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EDITORIAL

On the Use of Standard Terms and Abbreviations

WE were interested to receive a copy of a booklet of 51 pages entitled "A Few Notes on Words and Other Things" published by Messrs. A. Reyrolle and Co. Ltd., the switchgear manufacturers of Hebburn-on-Tyne. After three chapters on the English language by Mr. Thomas Carter, M.I.E.E., a member of the staff to whom, we suspect, the booklet owes its origin, there is a reprint of his recent article on Ohm's law in *Electrical Review*. This is followed by an article by E. H. W. Banner, M.Sc., on "Electrical Terminology," from *Electrical Review* of 2 June 1944, a humorous article by J. H. Quick from the *I.E.E. Students' Quarterly Journal* and sections dealing with hyphens and abbreviations, and what words to use and what to avoid. The last dozen pages are devoted to definitions of the electrical terms most likely to be used in Reyrolle's Works, based generally on the Glossary of Terms used in Electrical Engineering, B.S. 205 : 1943, but giving their own preference in the case of alternatives, and introducing some terms that are not in the Glossary. It is very interesting to know that one of our leading electrical manufacturing firms is making such a praiseworthy effort to bring home to the members of its staff the importance of the correct use of our language and the fact that "there are rules for the construction of phrases, clauses, sentences and paragraphs . . . just as there are rules for the construction of switchgear."

In his recent inaugural address as President of the Institution of Electrical Engineers, Mr. P. Good of the British Standards Institution, said, "It can be asserted with confidence, I believe, that at the conclusion of the chain of transactions from raw material to ultimate consumer there is a loss in efficiency, as well as in commercial integrity, due to loose terminology."

The following quotation from Jean-Louis Guez de Balzac (1597–1664): "Ce n'est pas assez de savoir la Théologie pour écrire de la Théologie ; il faut encore savoir écrire, qui est une seconde science" might be mistranslated as follows: it is not enough to know all about wireless engineering in order to write about wireless engineering ; it is also necessary to know how to write, which is quite another matter. Mr. Carter says, "I think that we can only acquire this by study of the masters—prose if you shudder at poetry ; but better both." We are prompted to refer to this by the fact that, although many of the articles submitted to us for publication show signs of considerable care, others suggest that the authors have little knowledge of English composition or interest in the proper construction of sentences and use of words.

We need not discuss the article on Ohm's law as we have referred to it in two recent editorials, but we would draw attention to Mr. Banner's article on the importance of using standard terminology and avoiding obsolete symbols and expressions. There are bound to be some differences of opinion as to the most desirable terms and symbols, but after representative committees of the British Standards Institution have spent months discussing all the pros and cons and have finally agreed on certain terms and symbols, which are then published by the B.S.I., it is surely the duty of teachers and authors to adopt these recommendations, even if it entails a little readjustment of their previous habits. Probably the most ill-treated letter of the alphabet is "μ," for one comes across ma instead of mA for milliamperes, mfd instead of μF for microfarads, and mc instead of Mc/s for megacycles per second.

It is unfortunate that there have been a few misprints in the B.S.I. publications. In the 1926 glossary the definition of the dyne referred to an acceleration of one centimetre per second; this was due to an over-zealous printer who, seeing 'per second per second,' corrected such an obvious mistake at the eleventh hour. Then in the 1936 glossary it was intended to use a capital script $\mathcal{E}$ as the symbol for electric force, to distinguish it from E which was used for electromotive force, but unfortunately it appeared in the printed edition as a small Greek epsilon ϵ, which was the cause of much trouble. Magnetic susceptibility suffered a worse fate, for before going to the printer it fell into the hands of a non-technical editor who, presumably because capacity had been replaced by capacitance, altered susceptibility to susceptance and then, to emphasize still further the truth of the old adage about a little knowledge, altered the symbol from κ to b, so that the definition read as follows: "Susceptance—The ratio of the intensity of magnetisation to the magnetising force. Symbol: b (or B); $4\pi b = \mu - \mu_0$." This almost incredible muddle was, of course, corrected in the 1943 edition of the glossary, which gives correct definitions and symbols for both susceptibility and susceptance. There were certain points that were left in an unsatisfactory state in the current (1943) glossary; for example, the abbreviation for kilowatt-hour was given as kWh, and that for reactive volt-ampere hour as VARH in one place and Varh in another. The Reyrolle booklet gives VAR and VAh but shies at the four initials and gives VAR-hour-meter. It must be remembered that the current edition of the glossary was produced in 1943 when conditions were far from favourable, and any such slight defects should not be made an excuse for not adopting its recommendations.

The sections of the Reyrolle booklet dealing with hyphens and abbreviations and with words are based largely on Fowler's "Modern English Usage." Fowler says: The chaos prevailing among writers or printers or both regarding the use of hyphens is discreditable to English education." A walking stick would be a wonder, but a walking-stick is a common object; similarly a red hot poker is a very different thing from a red-hot poker. The section on words is mainly a plea for the use of simple normal words and the avoidance of stylish words and genteelisms.

One of the difficulties in following the recommendations of the B.S.I. is that of keeping up to date. In the Reyrolle booklet, for example, the recommended abbreviations are lb., oz., cm., h., etc., all in accordance with B.S.560: 1934. Now these are all obsolete, for an amendment was published in September 1945 deleting the unnecessary dots after all these abbreviations, and recommending lb, oz, cm, h, ft/min, cm/sec, m/sec, radn/sec, etc, all without any dots and without any "s" in the plural.

When the name of Ampère was first suggested for the unit of current Heaviside objected on the grounds that it would soon be shortened to amp, which he thought was horrible. With a twinkle in his eye he suggested shortening it to père, since Ampère was the father of electrical science. His fears were well founded, for in factory slang the unit of current soon became the amp, and this appeared as a recognized abbreviation in some editions of the B.S.I. Glossary; although it has been banished from the more recent editions and its place taken by A, one still comes across it, even in the *Journal of the Institution of Electrical Engineers*.

We wonder if we dare suggest that it might be advisable to consider the substitution of amp for ampere as the name of the practical unit of current, or, if this is internationally impossible, making them alternatives of equal status. A little consideration will do much to remove any feeling of horror that such a suggestion may cause. The name of Volta was shortened to give the volt and the name of Faraday to give the farad; why should the name of Ampère be more sacrosanct? It has already been mutilated by the removal of the accent, but still its final syllable is more obtrusively drawn out than are those of Volta or Faraday. The second syllable of farad gives a snappy termination to the word, whereas that of ampere does just the opposite. The other most commonly used electrical units are monosyllabic. Phonetically an ampere does not fit in at all well with the volt, ohm, and watt, but an amp certainly does. Although there are some exceptions the great majority of commonly used units are monosyllabic ; inch, foot, yard, mile, knot, pound, stone, ton, gill, pint, quart, etc. If this suggestion were adopted one could speak and write of a current of 40 amps with a clear conscience. At present we read in the *Journal of the I.E.E.* of " a current of 40 amp on short-circuit," a strange hybrid, since the suppression of the dot and plural after abbreviations was subsequent to the abolition of amp. as an abbreviation of ampere.

A distinction should be made between typographical abbreviations and verbal abbreviations. One writes in, ft, yd, V, W, Ω, cm, but reads them as inches, feet, yards, volts, watts, ohms, and even centimetres ; these are merely typographical abbreviations. If a change in the name of the international unit of current is out of the question, it may be desirable for the B.S.I. to recognise two abbreviations of ampere, viz., amp as a verbal abbreviation and A as a typographical abbreviation, so that one could read 20A as 20 amps. A similar suggestion might be made with regard to 2,000 r.p.m. so that one could read it as 2,000 revs per minute without infringing any recognised standards. The plural of the verbal abbreviations would be formed in the usual way by the addition of " s ".

Our argument is supported by the names of the measuring instruments ; volts are read on a voltmeter, ohms on an ohmmeter, and watts on a wattmeter, but only those who are very old can remember when amperes were read on an amperemeter. The name was too cumbersome and was soon-replaced by ammeter.

Presumably the suggested alteration of the name of one of the most important electrical units could only be made by an International Electrical Commission, but it will be interesting to see whether it meets with the general approval of our readers or raises a storm of protest.

Although somewhat away from our subject we might mention, as an interesting example of the use of unsuitable units, an article in a recent number of the *Brown Boveri Review*, in which a new motor-coach was stated to be 22,700 mm long, 3,000 mm wide, and 3,560 mm high. The dimensions would look much more impressive if expressed in microns.

G. W. O. H.

REACTANCE MODULATOR THEORY*

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

SUMMARY.—In the simplified analysis of reactance modulators, the loading effect of the phase-shifting circuit is commonly neglected. By considering the impedance of this circuit in parallel with that of the modulator valve it is possible to select the component values of the circuit so that the complete two-terminal network is purely reactive.

The analysis covers both the conventional reactance modulator and the cathode-driven modulator. A simple bridge circuit is described, which may be used to determine the exact point of resistance neutralization.

1. Introduction

ASSUMING operation over a linear characteristic, the alternating component of anode current drawn by a valve which is subjected to simultaneous changes of anode and grid potential is given by the expression :—

$$I_a = \frac{E_a + \mu E_g}{r_a}.$$

Let $\mu E_g = (a + jb)E_a$,

where a and b are coefficients devised to represent, in magnitude and phase, the relationship between the anode and grid voltages. Then,

$$I_a = \frac{E_a}{r_a}(1 + a + jb).$$

Let Z_i = input anode impedance.

We have $Z_i = \dfrac{E_a}{I_a} = \dfrac{r_a}{1 + a + jb}$.

Three special cases can be distinguished. If $b = 0$, while a is real and positive, the input impedance is a pure resistance. When $b = 0$, a being negative and numerically

* MS. accepted by the Editor, January 1947.

greater than unity, the input resistance is negative. Under these conditions, the valve can be associated with an external parallel tuned LC circuit to form an oscillator. To generate sustained oscillations, the dynamic resistance of the tuned circuit must be numerically equal to, or greater than, the input resistance of the valve.

If $b \gg 1 + a$, the input impedance approaches a pure reactance. Alternatively, if $1 + a = 0$, the pure reactance condition is attained, while allowing an arbitrary choice of the coefficient b. The nature of the reactance is settled by the algebraic sign of b.

A valve which is operated in such a way as to draw current in quadrature with the applied anode voltage is defined as a reactor valve. The magnitude of the reactance is fixed by the valve parameters. Of these, the mutual conductance is readily variable by changes in control- or auxiliary-grid potentials.

There are two practical applications of reactor valves. The first relates to automatic frequency control of the oscillator of a superheterodyne receiver. Under certain conditions, the thermal drift of oscillator frequency may be sufficient to displace the intermediate frequency from the centre of the pass-band of the i.f. amplifier, causing distortion and, in extreme cases, total loss of the signal. Ganging errors cause similar trouble. The difficulty is resolved by detecting an error signal, represented by the difference between the nominal and actual intermediate frequencies and converting it into a steady potential, the magnitude and polarity of which are respectively proportional to the extent and to the sign of the frequency error. A reactor valve is shunted across the tuned circuit of the receiver oscillator and the control voltage is applied to it in such a manner as to reduce the initial error by an appropriate electronic change of the shunt reactance.

The second application is in frequency modulation. Here the reactor valve is connected in parallel with the tuned circuit of the primary oscillator of the f.m. transmitter. The audio-frequency modulation voltage is used to vary the instantaneous reactance and hence to swing the radio frequency of the oscillator between the deviation limits corresponding to full modulation. A valve used in this way is termed a reactance modulator. The electrical equivalent circuit of such a valve may be represented by a series or parallel combination of resistance and reactance. During modulation, the magnitudes of these quantities are subject to a cyclic variation. The change in reactance causes the desired frequency deviation of the transmitter. The resistance variation causes undesired amplitude modulation. Various methods have been devised to reduce or to eliminate the second effect. All are similar in that they cause the grid-cathode r.f. excitation voltage to lag or lead on the alternating anode potential by an angle greater than 90 degrees. The choice of method is based on considerations of circuit simplicity and on the effectiveness of resistance neutralization.

The mathematical expressions for the input impedance are often formulated as a series combination of resistance and reactance. In practice, it is more convenient to consider the equivalent parallel arrangement, which shows directly the effective damping and de-tuning components. For simplicity, the loading effect of the phase-shift circuit is usually neglected. It is this circuit which establishes the desired relationship, in magnitude and phase, between the anode and grid excitation voltages. It will be shown that when the auxiliary circuit impedance is included, the resulting complex expressions can be much simplified if it is stipulated that the input impedance of the entire network is to be purely reactive. This condition is readily attainable by the inclusion of simple additional components in the phase-shift network.

The general case will be considered and two simple resistance-cancellation circuits will be described.

2. Earthed-Cathode Reactance Modulator

Fig. 1 shows a typical reactance-modulator circuit. A triode valve is shown, although in practice it is more usual to employ a hexode mixer, isolating the audio- and radio-frequency circuits by the use of separate grids. Using the symbols shown on the diagram, the network equations become :—

$$I_1 = \frac{E_a}{Z_1 + Z_2}\,;\quad I_2 = \frac{E_a + \mu E_g}{r_a}\,;$$

$$E_g = \frac{Z_2}{Z_1 + Z_2} \cdot E_a\,.$$

If Z_i = total input impedance, then :—

$$Z_i = \frac{E_a}{I_1 + I_2} = \frac{(Z_1 + Z_2)r_a}{r_a + Z_1 + (\mu + 1)Z_2} \quad (1)$$

Setting $Z_1 = R_1 + jX_1$, $Z_2 = R_2 + jX_2$, substituting in Equ. (1) and separating the real and imaginary parts :—

$$Z_i = \frac{(R_1 + R_2)\{r_a + R_1 + (\mu + 1)R_2\} + (X_1 + X_2)\{X_1 + (\mu + 1)X_2\}}{\{r_a + R_1 + (\mu + 1)R_2\}^2 + \{X_1 + (\mu + 1)X_2\}^2} \cdot r_a$$
$$+ j \cdot \frac{(X_1 + X_2)\{r_a + R_1 + (\mu + 1)R_2\} - (R_1 + R_2)\{X_1 + (\mu + 1)X_2\}}{\{r_a + R_1 + (\mu + 1)R_2\}^2 + \{X_1 + (\mu + 1)X_2\}^2} \cdot r_a \quad \ldots \quad \ldots \quad (2)$$

The input impedance is purely reactive if the real term vanishes in Equ. (2). If this is the case :—

$$(R_1 + R_2)\{r_a + R_1 + (\mu + 1)R_2\} = -(X_1 + X_2)\{X_1 + (\mu + 1)X_2\} \quad \ldots \quad \ldots \quad (3)$$

The input reactance then becomes :—

$$X_i = \frac{(X_1 + X_2)\{r_a + R_1 + (\mu + 1)R_2\} - (R_1 + R_2)\{X_1 + (\mu + 1)X_2\}}{\{r_a + R_1 + (\mu + 1)R_2\}^2 + \{X_1 + (\mu + 1)X_2\}^2} \cdot r_a \quad \ldots \quad (4)$$

Eliminating $r_a + R_1 + (\mu + 1)R_2$ between Equs. (3) and (4) :—

$$X_i = -\frac{r_a(R_1 + R_2)}{X_1 + (\mu + 1)X_2} \quad \ldots \quad (5)$$

Using Equ. (3) again, an alternative expression is :—

$$X_i = \frac{r_a(X_1 + X_2)}{r_a + R_1 + (\mu + 1)R_2} \quad \ldots \quad (6)$$

The conditions under which Equ. (3) is satisfied may be realised by proper choice of the circuit elements forming Z_1 and Z_2.

The left-hand side of the equation is essentially positive, so that the factors on the right are necessarily of opposite sign and the reactances X_1 and X_2 must be of different kinds. Even with this proviso, there is still a wide choice of possible component values. The final selection is determined by the desired magnitude of the input reactance.

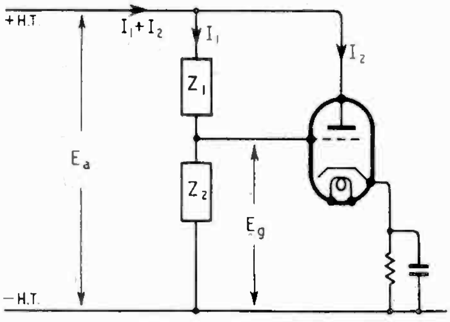

Fig. 1. *Basic circuit of typical reactance modulator.*

Fig. 2 shows two practical circuits which may be adjusted for resistance cancellation. In the first, the main phase-shifting elements are L and R. Since a blocking capacitor is essential, resistance neutralization is effected by variation of the capacitor C ; i.e., without the addition of further circuit elements.

The circuit of Fig. 2 (*b*) simulates an inductance and necessarily uses more components than the other. The effects of the blocking capacitor C_B and of the grid-leak R_g can be compensated by slight variation of the theoretical value of the compensating inductor L. The phase-shifting elements are in this case C and R.

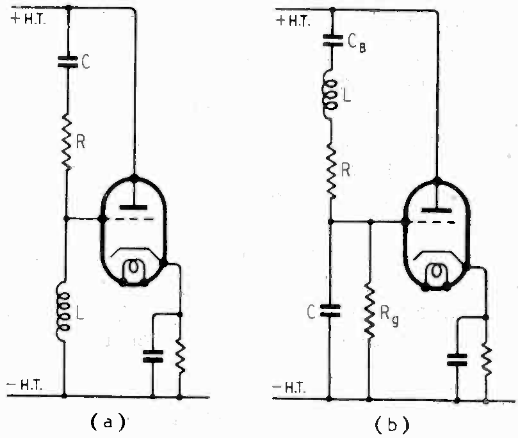

Fig. 2. *Practical circuits with resistance cancellation ; (a) simulates a capacitance and (b) an inductance.*

Referring to Fig. 1, the reactance modulator circuit becomes identical with that of the Miller integrator, used extensively in radar practice, if Z_1 is a pure capacitance and Z_2 a pure resistance. In the Miller circuit, a step-voltage applied to the grid is succeeded by a linear fall of anode potential. During this phase of the operation, the output impedance is known to be represented approximately by a resistance $1/g_m$ in series with a capacitance g_mCR where g_m is the mutual conductance of the valve, C is the anode-to-grid capacitance and R is the grid-circuit resistance. The same expression for the impedance can be derived from Equ. (1) by writing $Z_1 = 1/j\omega C$, $Z_2 = R$ and assuming that $r_a \gg 1/j\omega C$ and that $g_mR \gg 1$.

3. Input Admittance

Corresponding to Equ. (2), there is an expression for the input conductance and susceptance of the reactor valve shown in Fig. 1.

Let Z_i = input impedance, G_i = input conductance

Y_i = input admittance, B_i = input susceptance

$$\text{Then } Y_i = \frac{I_1 + I_2}{E_a} = G_i + jB_i = \frac{r_a + R_1 + jX_1 + (\mu + 1)R_2 + (\mu + 1)jX_2}{r_a\{R_1 + R_2 + j(X_1 + X_2)\}}.$$

$$\text{Hence } Y_i = \frac{(R_1 + R_2)\{r_a + R_1 + (\mu + 1)R_2\} + (X_1 + X_2)\{X_1 + (\mu + 1)X_2\}}{r_a\{(R_1 + R_2)^2 + (X_1 + X_2)^2\}}$$
$$+ j \cdot \frac{(R_1 + R_2)\{X_1 + (\mu + 1)X_2\} - (X_1 + X_2)\{r_a + R_1 + (\mu + 1)R_2\}}{r_a\{(R_1 + R_2)^2 + (X_1 + X_2)^2\}} \quad .. \quad (7)$$

To make the conductance zero, the requisite condition is :—

$$(R_1 + R_2)\{r_a + R_1 + (\mu + 1)R_2\} = -(X_1 + X_2)\{X_1 + (\mu + 1)X_2\} \quad .. \quad .. \quad (8)$$

This is precisely the same as that which previously caused the series resistive component to vanish. The equivalent parallel resistance is the reciprocal of the real part of Equ. (7) and the corresponding parallel reactance is the reciprocal of the imaginary term. The change of sign due to inverting the operator j should be noted. Using the condition described by Equ. (8) and following the same procedure as in Equs. (3), (4) and (5), the parallel reactance is given by :—

$$X_p = -\frac{r_a(R_1 + R_2)}{X_1 + (\mu + 1)X_2} \quad .. \quad (9)$$

The result is identical with that of Equ. (5).

4. Cathode-Input Reactance Modulator

Fig. 3 shows the schematic circuit of a cathode-input reactance modulator, which may be analysed by the method previously employed.

The network equations are :—

$$E_0 = I_1Z_1 + (I_1 + I_2)Z_2$$
$$= I_1(Z_1 + Z_2) + I_2Z_2\,;$$
$$E_g = (I_1 + I_2)Z_2\,;\; E_0 = E_a + E_g\,;$$
$$I_2 = \frac{E_a - \mu E_g}{r_a}.$$

$$Z_i = \text{input impedance} = \frac{E_0}{I_1 + I_2}$$
$$= \frac{r_a(Z_1 + Z_2) + (\mu + 1)Z_1Z_2}{r_a + Z_1} \quad .. \quad (10)$$

Setting $Z_1 = R_1 + jX_1$, $Z_2 = R_2 + jX_2$, substituting in Equ. (10) and separating the real and imaginary parts as before, we have :

$$Z_i = \frac{(r_a+R_1)\{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)\}+X_1\{r_a(X_1+X_2)+(\mu+1)(R_1X_2+R_2X_1)\}}{(r_a + R_1)^2 + X_1^2}$$
$$+ j \cdot \frac{(r_a+R_1)\{r_a(X_1+X_2)+(\mu+1)(R_1X_2+R_2X_1)\}-X_1\{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)\}}{(r_a+R_1)^2+X_1^2} \quad .. \quad .. \quad .. \quad (11)$$

The input impedance is purely reactive when :—

$$(r_a+R_1)\{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)\} = -X_1\{r_a(X_1+X_2)+(\mu+1)(R_1X_2 + R_2X_1)\} \quad .. \quad .. \quad .. \quad (12)$$

Eliminating $r_a(R_1 + R_2) + (\mu + 1)(R_1R_2 - X_1X_2)$ between Equs. (11) and (12), the input reactance becomes :—

$$X_i = \frac{r_a(X_1 + X_2) + (\mu + 1)(R_1X_2 + R_2X_1)}{r_a + R_1} \quad .. \quad .. \quad .. \quad (13)$$

Again using Equs. (11) and (12) :—

$$X_i = \frac{r_a(R_1 + R_2) + (\mu + 1)(R_1R_2 - X_1X_2)}{X_1} \quad .. \quad .. \quad .. \quad (14)$$

The equivalent parallel resistance and reactance may be found by a repetition of the process described under Section 3. The results are :—

$$R_p = \frac{\{r_a(R_1 + R_2) + (\mu + 1)(R_1R_2 - X_1X_2)\}^2 + \{r_a(X_1 + X_2) + (\mu + 1)(R_1X_2 + R_2X_1)\}^2}{(r_a+R_1)\{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)\}+X_1\{r_a(X_1+X_2)+(\mu+1)(R_1X_2+R_2X_1)\}} \quad .. \quad .. \quad .. \quad (15)$$

$$X_p = \frac{\{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)\}^2+\{r_a(X_1+X_2)+(\mu+1)(R_1X_2+R_2X_1)\}^2}{X_1\{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)\}-(r_a+R_1)\{r_a(X_1+X_2)+(\mu+1)(R_1X_2+R_2X_1)\}} \quad \ldots \quad (16)$$

When the resistance is neutralized, the input reactance is given by :—

$$X_p = \frac{r_a(X_1+X_2)+(\mu+1)(R_1X_2+R_2X_1)}{r_a+R_1} \quad \ldots \quad (17)$$

or, alternatively :—

$$X_p = \frac{r_a(R_1+R_2)+(\mu+1)(R_1R_2-X_1X_2)}{X_1} \quad \ldots \quad (18)$$

5. Generalized Reactor Valve

In the circuits so far discussed, resistance cancellation has been secured by the proper choice of the phase-shift circuit elements. The same effect may be attained by connecting particular impedances in series with the reactor valve. A typical circuit is shown in Fig. 4. The use of a correcting impedance in the anode circuit has already been described by Emrys Williams. His circuit is a particular case of the more general arrangement shown in Fig. 4.

The circuit equations are :—

$$E = I_1(Z_1+Z_2) = E_a + (Z_3+Z_4)I_2$$

$$E_1 = \frac{Z_2}{Z_1+Z_2}\cdot E = E_g + E_c;$$

$$I_2 = \frac{E_a+\mu E_g}{r_a};\ E_c = I_2Z_4.$$

Following the same procedure as before, the input impedance $\frac{E}{I_1+I_2}$ may be calculated, and the condition determined that it shall be a pure reactance. Writing the impedances Z_1, Z_2, Z_3 and Z_4 in the forms $Z_1 = R_1 + jX_1$, etc., and performing some tedious algebraic manipulations, it will be found that the necessary condition is :—

$$\{r_a+R_1+R_3+(\mu+1)(R_2+R_4)\}\big[(R_1+R_2)\{r_a+R_3+(\mu+1)R_4\} - (X_1+X_2)\{X_3+(\mu+1)X_4\}\big]$$
$$+\{X_1+X_3+(\mu+1)(X_2+X_4)\}\big[(X_1+X_2)\{r_a+R_3+(\mu+1)R_4\} + (R_1+R_2)\{X_3+(\mu+1)X_4\}\big] = 0 \quad \ldots \quad (19)$$

The input reactance is then given by :—

$$X_i = -\frac{(R_1+R_2)\{r_a+R_3+(\mu+1)R_4\}-(X_1+X_2)\{X_3+(\mu+1)X_4\}}{X_1+X_3+(\mu+1)(X_2+X_4)} \quad \ldots \quad (20)$$

An alternative expression is :—

$$X_i = \frac{(X_1+X_2)\{r_a+R_3+(\mu+1)R_4\}+(R_1+R_2)\{X_3+(\mu+1)X_4\}}{r_a+R_1+R_3+(\mu+1)(R_2+R_4)} \quad \ldots \quad (21)$$

Setting R_3, R_4, X_3 and X_4 all equal to zero, Equs. (20) and (21) reduce, respectively, to Equs. (5) and (6).

It is necessary to adopt some simplifying principle in applying the condition described by Equ. (19), since there are so many variables. Resistance neutralization can be attained by using either compensating element Z_3 or Z_4. One of them is normally redundant. In this case, the method of approach is to assign definite values to the components forming the phase-shift potential divider, and then to calculate the elements required in the correcting network. Much smaller reactance values are required when the compensation circuit is connected in the cathode lead of the reactor valve.

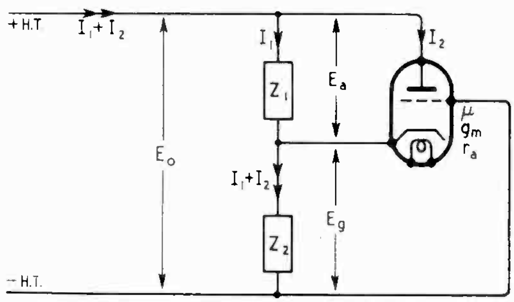

Fig. 3. *Cathode-input reactance modulator in skeleton form.*

6. Measuring Bridge

The equivalent resistance and reactance of a frequency-modulator valve may be determined by a simple bridge method. Fig. 5 shows a circuit designed for measurements at a high audio frequency.

The reactance modulator simulates a capacitance shunted by a certain loss resistance. It may be balanced against suitable standards in the bridge. When resistance-cancellation is secured, by proper adjustment of the resistance *R* and of the capacitance *C*, the shunt resistance in the adjacent arm of the bridge increases to infinity. At high frequencies, balanced and shielded transformers should be used. Precise measurements are quite difficult since changes in the phase-shift circuit affect simultaneously the impedance and phase angle of the reactor valve. The main purpose of the circuit shown is to demonstrate the possibility of resistance neutralization at one selected frequency.

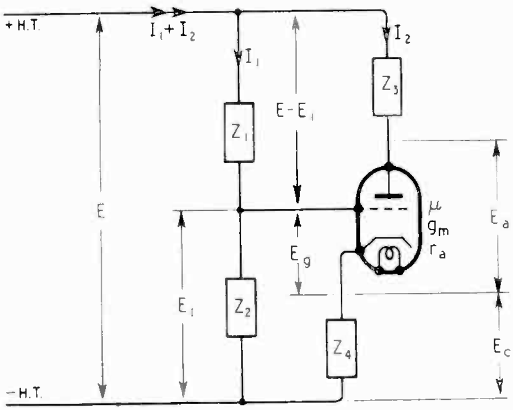

Fig. 4. *Generalized circuit of reactance modulator.*

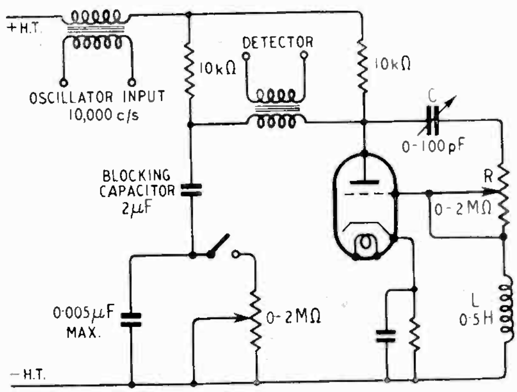

Fig. 5. *Bridge method of measuring modulator impedance.*

7. Conclusion

Efforts have been directed towards reducing the energy dissipation in reactance modulators by methods which are effective at one particular frequency. At other frequencies, resistance cancellation is not complete. If it is desired entirely to avoid the presence of amplitude modulation, it might be preferable to accept some dissipation in the modulator valve and to arrange that there is the minimum variation in the equivalent damping resistance during modulation. This possibility has not been examined theoretically in the present paper, though experiments have shown that with certain values of the phase-shift circuit components, even wide-band frequency modulation does not cause excessive amplitude changes in the oscillator output. These settings cannot be made to correspond to those which achieve resistance cancellation.

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TRANSMITTER-BLOCKER CELLS

Cm-Wave Fixed-Tuned Types

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(Concluded from page 61 of the February issue)

2. Further Considerations of the T.B.-cum-Magnetron Branch

IN this section we give an outline of the methods for assessing quantitatively the reception loss in a system employing a cell of a given edge-band impedance when operating with magnetrons of given cold resonance characteristics.

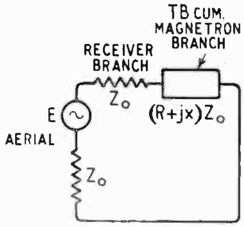

Fig. 10. *Equivalent circuit of the system for reception.*

We will follow the method first outlined by Samuel[3] to represent the conditions obtained in a system such as that depicted in Fig. 1. The circuit we will use is shown in Fig. 10 and, though the simplest possible, it does show very good agreement with conditions obtained in practice.

We will define the reception efficiency η as the ratio of the power absorbed in the receiver branch to the total available power, or the loss expressed in decibels below the value unity.

In the equivalent circuit of the complete system (Fig. 10) we envisage the system as being fed by a constant-voltage generator of internal resistance $Z_0 = 1$ at the aerial end of the main guide run, the load impedance consisting of the matched receiver branch in series with the effective impedance of the t.b.-cum magnetron branch $R + jX$ (normalized impedance) *referred to the plane of the receiver.*

Let E be the voltage of the generator, then the power absorbed in the receiver is

$$\frac{E^2}{(R+2)^2 + X^2}$$

The maximum available power is $\frac{E^2}{4}$, thus the receiver efficiency is

$$\eta = \frac{\text{Power absorbed in the receiver}}{\text{Maximum available power}}$$

$$= \frac{4}{(R+2)^2 + X^2} \quad \ldots \quad (2.0.1.)$$

From the form of the efficiency equation, it is seen that contours of constant efficiency, or constant loss, will be circles in the impedance (X, R) plane, and by inversion these will be transformed into circles in the reflection-coefficient plane (Smith Chart) also.

Since we are at the moment most interested in the effect which the magnetron impedance *transferred to the plane of the t.b.* has on the loss values, it will be more convenient to

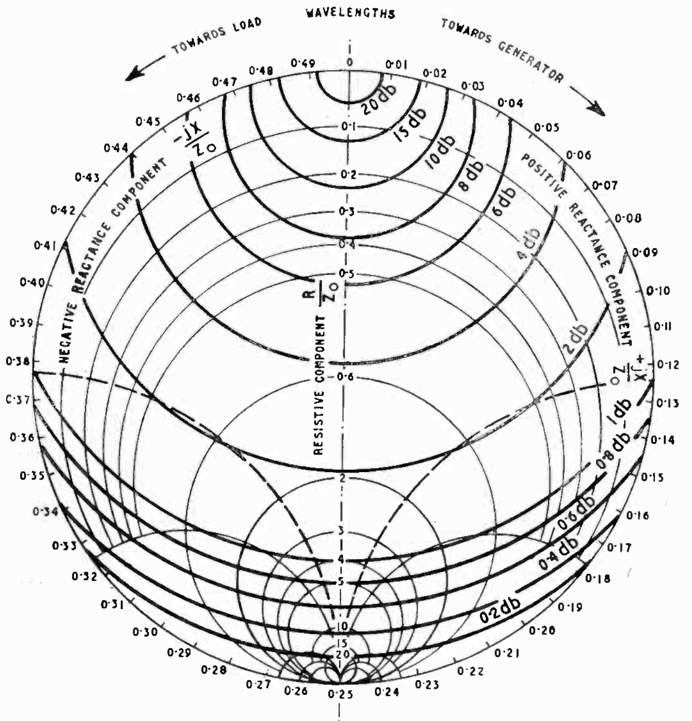

Fig. 11. *Smith chart showing contours of constant loss, referred to the impedance at the plane of the t.b.*

construct contours of constant loss referred to the impedance at the plane of the t.b. rather than to the receiver plane.

In the case of series-connected t.b. cells the distance between the t.b. and the receiver branch is usually $(2n + 1)\frac{\lambda_g}{4}$ where λg is the wavelength in the guide. Let $\star R + j\star X$ be the impedance of the t.b.-cum-magnetron *referred to the t.b.* plane, then transforming this impedance through a length of line of $(2n + 1)\frac{\lambda_g}{4}$ and substituting in the efficiency equation, yields :

$$\eta = \frac{4(\star R^2 + \star X^2)}{4(\star R^2 + \star X^2) + 4\star R + 1} \quad \ldots \quad \ldots \quad (2.0.2)$$

or

$$\left[\star R - \frac{\eta}{2(1 - \eta)}\right]^2 + \star X^2 - \frac{\eta}{4(1 - \eta)^2} = 0. \quad (2.0.3)$$

In this case (impedances referred to the t.b. plane), as in the case of impedances referred to the plane of the receiver, the contours of constant loss are circles. One family of circles may be obtained from the other by a rotation about the centre of the Smith Chart through an angle appropriate to the distance in guide wavelengths between the t.b. and the receiver branch. In Fig. 11 contours of constant loss are plotted referring the impedance to the plane of the t.b.

Such a plot is very convenient and gives constant loss contours in terms of the combined impedance of the t.b. cum-magnetron, referred to the plane of the t.b. It is possible now to go a step further and plot contours of constant loss showing regions within which the magnetron impedance must lie for the loss to be within certain specified values for a given value of t.b. impedance. The resistance and reactance values of the Smith Chart now refer to magnetron impedance at the plane of the t.b. whose impedance is some specified value $Z'_T = \star R'_T + j\star X'_T$. The equation of constant loss contours is then :

$$\left[\star R_m + \star R'_T - \frac{\eta}{2(1 - \eta)}\right]^2 + (\star X_m + \star X'_T)^2 - \frac{\eta}{4(1 - \eta)^2} = 0 \quad \ldots \quad (2.0.4)$$

where $\star R_m + j\star X_m$ is the impedance of the magnetron : $\star R_m$ and $\star X_m$ are the Smith Chart coordinates. Thus in Fig. 12 where $Z'_T = 3.8 + j13.8$ the equation of constant loss circles is :

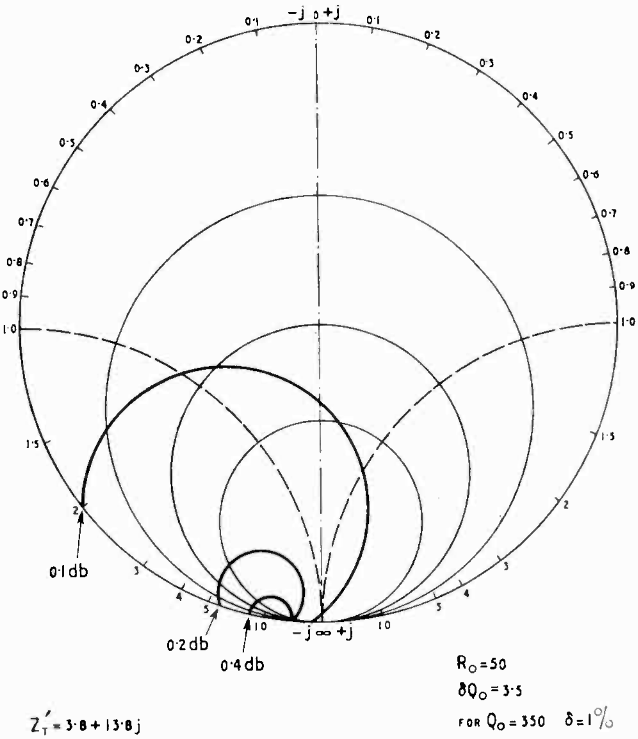

Fig. 12. *Contours of constant loss for $Z'_T = 3.8 + 13.8j$.*

$$\left[\star R_m + 3.8 - \frac{\eta}{2(1 - \eta)}\right]^2 + (\star X_m + 13.8)^2 - \frac{\eta}{4(1 - \eta)^2} = 0. \quad \ldots \quad \ldots \quad (2.0.5)$$

Similar contours are given in Figs. 13 and 14 for a t.b. whose values of edge band impedances are : $1 + 7j$ and $0.45 + 4.7j$.

For the other edge of the band (i.e., for t.b. impedances such as : $3.8 - 13.8j$, $1 - 7j$ and $0.45 - 4.7j$) the contours of constant loss are circles symmetrical to those of Figs. 12, 13 and 14, with respect to the real axis.

2.1 *Corrections*

The discussion so far has been rather idealized. Some corrections must be made ; first, for the fact that, though the t.b. may be positioned exactly an odd number of quarter guide-wavelengths behind the receiver branch at mid-band, at edge-band this figure will not be exact. A further correction is necessary to take into account the effect of the tolerance which must be placed on the mechanical distance t.b.-to-receiver branch in actual production sets.

The effect of a small change in the electrical distance between the t.b. and the receiver branch is considered in greater detail in Appendix III. In Fig. 14 dotted lines are contours of constant loss after correcting for the variation in guide wavelength over the frequency band. The distance between the t.b. and the receiver branch is exactly $\frac{3}{4}\lambda_g$ at mid-band; the correction is for 10,000 Mc/s; 1in × ½in internal dimensions, rectangular guide; a band of 3 per cent.

In practice it may well happen that the two corrections just discussed may be in opposing senses and tend to cancel one another; e.g., at lower frequencies, the correction for a positive tolerance in the length would be in a direction tending to cancel the wavelength correction and vice-versa. However, if the magnetron were replaced by one working at the other end of the band, the maximum tolerance would again have effect and it is well to use this figure to assess maximum possible loss.

To assess the loss likely to occur in a complete system, it now remains to consider in more detail the magnetron impedances likely to be met with in practice.

impedance," though a term universally employed to represent the impedance presented by the magnetron in the non-oscillating periods, may sometimes lead to inaccurate measurements if taken too literally. In making cold-impedance measurements, it is sometimes wise therefore to warm the valve from an external source so that the block is at the mean operating temperature reached by the valve when under pulsed operation and also to run the heaters to simulate the mean temperature of the cathode in operation.

Alternatively, the cold impedance characteristics may be measured while the valve is being pulsed under its normal operating conditions. This may be done with the apparatus shown diagrammatically in Fig. 15(a).

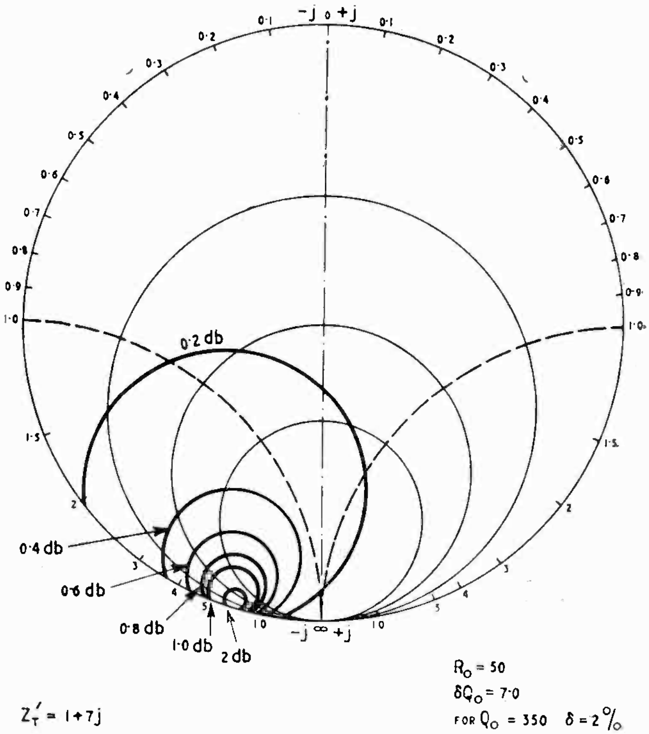

Fig. 13. *Contours of constant loss for* $Z'_T = 1 + 7j$.

2.2. *Magnetron "Cold-Impedance" Characteristics.*

The important characteristic of magnetrons, so far as the low-level operation of the magnetron-t.b. branch is concerned, is the spread in "cold impedance" characteristics at the operating frequency; i.e., the impedance at the operating frequency looking into a guide terminated by the valve in its quiescent state. In this condition it is sometimes important to remember that the magnetron block is often quite hot when the valve is running, so that the name "cold

The magnetron M is operated in the usual way into a matched load L, and c.w. power from a reflection v.m. oscillator R, working into matched load l, is fed into the guide in the direction shown via a directive feed D of good directivity and as large an attenuation as possible[4], the latter being dictated by the sensitivity of the receiver employed. A travelling-probe detector T near the magnetron extracts a sample of the power in the main guide and this power is led by coaxial cable to one of two possible display systems.

One may use one of the common types of spectrum analyser, in which the local-oscillator frequency in a u.h.f. receiver is varied by applying a saw-tooth sweep voltage to the reflector of the v.m. local-oscillator tube. The same sawtooth is used to supply the horizontal sweep of a cathode-ray tube, to which the output of the superheterodyne receiver is applied as a vertical deflection. The net result is a display in which horizontal deflection represents frequency, and vertical components represent the power in the main guide (e.g., in the magnetron spectrum) at those particular frequencies.

With the set-up described above, the radiation from the reflector oscillator, R, will also be displayed on the c.r. tube. This will in general display a very much narrower spectrum than that of the magnetron and by suitable adjustment of the sweep amplitude can be displayed as something so obviously different from the magnetron spectrum that there will be no possibility of confusing the radiation from the two sources. In any case, when measurements are being made at frequencies reasonably far from the magnetron frequency, and the mean frequency of the local oscillator in the receiver is adjusted so that the spectrum of R is displayed in the middle of the tube, the magnetron frequency will either be so far removed as to make confusion impossible or, depending on the sweep amplitude, may even be outside the tuning range of the receiver represented by the horizontal sweep and will not show up at all. When the oscillator R has its frequency within the bandwidth of the magnetron frequencies, the spectrum of R can still be distinguished as being always present whereas the magnetron spectrum will flicker. The vertical deflections occur only when the magnetron pulse happens to occur at such times as the local oscillator tuning in the receiver is at some value appropriate to a component in the magnetron's spectrum.

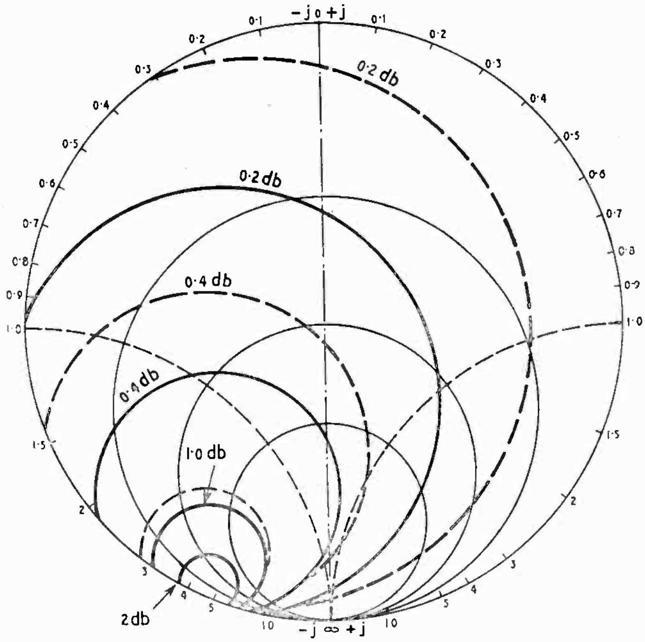

Fig. 14. *Contours of constant loss for* $Z'_T = 0{\cdot}45 + 4{\cdot}7j$.

The general appearance of this display is shown in Fig. 15(b) where R represents the steady and narrow reflection oscillator spectrum and M the broader magnetron spectrum. The vertical deflections M occur at random positions within the envelope of the magnetron spectrum (dotted) and their number in each sweep is determined by the ratio of the sweep time from x to y to the pulse repetition frequency of the magnetron. Their separation is determined by the sweep amplitude and by adjusting these two sweep parameters it is possible to obtain a convenient display for all values of the frequency of R.

If confusion is still unavoidable when the frequencies of R and M are equal, then some gating circuit may be employed so as to render the receiver inoperative when the magnetron is transmitting. In this case, the response due to the power fed in from R will be displayed alone.

An alternative display consists in leading the sampled power to a similar superheterodyne receiver whose local oscillator is not swept but is kept tuned to the frequency of R. The output of this receiver is applied as a vertical deflection to a c.r. tube, in which the horizontal deflection is a high-speed synchronous time base initiated by the circuits which pulse the magnetron simultaneously. The oscillator R is also pulsed with a fixed delay after the magnetron pulse so that the response of the receiver to this frequency is always separated by some convenient time interval from the magnetron breakthrough, whatever the frequency separation of M and R [see Fig. 15(c)].

Using either of these possible displays, the standing-wave ratio in the guide fed at the frequency of R is best measured by inserting a calibrated variable attenuator in the line to the receiver, such as an adjustable length of beyond-cut-off guide A in Fig. 15(a) and adjusting this attenuator so that the oscillator pip R on the c.r. tube display is at the same height when the travelling probe is first set on the maximum of standing waves and then at the minimum. The difference of the calibrated attenuator settings gives the power-standing-wave ratio directly in decibels.

The phase of the reflection from the magnetron is readily determined from the position of the probe which gives minimum pick up.

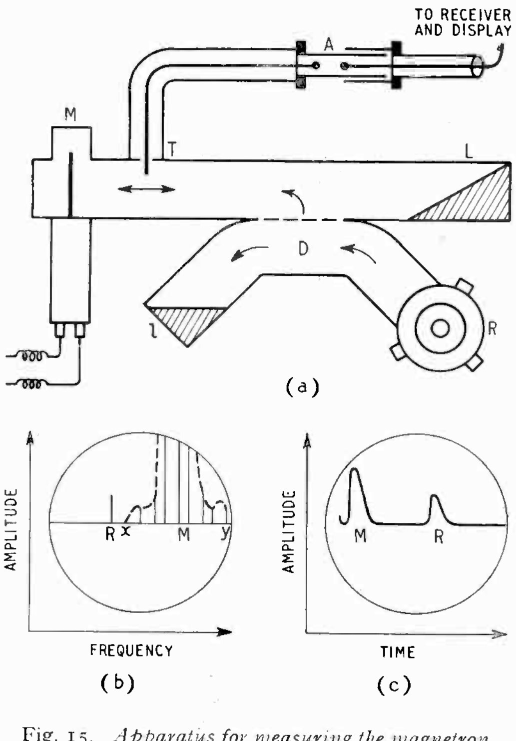

Fig. 15. *Apparatus for measuring the magnetron cold impedance* (a) *and two different forms of c.r. tube display* (b *and* c).

The frequency of R at which the measurements are being taken is determined by means of a cavity resonator inserted in the line to the receiver.

A typical plot of v.s.w.r. and position of the minimum of standing waves against frequency is shown in Fig. 16; d_m is the distance of the minimum from some arbitrary mechanical fixed origin, measured by a scale and vernier system relating the positions of the moving probe with respect to the slotted-guide section along which it travels. It is necessary therefore to convert these values of d_m to electrical distances in terms of the wavelength in the guide at each frequency. This is done by superimposing on d_m the grid of sloping lines converting d_m to distances in terms of λ_g from some convenient reference point on the magnetron; e.g., the output probe in this case. Their slope in the diagram is determined by the change in λ_g with λ and their position by the distance of the arbitrary origin of the mechanical scale from the probe. They may either be calculated, or plotted experimentally by replacing the magnetron by a short-circuited length of line, with the short-circuit occupying the position originally held by the magnetron output probe, to give the zero line of this grid.

The cold impedance characteristics observed can conveniently be divided into four distinct classes. These are illustrated in Fig. 17.

For convenience, the impedance is measured with reference to the position of the output probe of the valve and the impedance types met with correspond to four main types of valves :—

1. Valves which can be described as behaving as a parallel equivalent resonant circuit in series with the guide at the position of the output probe.

(a) When the movement of the minimum is as in Fig. 17(a) and the resistive component at resonance is then less than the characteristic impedance of the guide, or

(b) When the minimum moves as in Fig. 17(b) and the resistive component is greater than Z_0 at resonance.

2. Valves which behave as series equivalent resonant circuits inserted in series with the guide at the position of the output probe.

(a) When the movement of the minimum is as in Fig. 17(c) then $R < Z_0$ at resonance in the equivalent circuit, or

(b) When the minimum moves as in

Fig. 17(d) and $R > Z_0$ at resonance. Alternatively, types 2(a) and 2(b) may be regarded as valves of types 1(b) and 1(a) respectively with a transformation of one-quarter wavelength occurring somewhere before the probe output, to each of the cold impedance constants, L, R and C.

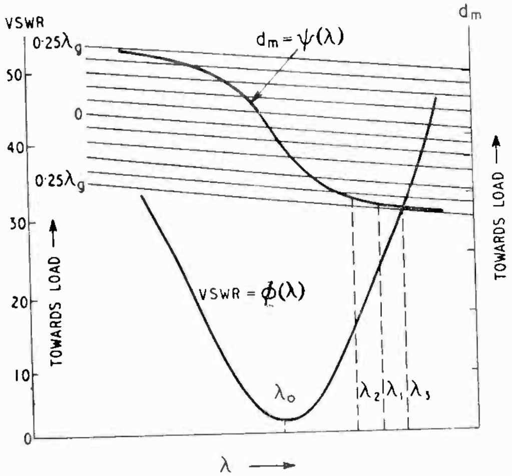

Fig. 16. *Typical plots of voltage-standing-wave ratio and position of standing-wave minima.*

In practice most valves are found to be of types 2(a) and sometimes types 1(b). The other types will obviously be unusual valves in practice and would be associated with very low efficiencies, since most of the power would be dissipated within the valve rather than in the load.

From the value of Q_L thus computed, Q_0 may be calculated in the two cases by employment of relations $Q_0 = Q_L(K_0 + 1)$ for the former type of resonance curve and

$Q_0 = Q_L\left(\frac{K_0 + 1}{K_0}\right)$ for the latter.[2]

The next characteristic to be looked for is the operating frequency of the valves. In general this will be well separated from the cold resonant frequency and on the low-frequency side of it (λ_{0p} in Fig. 17).

We can now assess the spread in cold impedance at the operating frequency. In general it will be found that the separation between the cold resonant frequency and the operating frequency will be of such a value that the curve of the position of the minimum of standing waves will have approached very near to the asympototic value (see λ_{0p} in Fig. 17) and the voltage-standing-wave ratio will be the order of 20 : 1 or 30 : 1. These values will usually be held within reasonable limits when the valves in question are "preplumbed" types, pretuned for operation with a restricted frequency pulling figure and with a certain efficiency.

By a series of such measurements on many valves we can arrive at a figure for the spread in cold impedance likely to occur at the operating frequency.

It may be instructive to outline the procedure in a case which actually occurred in practice. Measurements were made on

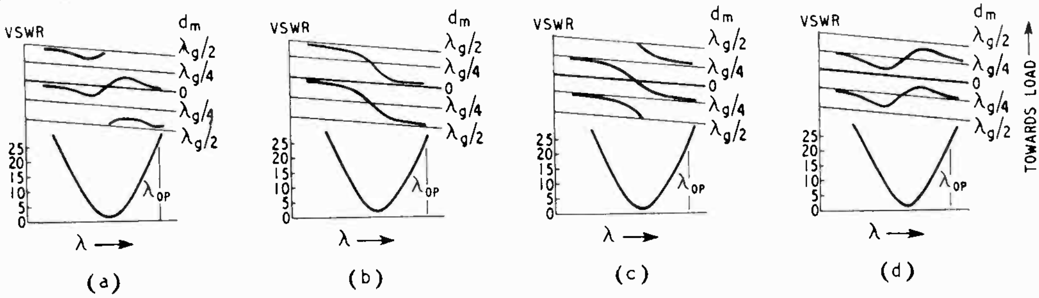

Fig. 17. *Magnetron cold-impedance characteristics corresponding to four types of valve.*

From curves such as these, the complete cold resonance characteristics of the valves may be plotted. For instance, Q_L may be computed as in Appendix A.1 for valves with cold resonance curves as in Figs. 17(b) and (c); the other type of curve may be dealt with by identical construction but here Q_L is given by $Q_L = \frac{\pi \Delta d/\lambda_{g0}}{\Delta f/f_0}(K_0 - 1)$.

many valves which were to be used in the system for which the t.b. of Fig. 8 was designed.

From these results it was found that the resonance curves were mostly of the type shown in Fig. 16 and that, the Q-values for all the valves being virtually the same, the resonant impedances were also the same; therefore conditions could be simulated for all these valves to a first approximation by inserting the abscissae λ_1, λ_2, λ_3 in Fig. 16, which is a plot for one valve of v.s.w.r. and d_m against λ. These are obtained by a determination of the values $\Delta\lambda = (\lambda_{RESONANT} - \lambda_{OPERATING})$ for many valves and then, using λ_0 in Fig. 16 as zero, plotting λ_1 at such a position that it is separated from λ_0 by the mean value of $\Delta\lambda$ and λ_2, λ_3 at the minimum and maximum values of $\Delta\lambda$ distant from λ_0. In this way, provided Q and K_0 are similar for all the valves concerned, their behaviour can be predicted with the aid of a complete cold resonance plot for one valve only, together with a measurement of $\Delta\lambda$ on a large number of valves—a much speedier process than making complete cold resonance plots for each. At the mean value, λ_1 it will be seen from Fig. 16 that the v.s.w.r. is 25 : 1 and that the minimum is displaced by $0.23\lambda_g$ from the probe toward the generator. The cold impedance referred to the probe is in this case $\approx 2.4 + 7j$. Corresponding values at the extremes, λ_2, λ_3 are $7 + 13j$, $2.2 + 4.2j$. With this type of valve we may therefore expect the cold impedance to be of order $(R \geqslant 2\,;\ X \geqslant 4)$ referred to the probe.

2.3. *Probable Loss and Magnetron Position*

It must be remembered that the mirror image of loss curves also exists for the other end of the band. We see immediately from Figs. 12-14 that the loss will be least serious for edge-band valves, with the above spread in cold-impedance characteristics, if the effective distance between the magnetron probe and the t.b. is fixed as an odd number of quarter-wavelengths. This gives but a low reactive component in the effective impedance of the magnetron at the plane of the t.b.

The data arrived at so far are usually sufficient to fix the best value of distance. To assess accurately the loss which may be expected to occur, it will usually be necessary to determine the cold impedance at the operating frequency for many valves and, once the spread has been determined, (1) to transform these values through the distance between the t.b. and magnetron to arrive at the spread in impedance presented by the magnetrons at the t.b. plane, (2) to correct this spread for the mechanical tolerances which, in practice, have to be placed on the distance. It is best to perform this correction on the assumption that the tolerance will add in such a direction as to increase the overall spread possible. (3) to correct for the change in guide-wavelength over the band, the distance initially being fixed as an odd number of quarter mid-band wavelengths.

Fortunately, this correction may well occur in the opposite direction to that for mechanical tolerance; e.g., for the lower-frequency valves the correction for a positive tolerance in the guide length will be partly cancelled by this latter correction and vice versa. However, it is best to consider the case when all corrections are adding so as to increase the effective possible spread in magnetron impedance referred to the t.b.

From the corrected spread in impedance just arrived at it is obviously possible to assess the probable loss by plotting the figures obtained into characteristics such as those illustrated in Figs. 12–14.

A modified procedure is desirable in systems employing magnetrons whose cold-impedance characteristics are not to some extent limited in spread by "pre-plumbing," with reasonably close tolerances placed on other parameters of the valve, such as, pulling figure, efficiency, frequency band; or with magnetrons tunable by a piston behind the output probe or by "crown of thorns" cavity tuning. It is also employed in systems in which the bandwidth is so great that corrections (1)—(3) outlined above are sufficient to increase the possible spread in the impedance, referred to the t.b. plane, to the order of a complete circle in the reflection coefficient plane. The best approach is merely to determine the average magnitude of the voltage-standing-wave ratio set up by the magnetrons when terminating a guide fed at the operating frequency, without taking further time to analyse this parameter into $\star R_m$ and $\star X_m$ components; i.e., taking no note of the position of the standing-wave pattern with respect to the valve. This can be done quickly by one of the methods outlined in Section 2.2. and from these data alone we can arrive at reasonable impedance criteria to be aimed at in designing a fixed-tuned t.b. for the system. Two further curves will be found useful for this purpose :—

We have seen that the impedance components of the t.b. may be related to R_0, the resistive component at resonance, Q_0 the unloaded Q, and $\delta = \dfrac{2\Delta\omega}{\omega_0}$ a measure of the frequency separation of the point being considered from the resonant frequency, by the relations

$$\star R_T = \frac{R_0}{1 + Q_0{}^2\delta^2}\,;\quad \star X_T = \frac{jR_0Q_0\delta}{1 + Q_0{}^2\delta^2}\,.$$

Also from Fig. 12 we can read off immediately the figures of Table I relating values of $\star R_T$ with the maximum possible loss, assuming a purely reactive magnetron. Thus :—

TABLE I.

$\star R_T$	9	7	6	5	4.5	4	2.6	1.9	1.5	1.2	1.0	0.85
Max. Loss Poss. (db)	0.5	0.6	0.7	0.8	0.9	1.0	1.5	2.0	2.5	3.0	3.5	4.0

Now, from these figures and using the relation $\star R_T = \frac{R_0}{1 + Q_0^2\delta^2}$ we can construct a further table giving values of R_0 which will limit the system to a given maximum loss at an operating point $Q_0\delta$ removed from ω_0. For example, for the maximum edge-band loss to be 1 db, $\star R_T$ at edge band must be 4, from Table I ; and the values of R_0 necessary to give this $\star R_T$ value at a point $Q_0\delta$ removed from ω_0 are given in Table II, for several values of $Q_0\delta$.

TABLE II.

$Q_0\delta$	1	2	3	4	5
R_0	8	20	40	68	104

If this process is repeated for various chosen values of maximum allowable edge-band loss (i.e., for various $\star R_T$) the curves of Fig. 18 can be plotted showing the values which R_0 must have for the edge-band loss to be limited to a given maximum value, for various $Q_0\delta$ values appropriate to the bandwidth.

If δ is known, these curves are useful for immediately arriving at an estimate of the possible R_0 and Q_0 values necessary to keep the loss within the specified limit over the bandwidth δ. For example, if the maximum loss tolerable is 2 db, we can have values of $Q_0\delta$ ranging from 1.3 to 7.8 for a range of R_0 from 5 to 100 determined by the curve marked " 2 db " in Fig. 18.

If, in addition, the bandwidth is, say, $\pm$ 2% corresponding to a δ value of 4%, we have, for the possible values of R_0, Q_0 values given by the modified scale of abscissae shown and ranging from $Q_0 = 32$ to $Q_0 = 195$ as R_0 varies from 5 to 100, the precise values being read off from the 2-db curve.

All values of R_0 and Q_0 to the left of this curve will produce a useful cell and from previous experience we can select a range of R_0 and Q_0 which we are likely to achieve.

In addition, once a cell is designed and R_0, Q_0, δ known we can use this curve for a quick estimate of the probable edge-band loss.

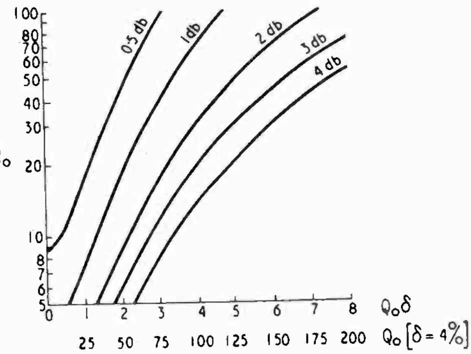

Fig. 18. *Curves relating R_0 and $Q_0\delta$ for various edge-band losses.*

We show in Fig. 19 how the initial value arrived at above, or the experimental values for R_0 and Q_0 may be used to estimate, in more detail, performance at edge-band and over the band. It will readily be seen that, if we decide on possible values for R_0 (20, 40, 60 have been chosen in Fig. 19), then using the relation $\star R_T = \frac{R_0}{1 + Q_0^2\delta^2}$ we can arrive at the resistive component of the cell at points separated from the resonant point by $Q_0\delta$ and then from Fig. 11 the maximum power loss may be read off directly from the value of the resistive component by following round the appropriate circle of constant $\star R_T$.

Thus the full curves of Fig. 19 can be drawn showing how the maximum possible loss for given R_0 will vary with $Q_0\delta$ (the separation of the operating frequency from the resonant frequency) ; i.e., how the loss will vary over the band for given R_0.

It is now possible to take into account the standing-wave ratio presented on reception by magnetrons which, as in practice, are not purely reactive. Suppose for example, that the experiment described at the beginning of this section indicates an average value for this field-standing-wave ratio of $m = 20:1$. With a system in which there are not tight tolerances it can be assumed that the impedance presented at the plane of the t.b. by the average magnetron may lie anywhere on the circle $m = 20:1$ on the Smith Chart (Fig. 20). We now wish to find how, for a given R_0, the loss will vary with $Q_0\delta$ when the t.b. cell is followed by a magnetron characterized by $m = 20:1$.

This is done by the following graphical construction.

For a given R_0 and $Q_0\delta$ the t.b. impedance components will be defined by a given $\star R_T$ and $\star X_T$. The locus of the possible range of effective impedance at the plane of the t.b. for this R_0 and $Q_0\delta$, when the t.b. is followed by the $m = 20:1$ magnetron, are determined by adding to the appropriate point $\star R_T + j\star X_T$ on the Smith Chart, all the impedance values possible with $m = 20:1$; i.e., all the points lying on the circle $m = 20:1$. The locus is a circle, for given $\star R_T + j\star X_T$ (i.e., for given R_0 and $Q_0\delta$) and a series of these loci are shown drawn in Fig. 20 for a cell characterized by $R_0 = 20$, for $Q_0\delta$ values 6, 5, 4, 3, 2, 1. The maximum loss possible under these conditions for each of these $Q_0\delta$ values can now be read off directly by superimposing Fig. 20 on Fig. 11. The result can then be plotted directly in Fig. 19

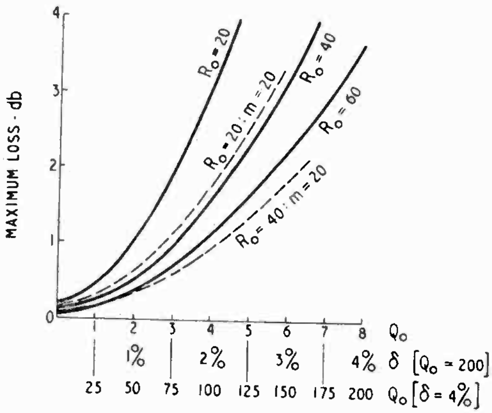

Fig. 19. *Variation of maximum loss with $Q_0\delta$ for various values of R_0.*

and this is illustrated by the dotted curves in that figure.

We have, then, in Fig. 19 a set of curves which can be regarded as showing how the loss varies over the band for given values

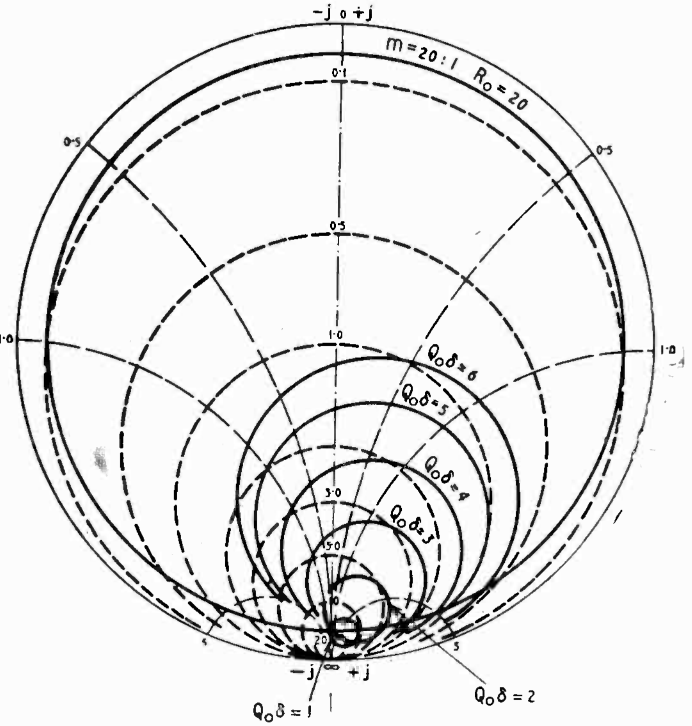

Fig. 20. *Smith chart used in relating loss with the t.b. cell and magnetron characteristics.*

of R_0 or, δ being known, can be used to determine what Q_0 values will be necessary to limit the loss to certain values, for given R_0. This curve can obviously be repeated for several values of m and R_0 giving the maximum loss as a function of $Q_0\delta$, and covering the spread in magnetron impedance (m) and cell impedance (R_0). However, initial estimates of the sort described above are sufficient, and the quickest check on loss over the band is the experimental method in which the receiver branch is terminated by a thermistor mount of the wide-band type to give adequate matching over the frequency range. C.W. power is fed in from the aerial end and the power entering the receiver branch is monitored by the thermistor as a piston is moved in the main guide beyond the t.b. cell. In this way reasonably informative plots of possible loss against t.b. impedance may be obtained at various frequencies. The oscillator feeding in c.w. from the "aerial end" should be isolated from possible effects on its power output and frequency, due to variation in r.f. loading as the piston is moved, by means of a pad attenuating by some 10 db inserted in the main guide run between the oscillator and the main apparatus. Power output can be monitored as a control figure, by means of a directive feed and thermistor between the oscillator and the attenuator.

3. Increasing Bandwidth

It will readily be seen that the factor limiting the Q value is in the main the Q of the resonant window itself since the Q of a quarter-wave stub alone is of order 1 and a decrease in the Q of the resonant window should result in a decrease in the Q of the t.b. as a whole.

In determining the Q value for the window experimentally, the usual method is to measure the variation of the input impedance to a guide across which the window W is mounted, followed by a well-matched load. [See Fig. 21 (a)]. It must be remembered, however, that in measuring the Q in this way, the window is loaded effectively by $Z_0/2$. However, when used in a t.b. cell the window is loaded by $2Z_0$ so that its loaded Q is increased by a factor of 4 [compare Figs. 21 (a) and (b)].

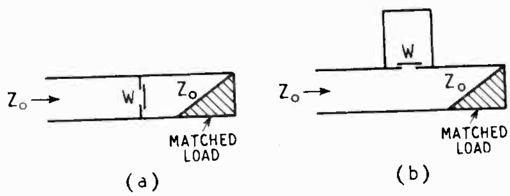

Fig. 21. *Loading of the window when measuring its Q (a) and loading when used in the t.b. (b).*

For the cell illustrated in Fig. 8 the Q of the window measured when loaded by $\frac{Z_0}{2}$ as above was found to be of order 3 to 4, so that as used in the cell itself its loaded Q will be increased by a factor 4 to, say, 12 to 16 and has then the same value as that found for the loaded Q of the complete t.b.; the loaded Q of the $\lambda_g/4$ stub alone being of order 1.

The useful bandwidth of the t.b. is then mainly limited by the loaded Q of the window. A reduction in this Q value can be obtained by increasing the size of the window in the direction of the electric vector in so far as this is consistent with limitations imposed by mechanical and high-power level requirements. Chief among the latter is the requirement for low transmission loss, which implies among other things that a high voltage should be developed across the window for ease in striking, and this in turn implies a not too low Q; a factor directly opposed to a large bandwidth under reception conditions. There is also the high-level requirement for short recovery time in most radar applications, so that echoes from near by targets will not be seriously attenuated. In practice, the long recovery time with noble gas fillings is reduced by the introduction of water vapour. The requirements are again rather opposed since water vapour causes an increase in the transmission loss in the arc. A compromise is therefore necessary in the face of these conflicting requirements.

Having obtained the best bandwidth by choosing the largest possible window, it may be further improved by decreasing the dispersion of the guide of which the stub is constructed, or of both waveguides.

Briefly, in the case of $b = c$, the bandwidth will increase if Y of Section 1 is decreased and the coupling coefficient is made nearer to unity. This may be done by decreasing the value of b/λ_g by :—

(1) decreasing b, the limiting factor being the power-handling capacity of the waveguide.

(2) increasing λ_g by decreasing a. The limiting factor here is the increase in the dispersion of the waveguide which may decrease the bandwidth.

(3) The coupling may also be increased by using a 120° Y-junction. The bandwidth may in this case be increased by a factor of approximately two.

The exact mathematical treatment of the factors mentioned above is complex to derive and cumbersome to handle, especially in the case $b \neq c$; in the case of Y-junctions it has not yet been fully worked out. The treatment is then usually experimental and Fig. 22 shows a comparison of the variation of r and v.s.w.r. with frequency for a cell similar to that of Fig. 8 when mounted in a T- and in an 120° Y-junction. The decrease in Q_0 and increase in R_0 in an 120° Y-junction are well illustrated. A further possibility for increasing the bandwidth of the system is to employ more than one t.b. cell. For example, the case of using two cells with the same resonance point and separated in the guide by $\frac{\lambda_g}{2}$ is equivalent to doubling $\star R_T$ at all frequencies provided the bandwidth is not so great as to make dispersion corrections important. There are of course countless further possibilities, such as using cells with staggered tuning points with various frequency differences between these resonances and separating such cells in the guide by various distances. These

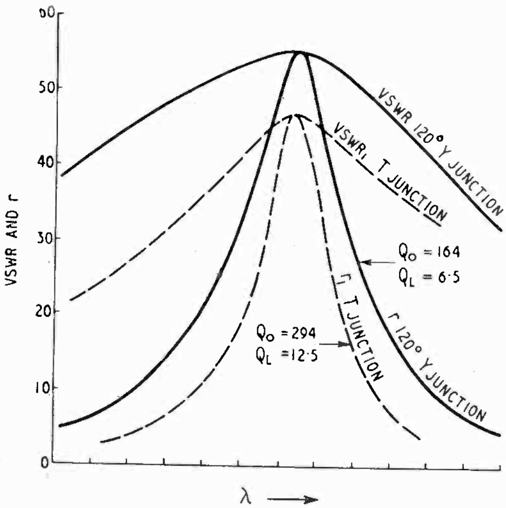

Fig. 22. *Variation of r and v.s.w.r. with frequency for a cell mounted in a Y-junction and also for a T-junction*

possibilities are in themselves an extensive subject and outside the scope of this paper which is concerned chiefly with the design of the single cells which would go to make up such a system.§ The mathematics of such complex systems is rather cumbersome and an experimental treatment is probably to be preferred since the many correcting factors mentioned in this report add further to the complexities of these multiple-cell systems.

Acknowledgment

Acknowledgment is due to all at A.S.E. who have helped with this work, to the Captain Superintendent for granting facilities for the preparation of the draft copy of these notes and to the Board of Admiralty, with whose approval this report is published.

§ Nicoll and Fry have reported investigations on multiple-cell systems in a Telecommunications Research Establishment report.

APPENDIX I

The Q values may also be determined from the movement of standing waves, measured near the t.b., as a function of frequency[2]. Again, this is not of particularly easy application here, when the movement of the minimum is in general small over a large range of frequencies. The method will, however, be outlined for the sake of completeness.

The method consists in plotting the position of the minimum of standing waves near the t.b. as a function of frequency. The curve $d_m = D(f)$ (where d_m is the distance of the minimum from some arbitrary origin) runs smoothly at frequencies far from the resonant frequency but in the vicinity of f_0 the slope changes, the rate of change being greater the greater the Q. The type of curve met with in the present case is shown in Fig. 23.

The Q values are determined by drawing the curves $d_s = D'(f)$, the asymptotes to which the curve $d_m = D(f)$ tends at frequencies far removed from f_0. This latter curve may alternatively be plotted directly by replacing the t.b. cell by a short-circuiting piston across the main guide at the plane AA, in Fig. 2 and plotting $d_m = D'(f)$ for this system. At f_0 the corresponding point on the curve $d_m = D(f)$ should be $\frac{\lambda_{g0}}{4}$ distant from the appropriate point on the curve $d_s = D'(f)$ where λ_{g0} is the wavelength in the guide at the resonant frequency f_0.

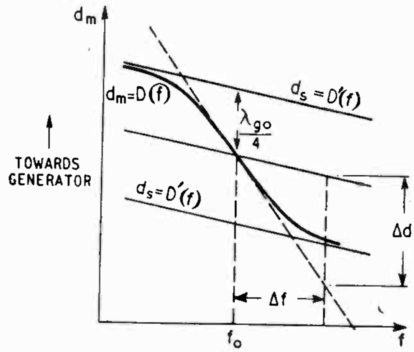

Fig. 23. *Movement of the minima of standing waves against frequency, plotted near the t.b. plane.*

If we now determine f_0 and k_0 and at the corresponding point on the curve $d_m = D(f)$ draw two straight lines, the one tangential to the curve $d_m = D(f)$ and the other parallel to the tangent to the curve $d_s = D'(f)$ and also, at a distance Δf from f_0, draw a line parallel to the d-axis cutting the other two lines in intersections separated by Δd (see Fig. 23), then Q_L is given by :

$$Q_L = \pi \frac{\Delta d}{\lambda g_0} (1 - 1/k_0) \Big/ \frac{\Delta f}{f_0} \quad \ldots \quad \ldots \quad (A.1.1.)$$

As mentioned initially, when low Q values are encountered in this type of circuit, the function $d_m = D(f)$ is not very sensitive to frequency change and the method would be rather laborious.

APPENDIX II

Let us assume that the average magnetron with which the t.b. is used in actual practice presents on reception v.s.w.r. of $m = \frac{20}{1}$ at its operating frequency. Assume for this purpose that the distance magnetron-to-t.b. has not been chosen to have any particular value. Then, in general, maximum loss *does not occur* when the reactance of the t.b. is cancelled by that of the magnetron.

To prove this statement, consider the particular case of the maximum possible reactance which such magnetrons may present to the t.b., then find out what are the resistive components of the magnetron and t.b. corresponding to this reactive component. The loss is then readily calculated, and comes out smaller than the loss corresponding to a case where there is no cancellation of reactive components. To calculate the maximum possible reactance of the magnetron, we explore along a line terminated with an impedance setting up a v.s.w.r. "m." There is a point at which the impedance is purely resistive and equal to m. Let this point be the origin. Then at any point along the line (see Fig. 24) the impedance is given by :

$$Z_x = \frac{m + j\tan\beta x}{jm\tan\beta x + 1} = \frac{m(1 + \tan^2\beta x)}{1 + m^2\tan^2\beta x} + j\frac{\tan\beta x(1 - m^2)}{1 + m^2\tan^2\beta x} \quad \ldots \quad \ldots \quad (A.2.1)$$

This may be rewritten as :

$$Z_x = \star R_m + j\star X_m \quad \ldots \quad \ldots \quad \ldots \quad (A2.2)$$

where :

$$\star R_m = \frac{m(1 + \tan^2\beta x)}{1 + m^2\tan^2\beta x}\,;\ \star X_m = \frac{(1 - m^2)\tan\beta x}{1 + m^2\tan^2\beta x}, \quad \ldots \quad \ldots \quad \ldots \quad (A.2.3)$$

β being the propagation constant.

Then solving the equation $\frac{\partial X_m}{\partial x} = 0$, it is found that the maximum value of $\star X_m$ is

$$\star X_M = \pm\frac{m + \frac{1}{m}}{2} \quad \ldots \quad \ldots \quad \ldots \quad (A.2.4)$$

and the corresponding value of the resistive components is :

$$\star R_m = \frac{m + 1/m}{2} \quad \ldots \quad \ldots \quad \ldots \quad (A.2.5)$$

Thus for a magnetron which sets up a v.s.w.r. of $m = 20/1$ on reception $\star X_m$ is always less than 10, and when it has its maximum value the resistive component is ≈ 10.

Now consider the equivalent circuit of the t.b. as depicted in Fig. 5. If the cell is regarded as introducing a series impedance Z, we have

$$Z = \frac{R_0}{1 + Q_0{}^2\delta^2} - j\frac{R_0Q_0\delta}{1 + Q_0{}^2\delta^2}$$

Where the *series* resistive and reactive components are

$$\star R_T = \frac{R_0}{1 + Q_0{}^2\delta^2} \text{ and } \star X_T = -\frac{R_0Q_0\delta}{1 + Q_0{}^2\delta^2}$$

respectively.

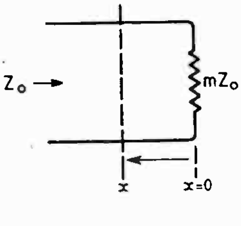

Fig. 24. *Equivalent circuit used in calculating maximum magnetron reactance.*

Taking the t.b. cell whose characteristics are plotted in Fig. 8 we have $R_0 = 50$. The series resistive component $\star R_T$ of the t.b. when its series reactive component is equal to $-\star X_m$, say to 10, is found by solving for $Q_0\delta$ the equation

$$\star X_m = \frac{R_0Q_0\delta}{1 + Q_0{}^2\delta^2} = 10,$$

then carrying this value of $Q_0\delta$ into the expression for $\star R_T$.

Performing this operation, the following two values of $\star R_T$ are obtained ; $\star R_T \approx 50$ and $\star R_T \approx 2$.

As the series resistive component of the magnetron is ≈ 10, then the combined resistances at the t.b. are ≈ 60 and ≈ 12, which correspond to a loss in the system of 0.08 db and 0.3 db, respectively, since the combined reactance is zero.

Now, for a certain distance of magnetron-to-t.b., which can be found from the Smith Chart, the resistive and reactive components of the magnetron ($m = 20/1$) at the t.b. plane are $\star R_m \approx 2$ and $\star X_m \approx 6$. Under these conditions in the most unfavourable case the combined series resistance (t.b. + magnetron) is $2 + 2 = 4$, and the combined reactance is $10 - 6 = 4$. The corresponding loss is 0.55 db. Thus, although the reactive components do not cancel, the loss is larger than in the preceding case. A similar state of affairs is clearly depicted in Fig. 14.

APPENDIX III

To assess the effect of a small change in the electrical distance between the plane of the t.b. and the receiver branch plane, consider the efficiency equation relating efficiency with impedances referred to the plane of the receiver branch $\left(\text{i.e., } \eta = \frac{4}{(R + 2)^2 + X^2}\right)$, and let $\star R + j\star X$ be the combined impedance referred to the plane of the t.b.

We assume now that the electrical length between the two planes is $(2n + 1)\frac{\lambda_g'}{4} + \Delta l$, then transforming $\star R + j\star X$ to the plane of the receiver branch we obtain :

$$Z = \frac{\star R + j\star X + j\tan\frac{2\pi}{\lambda_g'}\left[(2n + 1)\frac{\lambda_g'}{4} + \Delta l\right]}{j(\star R + j\star X)\tan\frac{2\pi}{\lambda_g'}\left[(2n + 1)\frac{\lambda_g'}{4} + \Delta l\right] + 1} \quad \ldots \quad \ldots \quad \ldots \quad (A.3.1)$$

Now, neglecting terms of higher order we may write,

$$1/t = \tan\frac{2\pi}{\lambda_g'}\left[(2n + 1)\frac{\lambda_g'}{4} + \Delta l\right] = -\frac{1}{\tan\frac{2\pi\Delta l}{\lambda_g'}} \approx -\frac{\lambda_g'}{2\pi\Delta l} \quad \ldots \quad \ldots \quad \ldots \quad (A.3.2)$$

where t is a small quantity. Then

$$Z = \frac{\star R + j\star X + j/t}{j(\star R + j\star X)/t + 1} = \frac{t(\star R + j\star X) + j}{j(\star R + j\star X) + t} \approx \frac{t(\star R + j\star X) + j}{j(\star R + j\star X)} = \frac{1}{\star R + j\star X} - jt = \frac{\star R}{\star R^2 + \star X^2} - j\star\frac{X + t(\star R^2 + \star X^2)}{\star R^2 + \star X^2}$$

Thus at the plane of the receiver branch the resistive and reactive components are :

$$R = \frac{\star R}{\star R^2 + \star X^2} \text{ and } X = -\frac{\star X + t(\star R^2 + \star X^2)}{\star R^2 + \star X^2}$$

Substituting these values in the efficiency equation, and after some simplification, we obtain :

$$\eta = \frac{4(\star R^2 + \star X^2)}{(4 + t^2)(\star R^2 + \star X^2) + 4\star R + 2t\star X + 1} \quad \ldots \quad (A.3.3)$$

when $t = 0$, we obtain, as we should do, the efficiency equation for impedances referred to the plane of the t.b

Now, neglecting t^2 and ηt^2 as compared to 4, and after some straightforward algebraic manipulation, we obtain the equation of constant loss circles as :

$$\left[\star R - \frac{\eta}{2(1-\eta)}\right]^2 + \left[\star X - \frac{t\eta}{4(1-\eta)}\right]^2 - \frac{\eta}{4(1-\eta)^2} = 0 \quad \ldots \quad (A.3.4)$$

Comparing this equation with the uncorrected one (Eq.2.3) we see that the effect of a small change in the electrical distance between the t.b. plane and the receiver branch, is to change the edge-band reactance.

In the impedance plane, where R and X are Cartesian co-ordinates, the circles of constant loss have the same radius as those of Equ. (2.03)., but their centre is displaced along a line parallel to the $\star X$ axis.

Considering the correction due to variation in guide wavelength at edge-band, it may be pointed out that $\star X_T'$, the reactance of the t.b. at edge-band, is always of the same sign as t. Thus for a frequency lower than the resonant frequency, $\star X_T'$ is positive ; λ_g' corresponding to the lower frequency is larger than λ_g at resonance, hence Δl is negative and t is positive. Thus for the same magnetron and edge-band t.b. impedance, the larger the correction due to variation in guide wavelength, the larger is the loss.

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Smith Chart :—See P. H. Smith "Electronics," 1939, Vol. 12, p. 29. 1944, Vol. 19, p. 130.

Willis Jackson : "High Frequency Transmission Lines," Methuen, 1945.

CORRECTION

The equation in the first part of this article in the February issue, page 60, line 13, should have read :—

$$Q_0 = \frac{1}{\delta} = \frac{f_0}{|f' - f''|}$$

NEW BOOKS

Very High-Frequency Techniques

By the staff of the Radio Research Laboratory, Harvard University. Pp. 1057 + xvii + viii in two volumes, with 895 illustrations. McGraw-Hill Publishing Co., Ltd., Aldwych House, London, W.C.2. Price 84s.

The Radio Research Laboratory operated by Harvard University was engaged on the development of countermeasures against enemy radar, and to this end it was concerned largely with the extension of continuous-wave technique to radio-frequencies above 100 Mc/s. The book "represents a summary of the methods, theories and circuits used by the Radio Research Laboratory that it is believed will be of general interest to radio engineers and physicists."

Most books dealing with the war-time developments at very-high frequencies which have hitherto been published have been primarily concerned with pulse techniques and radar methods. It is, therefore, quite a change to find a book in which pulses are hardly mentioned and it makes it one which is likely to be of unusually great value to the communications engineer.

In reading the book it is necessary to bear in mind that much of the work was carried out to provide apparatus for interception on the one hand and jamming on the other. Unless this is done, some statements in the book will appear rather extraordinary to those accustomed to ordinary practice. Thus, in dealing with the modulation of power oscillators it is stated that "the principal objectives are (1) maximum side-band power, regardless of phase shifts, distortion, and incidental frequency modulation and in most cases (2) maximum bandwidth." This is strange reading until one remembers that the discussion is about the requirements for a jamming transmitter.

The first 272 pages of the book are concerned primarily with aerials and include chapters on Broad-Band Antennas, Cone & Cylinder Antennas, Sleeve Antennas, Horns & Reflectors, Slot Antennas, Impedance Matching Transformers & Balun Measurements and Direction Finding Aerials. This section is extremely valuable. It is much less mathematical than is customary in aerial literature and it is profusely illustrated by measured impedance curves of a large number of aerial types.

The importance of impedance matching and its effect on the bandwidth is very clearly treated. The use of the Smith Chart calculator is explained and it is shown how with its aid the aerial resistance and reactance are easily calculable from voltage standing-wave measurements.

Following the aerial section, there is a short chapter on Homing Systems, and then comes a section on Power Generation. This occupies 317 pages and covers Triode & Pentode U.H.F., Oscillators, Coaxial-line Power Amplifiers & Oscillators, Power Output Coupling Methods, Modulation, General Design Considerations, the Resnatron, Magnetrons, and Power-Measuring Devices. In general, the treatment is not analytic and it is very largely non-mathematical. It is very practical, however, and well-illustrated by drawings of oscillator assemblies. The emphasis is always on apparatus for a wide frequency range rather than on equipment for spot-frequency operation. This makes design much more difficult and so greatly enhances the value of the treatment.

The chapter on Power Measurement covers most of the possible methods and explains their characteristics and the difficulties which are likely to be found.

The rest of the book is devoted to the receiving side and includes two very good chapters on the principles and design of transmission-line filters. They cover the use of lumped elements, coaxial resonators and cavity resonators for the elements in low-pass, high-pass and band-pass filters. Of particular importance is the description of a method of improving the termination by the use of end sections of greater bandwidth than the main filter instead of *m*-derived sections, which are often impracticable.

There are chapters on r.f. tuners and detectors and mixers, and three chapters on the local oscillator in which the butterfly, reflex-klystron, and coaxial line oscillators are treated. Chapters on i.f. amplifiers, receiver output circuits and receiver measuring equipment conclude the book.

The choice of intermediate frequency is covered and there are design charts for stagger-tuned amplifiers as well as for coupled-pair types. There are also many practical hints on the arrangement of parts and on decoupling.

Panoramic presentation is treated and the difficulties arising through spurious responses are well explained.

Throughout the book the emphasis is on practice and theory is given with a minimum of mathematics and in a readily understandable way. It is a book which should be invaluable to all concerned with frequencies of over 100 Mc/s. There is also a good deal in it of application to somewhat lower frequencies.
W. T. C.

Electromechanical and Electroacoustical Analogies and their use in computations and diagrams of oscillating systems

By BENT GEHLSHOJ. Pp. 142, with 81 illustrations. G. E. C. Gad. Vimmelskaftet 32, Copenhagen K. Price Kr. 12.

This is based upon work done for a higher degree in the Royal Technical College of Denmark. Although translated by a Dane and printed in Denmark there are relatively few grammatical or typographical errors. The paper, printing, and diagrams are all very good ; the book has a stout paper cover. The subject matter is divided into three parts, viz., mechanical systems, acoustical systems, and electro-mechanical transducers, followed by a bibliography of forty-six books and papers on the subject and a list of symbols.

It is shown that a simple mechanical system consisting of a mass with friction and a spring can be represented either by a series impedance diagram or by a parallel admittance diagram, and on page 12 the author says that, although the latter is called the inverse analogy in recent English and American literature, the designation is ambiguous because it seems to be the more direct. Again on p. 73 we read that " there are two schools ; one goes in for the use of impedance diagrams, while the other—led by Firestone—advocates the superiority of admittance diagrams. . . . When everything is taken into consideration it looks as if the admittance diagram should be given preference." Kirchhoff's laws, high-pass and low-pass filters, and magnetic and crystal pick-ups are all considered in detail.

In the second section electrical analogies and Kirchhoff's laws are applied to acoustic systems, microphones and telephone receivers, and employed to determine the frequency response. In the final section electromagnetic and electrostatic transducers are considered, including moving-coil, moving-armature, and electrostatic microphones and loudspeakers ; piezo-electric devices are also very fully considered. The treatment throughout is very clear and the book can be strongly recommended.
G. W. O. H.

Grundgesetze der Regelung (The fundamental laws of control)

By WINFRIED OPPELT. Pp. 118 with 32 illustrations. Wolfenbütteler Verlagsanstalt, Wolfenbüttel, Hanover.

This is a paper-backed austerity publication intended primarily for students ; it is one of a series of Bücher der Technik, published as an emergency measure to bridge the gap until it is possible to publish larger text-books. In addition to the 32 figures there are 28 so-called tables which are really collections of small diagrams. Some of these small diagrams are intended to illustrate the mechanical and electrical devices to which the mathematical treatment applies, others give the various loci or graphical solutions of the equations. The treatment is compressed and mathematical. A knowledge of linear differential equations is assumed and Laplace transforms are introduced. In the introduction the author says that the book contains the basic mathematical equipment, the practical application of which will be dealt with in another volume. The book should prove of great interest to those who are concerned with the mathematical treatment of feed-back or servo-mechanism methods of automatic control. It is for students of this specialized subject and not for the ordinary undergraduate that the book is intended.
G. W. O. H.

The Measurement of Radio Interference by the Modified Reception Set R. 206, Mk. I. Technical Report, Reference M/T90

By S. F. PEARCE, B.Sc., A.Inst.P., A. TURNEY, B.Sc. (Eng.)| and D. C. G. SMITH, B.Sc. The British Electrical and Allied Industries Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Pp. 19 + 12 illustrations. Price 13s. 6d. (Postage 4d.)

I.E.E. RADIO CONVENTION

A convention on Scientific Radio is being held by the Radio Section of the Institution of Electrical Engineers on 7th and 8th April. Organized in collaboration with the British National Committee for Scientific Radio, which has been set up by the Royal Society, the convention is a preliminary to the meeting later in the year of the Union Radio Scientifique Internationale at Stockholm.

The sessions are to be held from 2–4.45 and 6–8.45 each day and cover : standards and measurements, propagation, radio noise and radio physics. The chairmen for the different subjects will be respectively Mr. P. Good, president of the I.E.E., Sir Edward Appleton, Sir Robert Watson-Watt and Prof. G. I. Finch, and the sessions are all to be held at the I.E.E., Savoy Place, London, W.C.2.

PHYSICAL SOCIETY'S EXHIBITION

The catalogue for the exhibition, which is to be held at Imperial College, South Kensington, London, S.W.7 on 6th-9th April inclusive, is expected to be available about the middle of March. The price is 2s. 6d. to members and 5s. to non-members, plus 1s. postage.

INTERFERENCE MEASUREMENT*

Effect of Receiver Bandwidth

By G. L. Hamburger, Dipl.-Ing., M.Brit.I.R.E., A.M.I.E.E.

(Formerly of the British Electrical and Allied Industries Research Association.)

(Concluded from page 54 of the February issue)

5. Repeated Impulses

If a number of impulses is impressed on the band-pass amplifier the output meter will give a steady reading owing to a state of equilibrium existing between the amount of charge accumulated due to the rectified transients and the amount of charge leaking away through the 5-MΩ diode load. By increasing the number n of pulses during the discharge time-constant (500 milliseconds) the meter reading will gradually approach the condition of crest value indication. The equation governing the meter reading[3] is derived in Fig. 13, and in a general case will be—

$$V_0 = \frac{\Delta f}{\pi f_m} V_p g \frac{0.54\frac{n}{\Delta f}}{1 + 0.54\frac{n}{\Delta f}} = V_T p,$$

$$\text{and } p = \frac{0.54\frac{n}{\Delta f}}{1 + 0.54\frac{n}{\Delta f}}$$

where

V_0 = steady meter reading
V_p = height of input voltage pulses
g = gain of the amplifier
Δf = bandwidth (3 db) in kc/s.

The curve for p is plotted in Fig. 14, and shows that the narrower the bandwidth and the more frequent the repetition of the impulses, the closer the reading approaches the crest values V_T of the transients. Experiments were made to check the validity of this formula. However it follows from the considerations of overload conditions discussed in the previous section that large errors will be incurred, particularly for conditions where the function p is substantially below the 100 per cent mark. Indeed the measurements show that the error increases very much for smaller values of $n/\Delta f$; i.e., for low repetition rates and large bandwidths.

The measuring procedure is indicated in Fig. 15. The repeated impulses of height V_p produce a certain steady reading X on the meter of the 5-Mc/s amplifier. The signal generator is then connected in place of the impulses, tuned to mid-band frequency, and its output voltage V_{SG} adjusted to give the same reading X on the meter. It can easily be seen from Fig. 15 that

$$\frac{\Delta f}{\pi f_m} p = \frac{V_{SG}\sqrt{2k}}{V_p}$$

where the left-hand side of the equation represents calculated, and the right-hand side, measured quantities, so that again comparisons can be drawn.

Three sets of measurement were made by

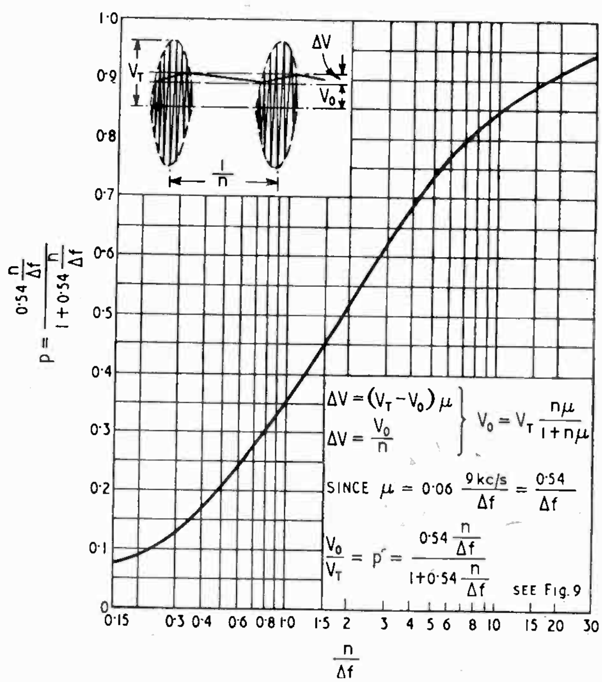

Fig. 14. *Meter reading for repeated impulses; Δf bandwidth in kc/s, n = number of pulses during discharging time constant.*

applying repeated pulses at the rate of 200 to 9,000 c/s to the 4 sets of filters and by going through this procedure three times. The errors incurred are plotted in the three diagrams of Fig. 16 (a), (b), (c), and it can be seen that there is quite a distinct tendency for the errors to decrease for a rising value of $n/\Delta f$; i.e., for higher rates of repetition and narrower bandwidths.

6. Measurement of Motor Noise

As a last item the effect of bandwidth upon the measurement of the noise produced by d.c. motors was investigated. This category of interference is very frequent, particularly in industrial areas, and it is therefore important to be able to assess its effect upon communication channels of different bandwidths.

Motor noise is of intermittent nature. During the actual commutation of the rotor windings abrupt electric disturbances occur while relative silence prevails between the commutations, assuming, of course, a clean commutator. At first sight, therefore, it seemed to be not too far fetched to expect a behaviour similar to that of repeated impulses as discussed in Section (5). Measurements, however, have revealed that this type of motor noise follows the same law as smooth fluctuation noise; viz., the square root of bandwidth law. In fact, three different motors have been measured under conditions analogous to those for smooth noise as described in Section (3), and this law was well borne out. The results of the measurements are given in Table 3.

The causes for this unexpected phenomenon were then examined more closely, and again an oscillographic method was employed to study the nature of motor interference.

An ordinary fractional horse-power series motor was made the object of this investigation. Fig. 11 (d) is an oscillogram of the total voltage across the brushes and shows the expected substantially wavy trend, where each cycle corresponds to one commutator segment. During the actual commutation, however, fine vertical lines are discernible which represent bursts of interference of essentially high frequency. In the following it will be shown how a number of bursts of interference, and their effect on a bandpass amplifier, belonging to one particular commutator segment, have been photographed individually as they occur after successive revolutions.

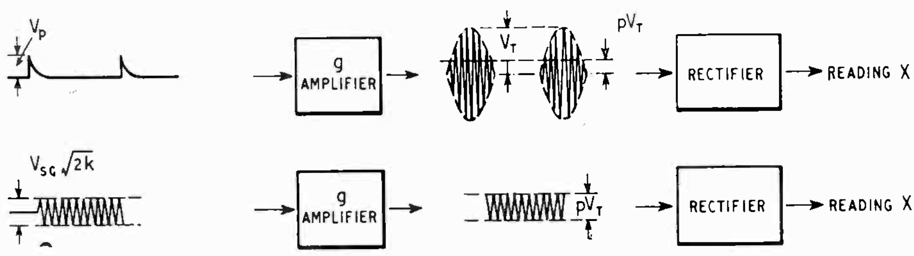

Fig. 15. *Measurement of repeated impulses.*

$$V_{SC}\sqrt{2kg} = pV_T = p\frac{\Delta f}{\pi f_m}V_P g. \therefore \frac{\Delta f}{\pi f_m}p = \frac{V_{SC}\sqrt{2k}}{V_P}$$

The problem consists of isolating the interference created by the commutation of one particular segment; viz., by taking a small time interval within which the interference occurs, expanding it right across the screen in order to resolve the fine structure of the interference, and then taking care that the oscillograms of successive commutations (of that particular segment) are allocated different portions on the screen so that they can be kept apart without moving the camera.

The principal lay-out of the scheme is sketched in Fig. 17. The interference

TABLE 3

D.C. Motor Noise measured at Different Bandwidths.

Filter		A	B	C
Response differences in db calculated from the $\sqrt{\Delta f_N}$ law		3.1	2.1	4.0
Response differences in db measured with-	Motor 1	3.15	2.06	4.5
	Motor 2	2.85	1.95	4.56
	Motor 3	3.05	2.00	4.45
	Diode noise	3.1	2.1	4.3

generated across the brushes is prevented from breaking down across the speed control potentiometer by inserting an h.f. choke so that its amplitude can be kept to the order of 5 to 10 V peak, enough to give a satisfactory direct indication on the c.r. tube without any amplification. Coupled with the motor is an alternator which generates one cycle of a.c. per revolution. This synchronizing a.c. is fed to a pulse generator which produces a short rectangular pulse of about 40-microseconds duration once per revolution. A simple variable phase shift RC network inserted between the alternator and pulse generator then permits the 40-microseconds pulse to be adjusted to occur at any particular phase of the armature revolution so that any of the commutations can be brought within the 40-microseconds interval and be observed. The pulse then triggers the single-stroke time base of the E.R.A. Universal Oscillograph. Therefore a single sweep of 40-microseconds duration occurs once per revolution. The same triggering pulse is also fed to the modulator grid of the c.r. tube, thus switching on the beam for the duration of the sweep and blanking it out for the rest of the revolutions. The oscilloscope really serves only as a monitor to enable the operator to adjust the phase so that the interference does indeed occur during the actual sweep. This adjustment is rather critical owing to slight speed variations of the fractional horse power motor and its ensuing frequency and phase drifts. It must be borne in mind that the 40 microseconds represent about one thousandth of one total revolution (i.e., about one-third of a degree) so that obviously a very slight drift in phase will bring the interference right outside the sweep range and, therefore, beyond observation.

As it was found important to observe the interference going into and coming out of the band-pass amplifier simultaneously, two more cathode-ray tubes were required. For this purpose a double c.r. tube unit was constructed, together with a separate power pack giving an e.h.t. variable between 2 and 4.5 kV. All X-plates were put in parallel with those of the monitoring oscilloscope, thus furnishing the recurrent single sweep to the double-tube unit. Also a photographic attachment was built so that the simultaneous oscillograms on both screens could be photographed on one plate.

Lastly, a scheme had to be devised which would permit the displacement of successive sweeps across the screen like the lines of a television raster (only with much wider spacing), and which would also switch on the, otherwise blacked out, beams for the time the raster pattern was being produced. It was realized that balanced deflection is absolutely essential for producing uniform sharpness all over the screen, and this was achieved by utilizing anode and cathode voltages of the triodes V_2 and V_3, which in turn are provided with equal anode and cathode loads. The common control grids are furnished with a rising voltage saw tooth arising from charging a capacitor C through

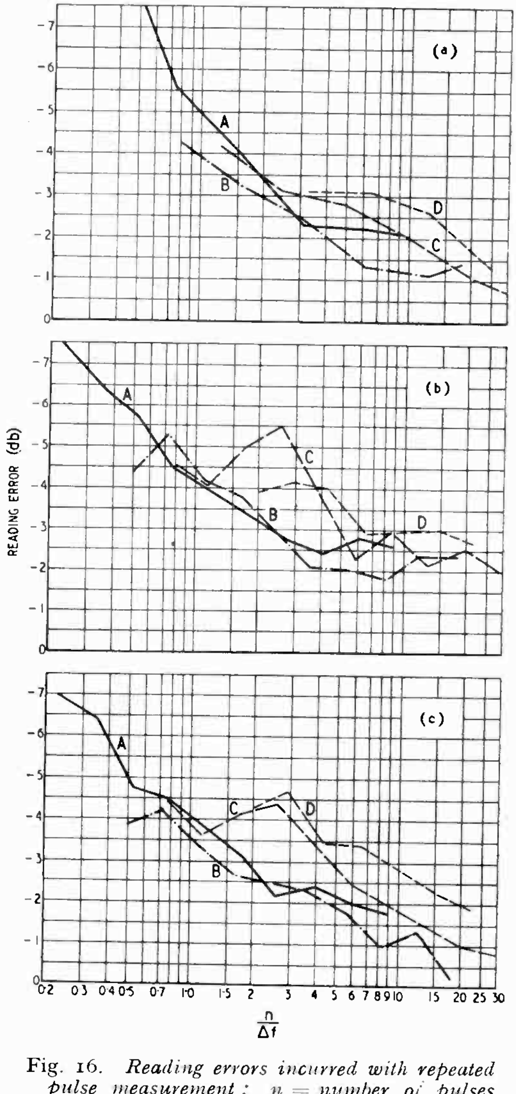

Fig. 16. *Reading errors incurred with repeated pulse measurement: n = number of pulses per discharge time (0.5 sec), Δf bandwidth of filter.*

a resistance R. This comparatively slow saw tooth is initiated by the opening contact spring of a relay. At the same time another contact spring of the same relay puts a negative bias on one grid of the double triode V_1, so that its anode voltage rises. V_3 via small capacitors, and is thus superimposed upon the two sets of balanced vertical sweep voltages. Therefore pressing the key operates the relay, produces a slow vertical sweep across the screen and switches on the beam so that the camera with open shutter records a dozen or so lines well displaced from each other and showing the interference created by one and the

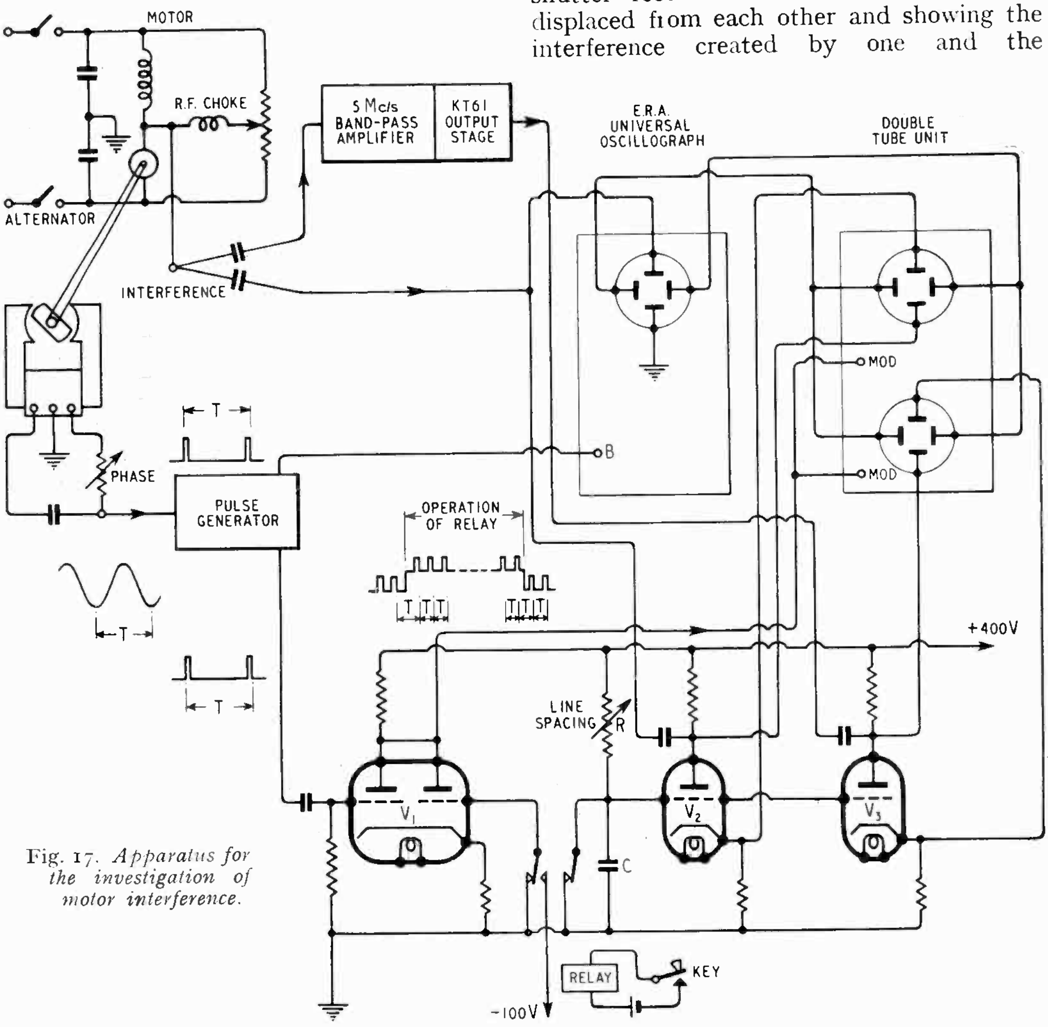

Fig. 17. *Apparatus for the investigation of motor interference.*

Since the other grid of V_1 is fed continuously with a negative 40-microsecond pulse, the total common anode voltage of V_1 is the sum of positive 40-microsecond pulses and a long positive pulse lasting as long as the relay operates. The common anode voltage of V_1 then modulates the two c.r. tube grids of the double-tube unit, hence the two beams are switched on for the sweep duration only as long as the relay is operating. The in- and outgoing interference itself is loosely coupled to the respective anodes of the triodes V_2 and same commutator segment at successive revolutions.

It is interesting to note that after a certain amount of experimenting it was found that the best photographic results were obtained by using tubes with green fluorescent screens with the maximum of about 4.5 kV accelerating potential and a highly sensitive orthochromatic film of the type 5G91. It was found that the available blue tubes gave a focus quality essentially inferior to that of the green tubes of the same type and, remarkably enough, their photographic efficiency was lower with panchromatic as well as orthochromatic material.

A number of photographs was taken with this complex arrangement and some samples are shown in Fig. 18. The first column shows the actual interference as created by one commutator segment after consecutive revolutions; i.e., each line depicts the voltage across the brushes for a duration of 40 microseconds within which that particular segment commutates. Each line represents another commutation of the same segment and it is interesting to note that no two such individual commutations give the same interference. The duration, strength and relative phase position of each successive burst of interference is different from the rest. The exact moment, for instance, when the interference occurs (i.e., its phase position) varies so much as to be completely outside the 40-microsecond interval at times. The photographs, therefore, lead to the conclusion that the bursts of interference have a random characteristic regarding their duration, strength and phase (relative to the geometrical position of the segments) inasmuch as those three quantities vary around a most probable and average value.

In the second column are photographs taken simultaneously with the first and show the transients initiated in the band-pass amplifier by each particular burst of interference. The events can be checked line by line. It is instructive to see how, for instance, the closely packed cluster of interference on the second line from the bottom of the top left picture produces a certain complex transient on the corresponding line of the top middle picture which is obviously the sum of a number of small overlapping transients whereas the more long drawn out interference pattern on the fourth line from the bottom produces smaller and more separated transients. Similar observations can be made at the corresponding pictures of row B, C and D. It can be seen how, with narrowing the bandwidth (by going from filters A to D) the transients become longer, and also how the delay time between, say, the centre of an interference cluster and the maximum of its corresponding transient increases as the bandwidth becomes narrower.

In the third column the photographs show transients as produced by amplified fluctuation noise under otherwise identical conditions of oscillographic display. An inspection of the second and third columns reveals a striking similarity between the transients caused by the bursts of commutation interference and those caused by smooth noise; i.e., fluctuation or genuine random noise. The only apparent difference is that the transients due to fluctuation noise go on uninterruptedly whereas those due to motor interference exist only for short spasms and occur as many times per revolution as there are segments on the commutator. As already indicated, it can be seen by close inspection of the actual interference (first column) that the individual pulses, as far as is discernible from the photographs, are indeed random in amplitude, frequency of occurrence, and mutual spacing. One can therefore stipulate the following conclusion from the oscillographic exploration of motor-interference :—

Motor interference can, in effect, be regarded as an intermittent fluctuation noise; that is to say, as a fluctuation noise which is switched on as many times per second as there are commutations per second, the switching on period varying randomly in length and exact phase position.

Owing to the limitations of the method of investigation employed (i.e., the limitation of resolution on the screen), it is difficult to say just how many individual pulses do occur within one particular cluster of interference. It is desirable to know this for one could then determine a certain (rather wide) bandwidth where the individual transients caused by the individual pulses would not overlap, so that the random vectorial addition caused by overlapping transients initiated by even identical pulses but of random spacing or phase, would be eliminated.

With an output meter of the C.I.S.P.R. type* the response to randomly spaced pulses of random height could then be found by a complicated elaboration of the method for regularly spaced pulses, as treated in Section (5)†.

The top middle photograph, and other samples not reproduced in this paper, show at times an apparent isolation between some individual transients, but it is difficult to say for certain whether this is due to the absence of impulses or whether the impulses

* C.I.S.P.R. = Comité International Spécial de Perturbation Radiophonique.

† It may be noted that, with a square-law output meter, the random noises need only be spaced sufficiently closely to be averaged by the meter in order to preserve the analogy with fluctuation noise.

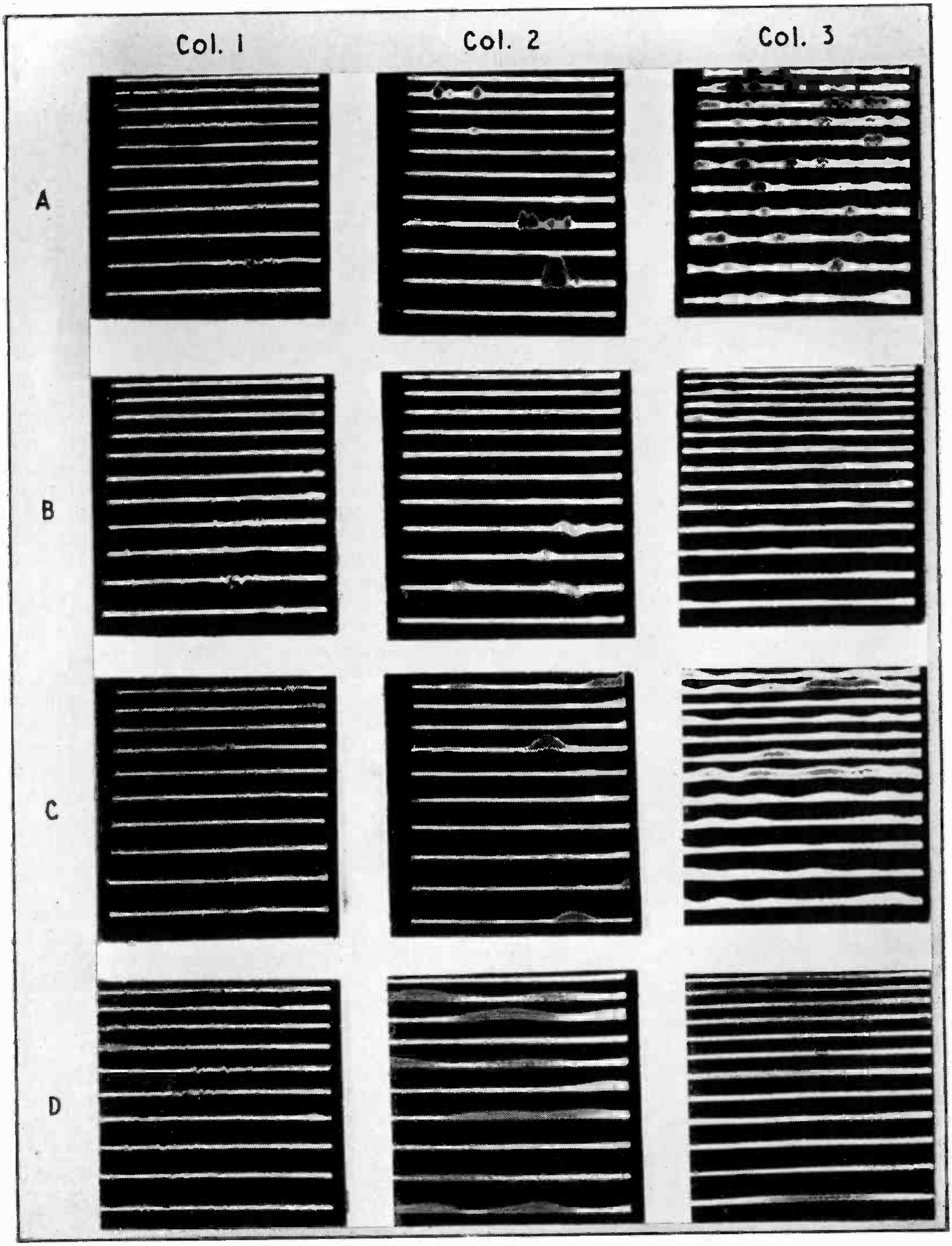

Fig. 18. *Motor interference as applied to the amplifier is shown in Col. 1. The transients due to motor interference and fluctuation noise respectively as illustrated in Cols. 2 and 3.*

are too small during these silence periods within one cluster to produce recognizable transients. In any case, the square root of bandwidth law holds up to 500 kc/s and it is not unreasonable to assume that at some bandwidth in the megacycle region motor-noise measurements might deviate from the square root of bandwidth law. However, this remains to be established.

Within the bandwidth range considered the indication of motor noise, of the type tested on the C.I.S.P.R.-meter circuit, is in conformity with the square root of the bandwidth analogous to true fluctuation noise. Assuming the same average magnitude of transients for both types of noise, the difference in reading will merely be a constant factor owing to the spells of silence in the case of motor noise.

This law for motor interference was first found empirically as mentioned in the introductory paragraphs of Section (6), and it appears that the conclusions drawn from the results of oscillographic investigations corroborate these findings.

7. Conclusions

Some aspects of radio interference were investigated and the following results were obtained :—

1. It appears that, with fluctuation and motor noise, the effective bandwidth of a band-pass amplifier containing more than one coupled filter, can be determined on the basis of 3-db attenuation on the overall response curve.

2. Fluctuation noise follows quite closely the expected square root of bandwidth law, if bandwidth is defined on the 3-db basis.

3. The crest values of band-pass filter transients initiated by single impulses agree with those of idealized filters to within ±3 db.

4. The bandwidth calculated from the duration of the main lobe of the response transient agrees fairly well with the equivalent noise bandwidth ; viz., that defined on the 3-db basis.

5. The steady meter reading due to repeated impulses agrees with the expected behaviour, subject to an error increasing with decreasing $n/\Delta f$, whereby the nature and general trend of this error is fully accounted for by the inherent properties of the measuring apparatus.

6. Motor noise follows the square root of bandwidth law within the range investigated.

Acknowledgment

The work described in this paper was carried out by the author while he was on the staff of the British Electrical and Allied Industries Research Association. It is a part of the Association's research programme on radio interference, and the author gratefully acknowledges the Director's permission to publish.

APPENDIX I

It can easily be shown that the response curve of a critically coupled filter consisting of two identical tuned circuits each independently of $Q = \omega L/r$ is given by the equation—

$$Z = \frac{L}{2rC} \frac{1}{1 - \frac{\epsilon^2}{2} + j\epsilon}$$

where—

L/rC is the resonance impedance of each tuned circuit, and

$$\epsilon = \left(\frac{f}{f_m} - \frac{f_m}{f}\right) Q \approx \frac{2(f - f_m)}{f_m} Q = \frac{\Delta f}{f_m} Q$$

Hence—

$$|Z| \propto \frac{1}{\sqrt{\left(1 - \frac{\epsilon^2}{2}\right)^2 + \epsilon^2}} = \frac{1}{\sqrt{1 + \frac{\epsilon^4}{4}}} = R$$

where R stands for response.

If n such filters are either successively loosely coupled to each other or used as interstage coupling networks of an amplifier as in the present case, the total response is given by—

$$R_n = \left[\frac{1}{\sqrt{1 + \frac{\epsilon^4}{4}}}\right]^n = \left(1 + \frac{\epsilon^4}{4}\right)^{-\frac{n}{2}}$$

In order to find the equivalent noise bandwidth Δf_N the total response has to be squared and integrated over the relative frequency deviation ϵ from zero to plus infinity and divided by $R_n{}^2{}_{max}$. Since for $\epsilon = 0$, $R_n = 1$,

$$\epsilon_N = \int_0^{+\infty} \frac{d\epsilon}{\left(1 + \frac{\epsilon^4}{4}\right)^n} = \left(1 - \frac{1}{4}\right)\left(1 - \frac{1}{8}\right)\left(1 - \frac{1}{12}\right) \cdots \left(1 - \frac{1}{4(n-1)}\right)\frac{\pi}{2}$$

where $\epsilon_N = \Delta f_N Q/f_m$. If n is great

$$\epsilon_N = \sqrt{2} \times 0.906/n^{\frac{1}{2}} = 1.28/n^{\frac{1}{2}}.$$

The squared response curves are plotted against half the relative bandwidth in Fig. 6 and the respective equivalent noise bandwidths, as computed from the last two formulæ, are indicated as vertical lines. Their intersection with the response curves corresponds to certain values of attenuation on the respective response curves and can be read off the diagram by converting the ordinate values of the intersections into decibels or they can be calculated by substituting the noise bandwidth in the expression for the response curve. Thus—

attenuation =

$$10\, n \log \left\{ 1 + \frac{1}{4} \left[\frac{\pi}{2} \prod_{\nu=2}^{\nu=n} \left(1 - \frac{1}{4(\nu - 1)}\right) \right]^4 \right\}$$

or, if the number n of filters tends towards larger and larger values, this expression becomes—

$$10n \log\left\{1+\frac{1}{4}\left(\frac{1.28}{n^{\frac{1}{4}}}\right)^{4}\right\} = 10n \log\left\{1+\frac{0.692}{n}\right\}$$

$$= 10n\,\frac{0.692}{n}\,0.434 = 3.$$

In other words, the bandwidth defining attenuation along the total response curve tends towards the definite value of 3 db if more and more filters are cascaded.

This trend is also borne out in the Table 4 computed from the foregoing formulæ. The squared response curves for n cascaded filters and their respective equivalent noise bandwidths are depicted in Fig. 6.

It is interesting to note that the attenuations corresponding to the true-noise bandwidth vary so little from the 3-db value so that one can stipulate the noise bandwidth Δf_N to be the bandwidth measured on the total response curve at 3-db attenuation without committing an error of more than + 3.1 per cent for two cascaded filters, or 1.5 per cent in the case of three, or 0.6 per cent for four filters, this error rapidly decreasing for greater n. This is a very convenient result for it enables the experimenter to find the equivalent noise bandwidth of a receiver or amplifier on this 3-db basis by means of a simple measurement with the signal generator without going to the trouble of plotting the whole response curve, squaring it, determining the area by some method and calculating the equivalent rectangle.

Appendix (1.1) (*Communicated by S. Whitehead*)

Proof of the integration involved in the calculation of the equivalent noise bandwidth in Appendix I.

$$I_n = \int_{-\infty}^{+\infty} \frac{dx}{\left(1+\frac{x^4}{4}\right)^n}$$

$$I_n - I_{n+1} = \frac{1}{4}\int_{-\infty}^{+\infty} x\,\frac{x^3\,dx}{\left(1+\frac{x^4}{4}\right)^{n+1}}$$

and integrating by parts.

$$I_n - I_n + 1 = -\frac{1}{4}\left\{\left[\frac{x}{n\left(1+\frac{x^4}{4}\right)^n}\right]_{-\infty}^{+\infty} - \int_{-\infty}^{+\infty}\frac{dx}{n\left(1+\frac{x^4}{4}\right)^n}\right\} = \frac{1}{4n} I_n$$

Thus—

$$I_{n+1} = \frac{n-\frac{1}{4}}{n} I_n$$

By contour integration it can be seen that—

$$I_n = \pi$$

whence—

$$I_n = \frac{\left(1-\frac{1}{4}\right)\left(2-\frac{1}{4}\right)\ldots\left(n-1-\frac{1}{4}\right)}{(n-1)\,!}\,\pi$$

$$= \left(1-\frac{1}{4}\right)\left(1-\frac{1}{8}\right)\left(1-\frac{1}{12}\right)\ldots\left(1-\frac{1}{4(n-1)}\right)\pi$$

as $n \rightarrow \infty$

$$I_n = \frac{\pi \text{ Lim}}{n\rightarrow\infty}\prod_{\nu=1}^{\nu=n}\left(1-\frac{1}{4\nu}\right) = \frac{\pi}{\left(-\frac{1}{4}\right)!}\,n^{-\frac{1}{4}}$$

$$(\ddagger) \quad = \frac{4}{n^{\frac{1}{4}}}\left(\frac{1}{4}\right)!\,\sin\frac{\pi}{4} = \frac{2\sqrt{2}}{n^{\frac{1}{4}}}\left(\frac{1}{4}\right)! = \frac{2\sqrt{2}\times 0.906}{n^{\frac{1}{4}}}$$

‡ See Jahnke & Emde Tables of Functions, Chapter on Factorial Function.

TABLE 4

n	1	2	3	4	8	16	32
ϵ_N	1.57	1.178	1.03	0.946	0.776	0.643	0.539
Attenuation (db) for equivalent noise bandwidth Δf_N	4	3.4	3.21	3.12	3.10	3.05	3.01
Deviation (per cent) of true-noise band width from 3-db bandwidth . .	10	3.1	1.5	0.6	—	—	—

APPENDIX II.

Pulse Spectrum.

It can be shown that the spectrum of a pulse of the form of Fig. 12(a) integrated over a passband Δf around a mid-frequency f_m gives a transient of the shape shown in Fig. 8, but of a crest value given by—

$$V_{T1} = \frac{2\Delta f}{\sqrt{\frac{1}{\tau^2}\quad\left(2\pi f_m\right)^2}}$$

where τ is the time constant of the trailing edge of the pulse. It is evident that, for a time constant τ large compared with $1/f_m$ the crest value becomes—

$$V_{T2} = \frac{\Delta f}{\pi f_m}$$

the same as for a perfect unit step. The ratio of the two equations gives a quotient whose deviation from unity represents the error due to replacing unit step by the experimental exponential pulse. Thus—

$$\frac{V_{T1}}{V_{T2}} = \frac{2\Delta f\,\pi f_m}{\Delta f\sqrt{\frac{1}{\tau^2} + \left(2\pi f_m\right)^2}}$$

$$= \frac{1}{\sqrt{1 + \left[\frac{1}{2\pi f_m\tau}\right]^2}} = \frac{1}{\sqrt{1 + \left[\frac{1}{\omega_m\tau}\right]^2}}$$

Some values are tabulated in Table 5 and show that, for the pulse used in the experiments, the error is the order of 0.2 db, since the time constant τ of the trailing edge is about 0.15 μsec as can be estimated from the oscillogram of Fig. 11 (a).

TABLE 5

f_m = 5 Mc/s			
τ (μsec)	$\omega_m\tau$	V_{T1}/V_{T2}	Error (db)
0.0318	1	0.707	3
0.0636	2	0.894	1
0.1	3.14	0.953	0.44
0.2	6.28	0.9875	0.108
0.15	4.7	0.978	0.195

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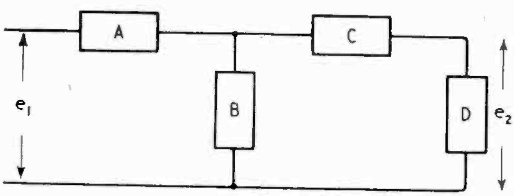
CORRESPONDENCE

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.

Network Mnemonic

SIR,—The mnemonic referred to below was devised and published some years ago in the text of a Patent Specification, where it is no doubt lost to sight. This note is intended to draw attention to it and to the formula it serves to bring to mind.



The figure illustrates an "unbalanced" network (the solution for the "balanced" case will be clear from this). The formula in question is that giving the loss-ratio e_2/e_1. It will be found much simpler to deal with e_1/e_2: we find that :—

$$e_1/e_2 = 1 + A/D + A/B + A/C + C/D + (A.C)/(B.D) \qquad \text{(a)}$$

This formula can be memorized if the symbols (1), (2), (3), (4) are substituted for the letters A, B, C, D. Note that the numbers refer to the order of the elements taken from left to right, the direction of the flow of power. Making this change we get :—

$$e_1/e_2 = 1 + (1)/(4) + (1)/(2) + (3)/(4) + (1).(3)/(2).(4) \qquad \text{(b)}$$

It will be seen that the 2nd, 3rd, and 4th terms resemble the series of fractions $\frac{1}{4}$, $\frac{1}{2}$, $\frac{3}{4}$, and that the 5th term is the product of the 3rd and 4th.

Once this has been memorized it is easy to write at sight the formula for the case in which the elements "B" and "D" are each made up of a pair of impedances in parallel. For suppose that "B" is to be replaced by "B_1" and "B_2" in parallel: then the 3rd term in (a) is duplicated into A/B_1 and A/B_2 and a suitable change made to the 5th term, replacing $1/B$ by $1/B_1 + 1/B_2$. And so for D_1 and D_2.

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Bournemouth, Hants.

Signal-Noise Ratio in Radar

SIR,—In your December issue Mr. M. Levy discussed the problem of the detection of a recurrent positive or negative pulse against a background of unrectified random noise, when displayed as a lateral deflection relative to a linear time base on an oscilloscope. The author assumed that the minimum detectable amplitude of the pulse is dependent on the observation of changes of brightness along the ideal line of scanning. Actually the presence of the signal is detected by observation of a bodily displacement of part of the noise pattern from this ideal line. Since in this type of display both the upper and lower edges of the pattern fluctuate and have no distinct contours, the signal is detected by the observation of a kink in the originally straight line of symmetry of the pattern. In determining the shape of this line a subjective sense of symmetry is applied to the whole visible height of the pattern and not to the gradient of brightness near its middle as assumed in the paper.

The fallacy of the latter assumption may be shown by the following example. Suppose that the electron beam is defocused into a uniformly-illuminated disc of diameter d and that the noise level is negligible. In the absence of signals the scanned pattern will then appear as a uniformly bright band of height d. If the criterion for detecting signals were a change of brightness along the ideal scanning line, which is the middle line of this band, any signals of amplitude smaller than $\frac{1}{2}d$ would not be detected, which is obviously not the case. In the second method of display discussed by the author, a part of the pattern, below or above a given level, is reduced by the action of a limiter to a distinct line of variable brightness parallel to the scanning line. In this case only is a change of brightness material to the detection of the signal.

Since the two methods of detection depend on different criteria of observation, they cannot be compared by calculation of the distribution of brightness in both cases, as is done in the paper. Thus, the advantage of the method of display suggested by the author remains unproved. It appears that a conclusive comparison of the two methods could be obtained only by means of a number of subjective observations in which various additional conditions of experiment were carefully taken into account (definition and brightness of the beam, colour and persistence of fluorescence of the screen, rate of scanning, etc.).

In the introductory part of the paper the author refers to Eqn. (1) as the law of distribution of "the peak amplitudes in the noise," while this and all the following equations actually apply to the distribution in time of the instantaneous values of the noise voltage. The difference between these two conceptions may best be demonstrated by indicating that the probability of the amplitude of the noise voltage being zero is infinitely small, while the probability of the instantaneous noise voltage being zero is maximum. It may be interesting to note that a similar error of terminology occurred some years ago in the fundamental paper by V. D. Landon: "The Distribution of Amplitude with Time in Fluctuation Noise" (*Proc. Inst. Radio Engrs*, 1941, Vol. 29, p. 50). In a lively polemic with K. A. Norton which followed (*Proc. Inst. Radio Engrs*, 1942, Vol. 30, p. 425), the error was fully clarified.

In Eqn. (5) three typographical errors should be corrected. The upper limits of integration in the last three integrals of this equation should be α, α and $k\alpha$ respectively.

S. DE WALDEN.

Haslemere, Surrey.

Positive and Negative Frequencies

SIR,—I have been much impressed by the important points raised by G. B. Madella[1] in a letter appearing in your October issue in which he discusses a proposed system of filtering by modulation[2]. In the mathematical treatment of frequency, negative values of frequency constantly present themselves, but one's usual mental picture of a pure frequency as being a sinusoidal variation in time is quite inadequate to represent a negative frequency. Thus a carrier frequency of 1000 c/s when modulated at 10 c/s produces sidebands at 1,010 and 990 c/s.; if modulated at 1,010 c/s. it ought logically to produce sidebands of 2,010 and minus 10 c/s. The matter is of more than theoretical interest now that D. G. Tucker[3] has designed a highly selective transmission-measuring equipment in which filtering is attained by modulation, and has described[4] a new type of radio receiver which selects the incoming signal by a modulation process.

Signor Madella points out that polyphase theory provides a clear representation of the sign of a frequency in terms of the phase sequence of the supply. It may be noted that the mathematical treatment of frequency is essentially polyphase since the complete spectrum of a complicated signal is two spectra, one of sine and one of cosine terms. Would it not be more useful to think of a pure frequency in terms of a vector voltage rotating with constant angular velocity? A vector fluctuating in any manner can always be regarded as the sum of a sufficient number of component vectors of constant amplitude rotating with various constant angular velocities, and the frequency spectrum in this sense extends from plus to minus infinity as the mathematics requires. When such a vector is referred to axes rotating with frequency f it is clear that all the component frequencies are reduced algebraically by an amount f and this is the effect aimed at in polyphase modulation. It is worth noting that a pure frequency f_0 in single phase may be regarded as the sum of two vector voltages rotating in opposite senses, that is with frequencies $+ f_0$ and $- f_0$. If referred to axes rotating with frequency f the new frequencies $(-f \pm f_0)$, numerically equal to sum and difference frequencies, arise in an obvious way.

As regards networks or filters for polyphase alternating current it should be observed that a network will, in general, show a different transmission coefficient when the phase sequence of the supply is reversed: as an example, it is possible to construct a simple delta bridge which will transmit a 50-c/s three-phase voltage when the phase sequence is 1, 2, 3 but will transmit nothing if the supply has a phase sequence 3, 2, 1. This is not surprising if it is understood that a reversal of sequence changes the sign of the frequency, for the bridge may then be expected to have a transmission characteristic which varies over the whole range of frequency from plus to minus infinity and which will not in general be the same in the positive and negative ranges.

These new points of view due to G. B. Madella have been more fully discussed by him in two articles in *Alta Frequenza*[5].

N. F. BARBER.

Teddington, Middx.

[1] G. B. Madella, "Single-phase and Polyphase Filtering Devices using Modulation." Letter to *Wireless Engineer*, October 1947, Vol. 24, p. 310.

[2] N. F. Barber, "Narrow Band-Pass Filter using Modulation," *Wireless Engineer*, May 1947, Vol. 24, p. 132.

[3] D. G. Tucker, "Highly-Sensitive Transmission-Measuring Equipment for Communication Circuits," *J. Instn elect. Engrs*, Vol. 94, Part III, p. 211, 1947.

[4] D. G. Tucker "The Synchrodyne: A New Type of Radio Receiver for A.M. Signals, *Electronic Engineering*, March, August and September, 1947.

[5] G. B. Madella, *Alta Frequenza* XIII, p. 31, p. 132, 1944.

CORRECTION.

In the letter from A. S. Gladwin on p. 34 of the January issue, the expression $\omega^2/\omega_0^2 = \Sigma V_n^2/n^2 \Sigma V_n^2$ should read $\Sigma(V_n^2/n^2)/\Sigma V_n^2$.

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.

AERIALS AND AERIAL SYSTEMS

586 578.—Aerial system comprising groups of dipoles arranged in circular tiers, and capacitance-loaded, to operate over a wide frequency spectrum.
E. C. Cork. Application date 2nd May, 1944.

587 071.—Aerial system for the transmission of two sets of waves having the same frequency, but polarized in two orthogonal planes.
Marconi's W.T. Co. Ltd. (assignees of G. H. Brown). Convention date (U.S.A.) 31st March, 1943.

587 677.—Supporting and energizing a system of loop aerials, each consisting of four semi-cylindrical radiating elements.
Standard Telephones and Cables Ltd., (assignees of A. G. Kandoian) Convention date (U.S.A.) 7th October, 1942.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

587 185.—Radiolocation equipment for artillery work, in which the gain factor is cyclically varied in order to permit the observation of shell-splashes and other comparatively weak echo signals.
Western Electric Co., Inc. Convention date (U.S.A.) 1st October, 1943.

587 351.—Radiolocation system in which two spaced land-stations, when interrogated by a mobile craft, radiate pulsed signals which indicate their respective distances from the craft.
F. C. Williams. Application date 15th September, 1943.

587 389.—Portable radiolocation system in which the echo-signal from a pulse-modulated exploring wave gives a beat-frequency indication of distance against a time-base synchronized with the pulsing-frequency.
A. G. Pocock. Application date 18th December, 1939.

587 390.—Radiolocation system in which each exploring pulse is initiated, through a saw-toothed voltage relay, by the preceding echo-signal, the distance of the target then being a function of the instantaneous value of the saw-toothed voltage.
J. Forman and Pye Ltd. Application date 4th July, 1940.

587 464.—Pulse-generating system, utilizing the shock excitation and damping of tuned circuits, particularly for radiolocation.
Western Electric Co., Inc. Convention date (U.S.A.) 13th May, 1943.

587 502.—Phase-correcting devices for combining the signals from spaced directive aerials with those from a sense-determining aerial.
Standard Telephone and Cables Ltd., and L. J. Heaton-Armstrong. Application date 28th November, 1944.

587 517.—Radiolocation equipment of the kind in which the echo-signals are presented to one or other of two selected time bases, suitable for close and long-range observation, respectively.
E. G. Bowen and A. G. Touch. Application date 9th March, 1943.

587 543.—Directive aerial of the rhombic type in which a dissipating loop, of the same general shape as the radiating loop, is arranged below it and close to the ground.
Press Wireless Inc. Convention date (U.S.A.) 15th June, 1943.

587 562.—Biased limiter-device for reducing interference, particularly the jamming of radiolocation observations by continuous oscillations of a relatively-low frequency.
A. D. Blumlein. Application date 31st March, 1941

587 592.—Radiolocation system in which distance is measured by identifying the echo-signal on the screen of the c.r. indicator with one or other of a series of local impulses generated at intervals by a steadily-rising voltage.
Western Electric Co., Inc. Convention date (U.S.A.) 5th October, 1943.

587 655.—Thermionic - valve counting - circuit, particularly for distinguishing the echo-signals in radiolocation from interfering pulses of a different frequency.
S. M. Taylor. Application date 30th December, 1944.

587 773.—Radiolocation system in which a frequency-modulated signal is used to interrogate and elicit a frequency-modulated response from a distant craft or point.
Bendix Aviation Corp. Convention date (U.S.A.) 15th October, 1942.

RECEIVING CIRCUITS AND APPARATUS
(*See also under Television*)

587 097.—Tuning and volume control device, combined with a pawl-and-ratchet motion for operating a wave-change switch.
E. R. Robinson and Masteradio Ltd. Application date 3rd January, 1945.

587 155.—Regulating the power-supply to a thermionic load, particularly when a number of different signal-carrying circuits are coupled to a common impedance.
Standard Telephones and Cables Ltd., (assignees of P. S. Christaldi). Convention date (U.S.A.) 26th November, 1942.

587 212.—Frequency - discriminating circuit, particularly for receiving f.m. signals that contain modulation elements comparable with the carrier-frequency.
Standard Telephones and Cables Ltd., (assignees of J. S. LeGrand and E. La Bossiere). Convention date (U.S.A.) 20th December, 1943.

587 222.—Tuning system for a valve oscillator

of the kind in which two branched circuits, of the transmission-line type, are used to couple the anode and grid.

Hazeltine Corpn., (assignees of L. R. Malling). Convention date (U.S.A.) 1st February, 1944.

587 274.—Controlling the speed and synchronism of the scanning and recording devices used in receiving facsimile telegraphy.

Marconi's W. T. Co., Ltd., (assignees of F. C. Collings Jr). Convention date (U.S.A.) 30th November, 1942.

587 319.—Utilizing negative reaction to stabilize a crystal-controlled valve oscillator.

Standard Telephones and Cables Ltd., (assignees of G. T. Royden). Convention date (U.S.A.) 9th November, 1943.

587 627.—Arrangement of an amplifier-stage at the centre of a dipole aerial in order to minimize the interference normally picked up by the down-lead.

D. Jackson and Pye Ltd. Application date 9th September, 1944.

587 638.—Clockwork mechanism with a time dial and two additional pointers which can be set to ensure the automatic switching-on of a radio-set at a given time and tuned to a selected wave-length.

G. Lan. Convention date (Canada) 15th April, 1944.

TELEVISION CIRCUITS AND APPARATUS

For Transmission and Reception

587 125.—Television receiver in which the action of the scanning beam momentarily volatilizes the sensitive coating of the screen, and so varies its transparency to an external source of light.

Co. Para La Fabrication De Contadores y Material Industrial S. A. and P. Viteau. Application date 22nd June, 1944.

587 370.—Circuit for inverting the light and shade contrasts, when producing a television picture from a photographic negative, without falsifying the relative half-tone values.

Cinema Television Ltd., and T. C. Nuttall. Application date 18th October, 1944.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

586 315.—Secret radio-communication system in which frequency-modulated signals are subjected to a narrow-swing modulation and are mixed with a masking signal having a wide swing.

Sperry Gyroscope Co. Inc. (assignees of W. W. Hansen). Convention date (U.S.A.) 10th April, 1943.

586 585.—Signalling system in which trains of pulses, modulated in time or amplitude, are transmitted through the ground by earthed electrodes, stated to be substantially immune from deliberate jamming.

Marconi's W.T. Co. Ltd. (assignees of C. W. Hansell). Convention date (U.S.A.) 24th June, 1942.

586 699.—Two-way set comprising two oscillators which are interlocked to ensure that both the transmitter and receiver work on the same signal-frequency.

J. Ruston and The Plessey Co. Ltd. Application date 10th November, 1944.

586 734.—Facsimile signalling system in which the shade distortion caused by static or like interference is automatically compensated.

Press Wireless Inc. Convention date (U.S.A.) 22nd September, 1943.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

586 275.—Construction of the resonator electrode in a tube of the velocity-modulation type, and a movable metal vane for tuning it.

Standard Telephones and Cables Ltd. and S. G. Tomlin. Application date 4th December, 1942.

586 351.—Construction of a thermionic valve of the type in which a sealed-in anode-disk divides the valve into two portions.

Standard Telephones and Cables Ltd., C. N. Smyth and A. S. Wade. Application date 8th December, 1943.

586 430.—Construction of velocity-modulating tube of the magnetron type to facilitate coupling it to a waveguide input or output.

J. Sayers, C. S. Watt and C. S. Wright. Application date 8th June, 1943.

586 511.—Construction and arrangement of the cavity-resonator in a discharge-tube of the velocity-modulation type.

The British Thomson-Houston Co. Ltd., W. J. Scott, and C. J. Milner. Application date 3rd July, 1942.

586 631.—Electrode arrangement for a velocity-modulation tube, designed to minimize intermediate-frequency "noise" when used as a local-oscillator in reception.

Standard Telephones and Cables Ltd. and J. H. Fremlin. Application date 19th April, 1943.

586 974.—Construction and assembly of the electrode system of high-voltage rectifiers, and like discharge devices, housed in a tubular envelope with flat-ends.

The M-O Valve Co. Ltd. and A. H. Howe. Application date 2nd August, 1944.

SUBSIDIARY APPARATUS AND MATERIALS

586 515.—Screened oscillation-generator, of the grounded-grid type, with anode and cathode coaxial-line circuits, particularly for use as a radio altimeter.

Standard Telephones and Cables Ltd. and F. H. Taylor. Application date 12th January, 1944.

586 714.—High-impedance device for preventing the escape of high-frequency energy from the gap between a lead-in conductor and the wall of a resonator or screen.

C. S. Bull. Application date 4th May, 1943.

587 020.—Construction and assembly of a dry-contact rectifier of the selenium type.

Standard Telephones and Cables Ltd. (assignees of M. F. Skinner, I. P. Denyssen and M. G. Holmes). Convention date (U.S.A.) 26th October, 1943.

587 045.—Attenuation-line for centimetre waves made by embedding a wire helix in a moulded mass of finely-divided magnetic material.

H. F. Garrett and Aladdin Radio Patents, Ltd. Application date 10th October, 1944.