "The a.c. current is . . ." instead of the correct "The alternating current is . . ." The correct form is so little longer than the incorrect abbreviated expression that we can see no objection to its use.

Mr Hart's point about the use of "a" and "an" is an important one which has caused us some thought. Surely the correct usage depends on whether the abbreviation is read as an abbreviation or mentally translated. If we read "f.m." as "frequency-modulated" then we should write "a f.m." but if we read it as "eff-emm" then it is necessary to use "an f.m."

While we agree that "K" is to be preferred to "k" as an abbreviation of "kilo" we do not feel that any change should be made without investigation into the reason which made the B.S.I. decide on "k." We cannot help feeling that there must be a good reason for what is apparently so illogical a choice.—ED.

SIR,—I have read, with interest and approval, both your Editorial and Mr. Scroggie's recent letter on the subject of abbreviations.

However, I have noticed in the technical writings in various issues of your journal, reference being made to a symbol—db. Now, it appears from the text that this is intended as an abbreviation for one of the diminutives of the Bel. Surely the correct abbreviation for one-tenth of a Bel should be dB, in spite of the anomaly fathered by BS.661.

If the db is considered *de rigueur,* we shall be expected to assimilate other horrors without protest.

Banstead, Surrey. L. F. ODELL

IGNITION INTERFERENCE

A campaign to persuade owners of motor vehicles to fit ignition-interference suppressors is being launched by the Radio Industry Council in order to improve television reception. Such interference suppression is now being adopted by Associated Iliffe Press and some 70 vehicles are being fitted; these include vans belonging to Cornwall Press Ltd., the principle printers in the Iliffe group.

British Standard Institution Year Book, 1947

Publication Sales Dept., British Standards Institution, 24, Victoria St., London, S.W.1. Price 3s. 6d. (post free).

This publication includes a subject index and a synopsis of the 1,400 British Standards now current.

The Fundamental Research Problems of Telecommunications

Pp. 80 + iv. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 1s. 6d.

This is the report of the Telecommunications Research Committee which was constituted in 1946 as an *ad hoc* committee of the Department of Scientific & Industrial Research. It includes the reports of nine working parties covering Wave and Line Propagation, Valve Fundamentals, Properties of Materials, Contact Phenomena, Circuitry, Luminescence, Photo-emission and Television.

Survey of Existing Information and Data on Radio Noise over the Frequency Range 1-30 Mc/s

By H. A. THOMAS, D.Sc., M.I.E.E., and R. E. BURGESS, B.Sc. Pp. 126 + vi. Department of Scientific & Industrial Research, Radio Research Special Report No. 15, H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 3s.

"The report consists of a critical survey of the available published literature on radio noise of all types whether of atmospheric, cosmic or man-made origin as affecting radio reception in the frequency range 1 to 30 Mc/s. In addition, it includes a summary of all the useful measurements of the intensity of atmospheric noise made in the past and available to the National Physical Laboratory as a result of specific enquiries."

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.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

589 407.—Time-base circuit producing a saw-tooth sweep of adjustable slope, particularly for use in radiolocation.

The British Thomson-Houston Co. Ltd. (communicated by the General Electric Co.). Application date 21st March, 1945.

589 593.—Radiolocation set in which the echo-signal transferred to a remote indicator is controlled by the position of the cursor on the main range-indicating tube, thereby cutting-out parasitic echoes.

Cie. pour la Fabrication des Compteurs et Materiel d'Usines à Gaz. Convention date (France) 8th March, 1940.

589 603.—Method of radiating a scanning-beam for use in radiolocation by varying the phase-velocity of the energy fed through a waveguide to a stationary aerial.

Western Electric Co. Inc. Convention date (U.S.A.) 27th July, 1943.

589 607.—System of space-scanning, for use in radiolocation, wherein waves of different frequency are fed to a plurality of fixed aerials in predetermined phase.

Hazeltine Corporation (assignees of H. M. Lewis). Convention date (U.S.A.) 27th October, 1943.

589 608.—Radiolocation receiver wherein modulation signals of different frequency are applied to control the directional response of a stationary aerial-system to the incoming echo-signals.

Hazeltine Corporation (assignees of H. M. Lewis). Convention date (U.S.A.) 27th October, 1943.

589 609.—Space-scanning system, for use in radiolocation, wherein waves of different frequency are fed to fixed aerials so as to allow both the rate of scanning and the shape of the scanning-beam to be controlled.

Hazeltine Corporation (assignees of H. M. Lewis). Convention date (U.S.A.) 27th October, 1943.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

589 228.—Phase-adjusting means for controlling the incidence of two sets of recurrent signals, say on the separate grids of a "gate" or switching valve.

A. D. Blumlein and E. L. C. White. Application date 13th December, 1939.

589 496.—Balanced circuit for detecting frequency-modulated signals, in which an upper and lower standard or "reference" frequency is provided by separate crystal-oscillators.

Philips Lamps Ltd. Convention date (U.S.A.) 27th October, 1943.

589 845.—Push-pull amplifier in which an unbalanced input is converted into two voltages of equal magnitude and opposite phase.

The General Electric Co. Ltd. and D. C. Espley. Application date 26th March, 1945.

589 867.—Receiver of the kind in which the tuning is swept constantly over a given frequency-band until a signal of worth-while strength is received, whereupon the tuning is automatically locked.

W. G. Johnston and C. S. Wright. Application date 14th September, 1944.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

589 345.—Construction and arrangement of an optical filter or mask, suitable for producing coloured effects in television.

The General Electric Co., Ltd., and L. C. Jesty. Application Date 12th June, 1944.

589 439.—Saw-tooth oscillation-generator for television, combined with a synchronizing-pulse amplifier, in the form of a triode-pentode or triode-hexode.

The Mullard Radio Valve Co., Ltd., and A. M. Spooner. Application date 5th September, 1944.

TRANSMITTING CIRCUITS AND APPARATUS

(*See also under Television*)

589 126.—Generating short pulses of high peak-power from low-rated valves fed with an alternating anode voltage which is a sub-multiple of the repetition-frequency.

Standard Telephones and Cables Ltd. and M. M. Levy. Application date 27th May, 1941.

589 303.—Variable-reactance device, comprising a thermionic valve and a phase-shifting circuit, suitable for frequency-modulation.

Marconi's W. T. Co., Ltd., (assignees of H. O. Peterson). Convention date (U.S.A.) 1st December, 1943.

589 659.—Modulating system in which the tuning of a klystron oscillator is controlled by a capacitor of the piezo-electric type.

Western Electric Co., Inc. Convention date (U.S.A.) 21st October, 1943.

SUBSIDIARY APPARATUS AND MATERIALS

587 399.—Coupling and voltage-biasing arrangements for stabilizing the frequency of a Rochelle-salt crystal oscillator.

"Patelhold" Patentverwertungs etc. A. G. Convention date (Switzerland) 12th April, 1943.

588 282.—Shaping-circuit for deriving two steep-fronted voltage-waves, in phase-opposition, from an a.c. supply, say for welding operations.

Radio Gramophone Development Co. Ltd. and E. T. Court. Application date 7th February, 1945.