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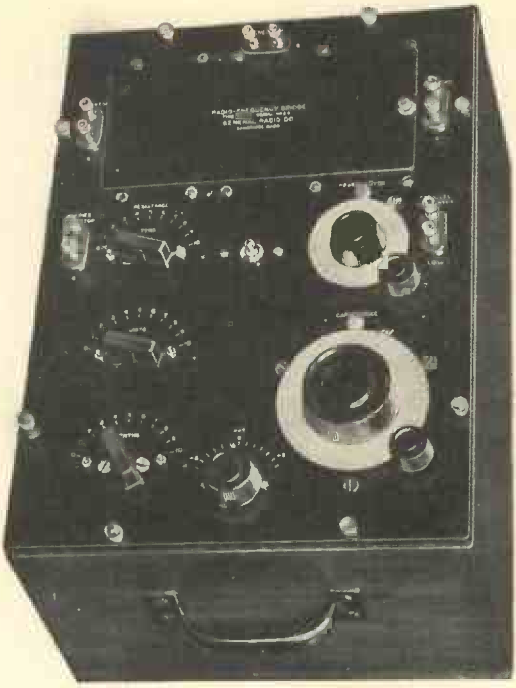

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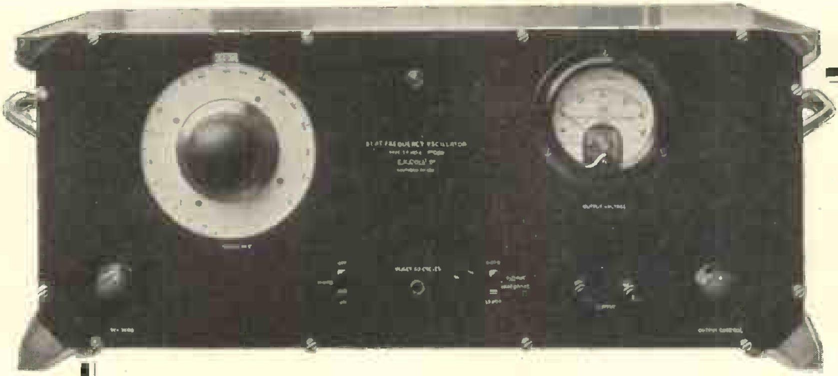

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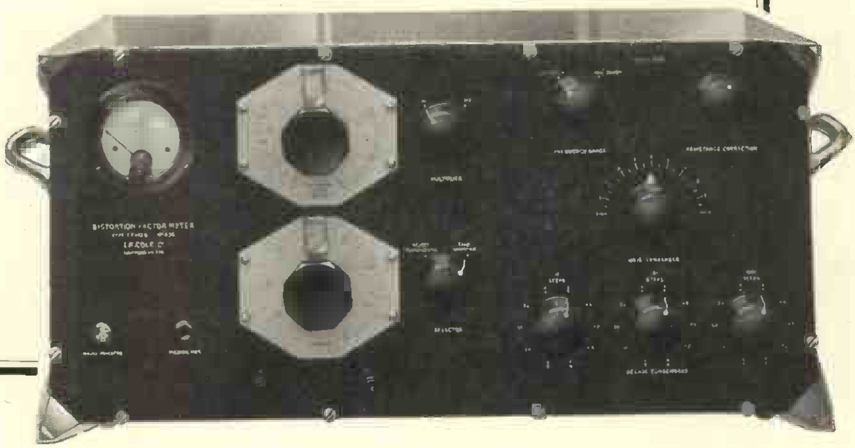

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THE
WIRELESS ENGINEER

VOL. XIII. DECEMBER, 1936. No. 159

Editorial

A New High Efficiency Power Amplifier for Broadcast Transmitters

WITH the ever-increasing power employed in broadcast transmitters, the efficiency of the final stage has become a matter of considerable importance. It is not merely the cost of the power, but also of the transformers, rectifiers, cooling equipment, etc., which is increased by the inefficiency of the amplifier. The ordinary linear radio-frequency power amplifier cannot be operated on the unmodulated carrier at an efficiency higher than about 33 per cent. without endangering the linearity of the peak amplification on 100 per cent. modulation. This means that the all-day efficiency will not greatly exceed 33 per cent., so that of a 150 kilowatt d.c. supply, 100 kilowatts are wasted and must be removed by the cooling equipment.

Several methods of obtaining an improved efficiency have been devised, such as the class B modulation system employed at the 500 kW. transmitter WLW, and the "outphasing" method of Chireix employed at several French stations. A new method has been devised by W. H. Doherty, of the Bell Laboratories, and was described by him in a paper read in May at Cleveland, Ohio.* This method allows the efficiency to be increased to over 60 per cent. while still retaining the advantages of the simple linear amplifier, such as modulation at a low power level, thus eliminating high-power audio equipment.

With the ordinary linear amplifier biased to the cut-off point, anode current flows during the positive half waves of applied grid voltage. If the pulses of anode current have a maximum value of $\hat{\imath}$, and the corresponding peak value of the anode voltage be $\hat{v}$, the power output of the valve will be equal to $\hat{v}\hat{\imath}/4$. The average value of the anode current is $\hat{\imath}/\pi$, and the power input therefore $V_0\hat{\imath}/\pi$. The efficiency would then be $\frac{\pi}{4}\cdot\frac{\hat{v}}{V_0}$. Since at its maximum swing with 100 per cent. modulation $\hat{v}/V_0$ should not exceed 0.9, the efficiency at zero modulation does not exceed about 33 per cent. At 100 per cent. modulation it increases to about 50 per cent., but this extreme modulation occurs during a relatively small proportion of the total programme time, and the efficiency taken over the whole programme time cannot be very much greater than 33 per cent.

In the new system the operation of the valve is normal when the amplitude of the

* *Proc. I.R.E.*, vol. 24, page 1163, Sept., 1936.

high frequency current is less than that of the unmodulated carrier, but when it exceeds this value—which it must do during a half of the audio cycle—the operation is abnormal in that, instead of the a.c. anode voltage increasing linearly it is automatically

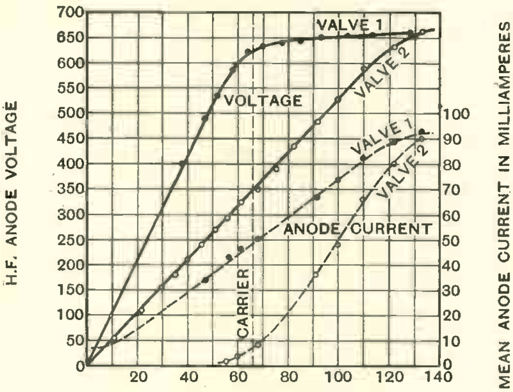


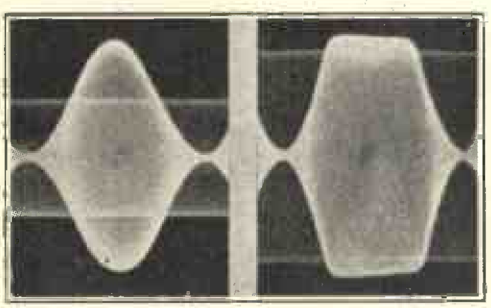
Fig. 1.

limited to the carrier value. Although the voltage does not increase, the anode current continues to increase in amplitude right up to the peak of the 100 per cent. modulated wave. This is achieved by reducing the anode load resistance into which the valve works; if the value of this was R for amplitudes below that of the carrier, it is automatically reduced as the amplitude increases until at the peak it is only $R/2$. Hence the anode current increases linearly throughout; over the first half due to an increasing anode voltage, R being constant, and over the second half due to a decreasing R, the voltage remaining constant. Since the anode voltage never exceeds the steady carrier value, the valve can be adjusted to work at maximum efficiency under these conditions, without any consideration of modulation peaks. This is illustrated in the voltage and current curves of valve 1 in Fig. 1, and in the right-hand oscillogram of the anode voltage in Fig. 2.

Beyond the carrier point, since the anode voltage is constant and the valve is working into a decreasing load resistance, it is not necessary for the grid voltage to increase linearly; instead of the normal 100 per cent. increase from the carrier point to the 100 per cent. modulation peak, an increase of about 40 per cent. in the grid voltage is found to be sufficient and special means are introduced to modify the grid voltage.

Two points will have occurred to the reader; firstly, the output of the valve which was increasing as the square of the current and voltage up to the carrier point, increased directly as the current beyond the carrier point, so that at the 100 per cent. modulation peak the output was only half of what it would have been; secondly, how can the system be given these assumed characteristics? Fortunately, the ingenious method devised by Doherty for limiting the anode voltage of the valve to its carrier point value automatically supplies the missing power.

A second valve is employed which is so biased that it is inoperative at amplitudes below the carrier value. For larger amplitudes it performs a double function, since it lowers the effective load resistance on the first valve in the desired manner, and in so doing supplies power to the load to compensate for the reduced output of the first valve. Its grid and anode voltages increase linearly throughout, but its current, which was zero until the carrier amplitude was reached, then increases so rapidly that at the 100 per cent. modulation peak its output is equal to that of the first valve. The combined output of the two valves has thus always the normal value. The voltage and current characteristics of the second valve



Valve 1. *Valve 2.*

Fig. 2.—*Anode voltage oscillograms for 100 per cent. modulation.*

are also shown in Fig. 1 and an oscillogram of the anode voltage on the left of Fig. 2.

The principle of the method can be explained with the aid of Fig. 3. Up to the carrier amplitude the right-hand valve is inoperative and need not be considered. The left-hand valve supplies the load $R/2$ through a special network consisting of a series inductance and two shunt condensers, each of which has an impedance of R ohms. It can easily be shown that this is equivalent

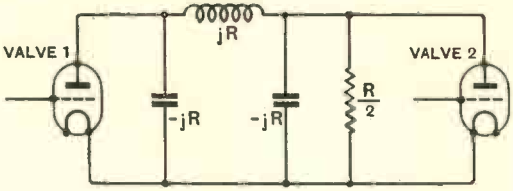


Fig. 3.

to a load resistance of $2R$ on the valve. As soon as the carrier amplitude is exceeded the right-hand valve becomes operative and supplies power to the load, but it should be remarked that since the network between the first valve and the load shifts the phase by 90°, it is necessary to produce a 90° phase difference between the two grid excitations. This is done by means of a somewhat similar network which is not shown. The supply of power from the second valve is equivalent to an increase in the resistance across the network, but the network being what is known as an impedance-inverting network, an increase in its output resistance causes an apparent decrease in the load resistance of the left-hand valve. By proper adjustment it can be arranged that at the 100 per cent. modulation peak the two valves are working in parallel supplying the same amount of power into the load $\frac{R}{2}$ which is therefore equivalent to each valve having its own load resistance of R ohms.

Tests on a 50 kilowatt transmitter have shown that the total power required, including all auxiliaries, which would have been about 230 kilowatts with the ordinary type, can be reduced to about 135 kilowatts by employing this system. G. W. O. H.

VALVE ASSEMBLY

The vastness of the Mazda valve assembly section at the Edison Swan Electric Company's works at Brimsdown can be judged from this picture.

The Behaviour of the Output Circuit to Transients*

By N. W. McLachlan

I. Introduction

THE design of an output transformer used between the power valve and loud speaker is now well established. To reproduce the bass register without undue attenuation a fairly large primary inductance is needed, whilst the magnetic leakage and the self and mutual capacitances of the windings must be very small if the highest audio frequencies are to be retained. A transformer worked below the knee of the magnetisation curve, and complying with the above conditions will give a uniform response characteristic for sine wave currents over a wide frequency range, provided the secondary load is entirely resistive.

It is often assumed that any device which has a flat response characteristic (for sine wave input) over a wide frequency range, will reproduce transients with but little distortion. Are we justified in making this assumption where an output transformer and its associated power valve are concerned? The object of the present article is to consider this question.

2. Circuit Arrangement.

The usual type of single valve output circuit, with a transformer whose core has a short air-gap to make the primary inductance independent of the valve feed current, is illustrated in Fig. 1. The equivalent circuit

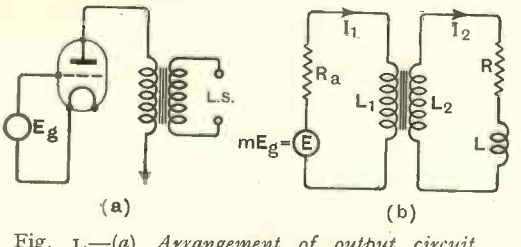


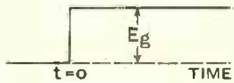
Fig. 1.—(*a*) *Arrangement of output circuit.* (*b*) *Equivalent circuit of Fig. 1a.*

of the loud speaker for all frequencies is rather complicated. To effect simplification, therefore, we shall confine our study to medium and high audio frequencies, where as a first approximation the electrical impedance of the speaker can be regarded as a resistance in series with an inductance. In Fig. 1, R_a = internal A.C. valve resistance of power valve; $L_1 L_2$ = inductance of primary and secondary windings of transformer, respectively; R = resistance of loud speaker in operation plus resistance (A.C.) of secondary winding; L = inductance of loud speaker in operation plus magnetic leakage between windings (assumed to be small). The self and mutual capacitances of the windings are taken as being negligible owing to sectionisation.

3. Transient of Form in Fig. 2.

This form of transient is known as Heaviside's unit function. At time $t = 0$ a D.C.

Fig. 2.—*Illustrating Heaviside's unit function.*



voltage E_g is applied to the grid of the power valve. This appears in the equivalent circuit as $-mE_g = -E$, where m is the magnification factor of the power valve. By mathematical analysis, using Heaviside's operator $p = d/dt$ and Bromwich's contour integral,* it is found that the current in the secondary winding of the transformer, i.e., through the loud speaker, is

$$I_2 = \frac{E}{L}(n_2/n_1)\frac{1}{(a-b)} \cdot (e^{-bt} - e^{-at})\,; \qquad (1)$$

where $\left.\begin{matrix} a \\ b \end{matrix}\right\} = \alpha \pm \sqrt{\alpha^2/4 - \beta}\,;$

$$\alpha = R/L + R_a/L_1 + R_a L_2/LL_1\,;$$

$$\beta = RR_a/LL_1\,;$$

and n_2/n_1 = turns ratio of secondary to primary.

* MS. accepted by the Editor, April, 1936.

* Jeffrey's "Operational methods in Mathematical Physics," Chap. 2.

To illustrate the form of curve represented by (1) we take the following practical data † and calculate the current I_2: $R_a = 10^3$ ohms; $R = 10$ ohms; $L_1 = 20$ henrys; $L_2 = 0.08$ henry; $L = 10^{-3}$ henry; $n_2/n_1 = 16/1$. Inserting these numerical values in (1) we obtain

$$I_2 = 4.46 \times 10^{-3}E(e^{-35.7t} - e^{-1.4 \times 10^4 t}). \quad (2)$$

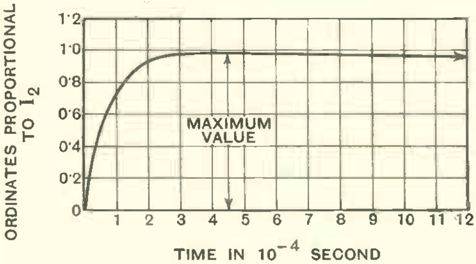


Fig. 3.—*Relationship between secondary current and time for circuit of* Fig. 1 *and transient of* Fig 2.

The form of relationship between I_2 and t is plotted in Fig. 3. The secondary current is not identical in form with that of the voltage applied to the grid of the power valve. It rises to a maximum value after a time lapse of 4.27×10^{-4} second, and then decays according to the law $e^{-35.7t}$. The initial portion of the current curve is linear, being given by the formula

$$I_2 = \frac{E(n_2/n_1)t}{L}. \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

From (3) it follows that the smaller the inductance L in the secondary circuit of the transformer, the more rapidly will the current reach its maximum value. If L were zero, the secondary current would be

$$I_2 = \frac{E(n_2/n_1)}{R + R_a\left(\frac{n_2}{n_1}\right)^2} e^{-t/\left(\frac{L_1}{R_a} + \frac{L_2}{R_2}\right)}; \quad \ldots \quad (4)$$

so it would reach the value $E(n_2/n_1)/[R + R_a(n_2/n_1)^2]$ immediately the transient was applied to the grid of the power valve, and then decay slowly. For a given value of L the time lag is reduced by increasing R_a, R and n_2/n_1. However, these reductions may not always be expedient in practice. Every effort should be made, therefore, to reduce the magnetic leakage as much as possible. The speaker inductance can be made small if either a separate unit is used for the upper audio frequencies, or only part of the driving coil is fed with current in this range.

The time taken for I_2 to reach its maximum value can be calculated from the formula

$$t = \frac{1}{a} \log_e a/b; \quad \ldots \quad \ldots \quad (5)$$

where $a \doteq \frac{R}{L} + \frac{R_aL_2}{LL_1} = 1.4 \times 10^4$, and $b = 35.7$. The maximum value of

$$I_2 \doteq \frac{E(n_2/n_1)}{L} \frac{e^{-bt}}{a}; \quad \ldots \quad \ldots \quad (6)$$

where t has the value given by (5).

$$\text{Thus } I_2\text{max.} \doteq \frac{E(n_2/n_1)}{L} \cdot \frac{e^{-\frac{b}{a}\log_e a/b}}{(R/L + R_aL_2/LL_1)};$$
$$= \frac{E(n_2/n_1)}{(R + R_aL_2/L_1)} (b/a)^{b/a};$$
$$= \frac{E(n_2/n_1)}{[R + R_a(n_2/n_1)^2]},$$
$$\text{very nearly} = K; \quad \ldots \quad (7)$$
$$\text{since } (b/a)^{b/a} \doteq 1.$$

This value is equal to that when $L = 0$ and $t = 0$, as will be seen from (4). Thus with

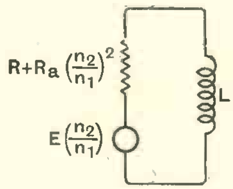


Fig. 4.—*Equivalent circuit for* Fig. 1 *under conditions specified in text.*

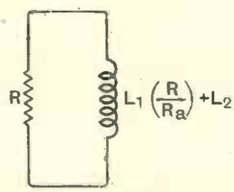


Fig. 5.—*Equivalent circuit for* Fig. 1 *under conditions specified in text.*

the given circuital constants there is no appreciable reduction in the maximum value of I_2 due to time lag. The circuit of Fig. 4 is approximately equivalent to that of Fig. 1 almost up to the time I_2 has its maximum value. Thereafter the equivalent circuit is approximately that shown in Fig 5, the current having the value given

† Provided the secondary load is resistive, i.e., $L = 0$, the response characteristic of an output circuit having these constants would be flat over a wide frequency range.

by (7), when it commences to decay according to the law $e^{-35.7t}$.

4. Transient of the Form $E_g e^{-at} \sin \omega t : L = 0$.

To simplify matters it is convenient first of all to discuss the case when there is no inductance in the secondary circuit of the transformer. The secondary current is given by

$$I_2 \doteq K[e^{-at} \sqrt{1 + \omega^2 c^2/(\omega^2 + a^2)^2} \sin\{\omega t + \tan^{-1} \omega c/(\omega^2 + a^2)\} - \frac{c}{\omega} e^{-ct}] ; \quad \dots \quad \dots \quad \dots \quad (8)$$

provided $a >> c$

where $c = RR_a/(RL_1 + R_a L_2) = 1/\left(\frac{L_1}{R_a} + \frac{L_2}{R}\right)$,

and the numerical values of the constants are those already stated. In practice $\omega >> c$, so (8) can be written

$$I_2 \doteq K[e^{-at} \sin \omega t - ce^{-ct}/\omega] ; \quad \dots \quad (9)$$

$$t > 0.$$

If $a = 0$ and $\omega >> c$, the applied "transient" is an undamped sine wave which starts at $t = 0$, so (9) becomes

$$I_2 \doteq K[\sin \omega t - ce^{-ct}/\omega) ; \quad \dots \quad \dots \quad (10)$$

$$t > 0.$$

From (9) and (10) we see that when $L = 0$ and $\omega >> c$, the output circuit does not introduce any appreciable distortion in the transient wave-form of the secondary current. In making this statement the exponential term ce^{-ct}/ω has been ignored, since it would not contribute anything of importance acoustically. The influence of this term is to reduce the initial rate of rise of the applied transient to a slight degree which is probably aurally insignificant. If the rate of rise were reduced considerably it might be a different matter, but the effect would depend upon the acoustical properties of the room where the sound is being reproduced.

At time $t = 0$ neither (9) nor (10) vanish, since they are approximate formulae derived from (8). If, however, $t = 0$ in (8) the current I_2 is zero. When t is very small the secondary current is given by

$$I_2 \doteq Kt, \quad \dots \quad \dots \quad \dots \quad (11)$$

which shows that the initial distortion is small, since $\frac{dI_2}{dt}$ from (11) is proportional to $\frac{d}{dt}(e^{-at} \sin \omega t)_{t \to 0}$.

When $t = 0$ in (8), the value of the exponential term is $-c/\omega$ which is much

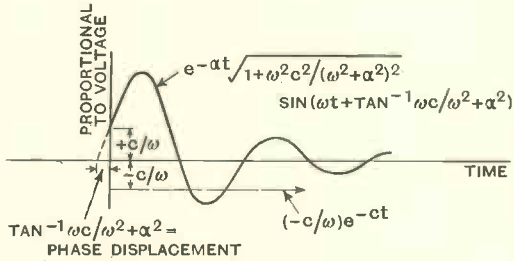


Fig. 6.—*Illustrating distortion of transient of form $e^{-at} \sin \omega t$ due to output circuit. The value of c/ω has been exaggerated for convenience of visualisation.*

less than unity. Since there is no applied voltage immediately before $t = 0$, the first term within brackets in (8) must have the value $+ c/\omega$, and this happens by virtue of the phase angle $\tan^{-1}\omega c/(\omega^2 + a^2)$. That is to say the oscillatory part of the transient is altered by an amount dependent upon this angle, as shown in Fig. 6. At high audio frequencies c/ω is negligible, so that both (9) and (10) will be represented quite accurately enough by their oscillatory terms alone, namely:

$$I_2 \doteq Ke^{-at} \sin \omega t ; \quad \dots \quad \dots \quad (12)$$

and $I_2 \doteq K \sin \omega t$; respectively. .. (13)

5. Transient of the Form $E_g e^{-at} \sin \omega t : L \neq 0$.

When L is not zero, the secondary current is given by an expression of the form

$$I_2 = E(n_2/n_1) \frac{\omega}{L} [Ae^{-at} + Be^{-bt} + Ce^{-at} \sin (\omega t + \theta)] ; \quad \dots \quad (14)$$

where a, b have the values stated in section 3, and A, B, C and θ are constants. Two exponential terms now appear instead of only one when L was zero. The first term, namely, Ae^{-at} decays very rapidly with increase in time, since a is so large. The second exponential term, Be^{-bt}, decays relatively slowly since $b <<< a$. The oscillatory part of the reproduced version of the original transient is given by the third term in (14). The influence of $L > 0$ is to modify the constant multiplier C and to introduce a larger phase angle θ, than that when $L = 0$. The initial part of the current curve is given by the formula

$$I_2 = E(n_2/n_1)\frac{\omega}{L}\frac{t^2}{2!}. \qquad \cdots \quad (15)$$

The slope of (15) is zero when $t = 0$, whereas that of the impressed transient is ωE_g. The initial zero slope is due to the presence of L in the secondary circuit. Thus the difference in wave-form between the impressed and reproduced transient depends largely upon the value of L, so it should be kept as small as possible in order to reduce distortion to a minimum.

Sounds can be divided broadly into two categories (*a*) steady, (*b*) transient. The latter are predominant in handclapping, footsteps, paper rustling, castanets, triangles and so on. Apart from the question of wave form, the basic frequencies of these sounds are high. We know from practical experience that the reproduced version of such sounds will not satisfy the ear unless the frequency range extends to 15,000 c/s. Even so we are unable to say in our present state of knowledge whether the ear can detect distortion of a transient when the alteration in wave-form is visible to the eye, although it may not appear to be serious. Since a loud speaker is used indoors, the ultimate wave-form which reaches the ear depends upon the acoustical properties of the room. Also in general non-linear effects and resonances may occur somewhere in the electro-acoustical system. Consequently, from an aural aspect the alteration in wave-form caused by a well-designed output stage may well be insignificant compared with that due to other influences.

There is an interesting field of research open to those who wish to study transient wave-forms in relation to aural perception. Until this research matures, all we can do is to indicate the visible differences between impressed and reproduced transient wave-forms. It may be remarked that in 1922 we were unable to say that, provided the alien frequency content of a steady complex sound was 26 or more decibels below the level of the whole, there was no distortion perceptible by ear. Although this does not go very far as a precise scientific statement, we should like to have an equally simple criterion applicable to transients, if this is at all possible.

Electronics and Electron Tubes

By E. D. McArthur. pp. 173+viii, 89 photographs and diagrams. John Wiley and Sons, New York, and Chapman and Hall, Ltd., 11, Henrietta Street, London, W.C.2. Price, 12s. 6d.

"In the last few years, electron tubes have been finding a rapidly expanding field of application in many kinds of industry. They are 'radio tubes' no longer; instead, they are recognised as extremely versatile devices which are supplying solutions to many difficult problems." In thus introducing his subject, the author ostensibly regards it in no narrow sense; and one may therefore feel just a little disappointed to find that the book tends to work out in the proportions of the General Electric Co. of America. Thus (as the author himself points out) gas-discharge tubes are given exceptional space; but no mention is made of the many important electronic devices associated with the names of Zworykin, Farnsworth, and Hazeltine. In fact, it is stated that "secondary emission phenomena are not used commercially to any great extent but do cause undesirable effects which must be recognised and eliminated." On the other hand, it is only fair to point out that the author does not claim to do more than describe "the fundamental principles which govern the action of all electron tubes," and this he has done in clear, readable, and well-arranged form.

No attempt is made to treat such a vast subject exhaustively, and it is not overloaded with mathematical proofs or with data on particular products; instead, the most useful formulae and working principles are concisely presented in a way that will appeal to the practical reader. The properties of electrons, atoms, molecules, and gases are briefly reviewed; subsequent chapters deal with the elements of electron tubes in general; two-electrode tubes; control of electron currents; applications of triode and multi-grid tubes; gas-filled tubes and their applications; a rather inadequate chapter on special tubes; and finally one describing briefly the construction of electron tubes.

The proportion given to gas-filled tubes is fully justified, for their principles are at once more complex and less generally known than those of high-vacuum valves; and the author is in a favourable position to give an authoritative account of them.

While the book has the advantages of conciseness, further study of any particular point is facilitated by a system of numbered references in the text to bibliographies at the end of each chapter. The European reader may consider it unfortunate that, with only one or two exceptions, these references are American; for some of the matters referred to have been more thoroughly investigated on this side of the Atlantic; but it is consistent with the author's policy of not straying beyond his experience.

It is a volume that can be highly recommended either to students approaching the subject for the first time or to experienced engineers as a "refresher." M. G. S.

Correspondence

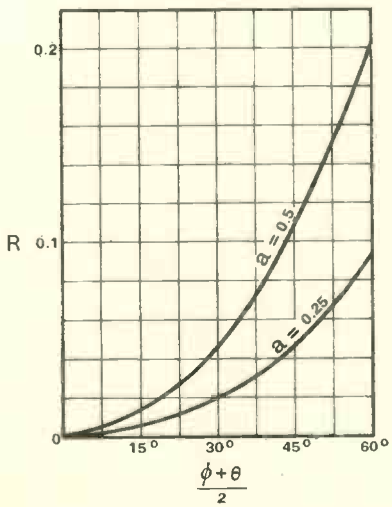
Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain

Side-Band Phase Distortion

To the Editor, The Wireless Engineer

SIR,—We were very interested in Prof. Howe's vectorial demonstration of side-band distortion given in the Editorial of your October issue, and we have to thank Prof. Howe for pointing out that Fig. 2 of our article in the same issue is inaccurate.

The errors, as he implied, are due to not using a sufficient number of terms in calculating the constant and fundamental components, and the formulae themselves appear to be quite accurate. We have found that by expressing formulae (6) and (7) as a series in powers of a^2 the arithmetical work is simplified. The distortion has been carefully recalculated, and our new results are given in the accompanying graph.



This graph is the result of the recalculations referred to in this letter.

We agree that in certain cases (e.g., when the distortion is small) it would be more convenient to calculate the fundamental and the first few harmonics and, as stated in our article, formulae for these harmonics may easily be picked out from our power series of cos ψ.

For high modulation coefficients we have under consideration a method of calculating the constant term in terms of elliptic integrals, and once the constant term is known Prof. Howe's approximation $(A-B)/(A+B)$ for the distortion could conveniently be employed.

Considerable distortion may easily occur in practice, especially when a comparatively low frequency carrier is modulated with high frequencies. For example, the effect has been observed in the I.F. stages of a superheterodyne receiver for television signals, particularly when "staggered" tuning is used for obtaining a wide band pass.

Since the publication of our note we have learned that the distortion for low values of modulation was calculated by M. V. Callender as long ago as 1932, by means of a binomial expansion.

In conclusion, may we point out that according to Prof. Howe's published figures the second harmonic is greater than the fundamental (his figures are 0.362 and 0.336). There is obviously a mistake somewhere.

Baird Television, Ltd.,
Crystal Palace,
Upper Norwood.
London, S.E.19

E. E. WRIGHT.
D. M. JOHNSTONE.

[The harmonics in the second column on p. 518 were obtained from a diagram drawn to ten times the scale of Fig. 2, and the values there given refer to a carrier amplitude of 10 and a fundamental of 3.36, as can readily be seen from the statement that 0.368 is 10.96 per cent. of the fundamental. It was perhaps misleading to assume three different carrier amplitudes, viz., 1, 100 and 10, on the same page, and the harmonics should certainly have been given to the same scale as the fundamental, viz., $A_1 = 0.336$; $A_2 = 0.0362$; $A_3 = 0.0065$; $A_4 = 0.002$. The correct values were, of course, employed in calculating the harmonic content.

G. W. O. H.]

To the Editor, The Wireless Engineer

SIR,—I was very interested to read the paper by Messrs. Johnstone and Wright on side-band phase distortion, and the Editorial on the same subject, in the October issue. As a matter of fact, however, I think that the presence of distortion under these conditions has been pointed out a number of times before. At any rate, in a paper in *Proc. Inst. Rad. Eng.* of September, 1932, I considered the more general case where the side-bands may have different amplitudes or phase differences or both: the results for a parabolic detector were obtained in addition to those for a linear detector. The basic equations agree with those in the present paper (*e.g.*, equation 4*b* is there implicit), but the numerical results are not so concordant: for example, for the case mentioned in the Editorial $\left(\frac{\theta+\phi}{2} = 45^\circ \text{ and } a = 0.5\right)$, my calculated second harmonic was about 9 per cent., agreeing roughly with the graphical method there used: Johnstone's figure for this case is 16 per cent. for total R.M.S. harmonic (not 22 per cent. as stated in the Editorial*). In my case the second harmonic was merely evaluated approximately from a simple expansion, since I was not equal to dealing with zonal harmonics in the facile way exhibited in the present paper.

The investigation of the introduction of phase differences between side-bands by tuned circuits is the natural corollary to Messrs. Johnstone and Wright's work: the simpler aspects of this problem were examined with some numerical examples in my paper, but there is still a considerable field here for investigation.

It is noteworthy that if we limit ourselves to the ordinary radio receiver comprising single circuits and symmetrical two circuit band-passes, this side-band phase distortion will only occur if we are off tune or, of course, if the alignment is incorrect: this will normally be accompanied by inequality of the two side-bands, though it is possible to imagine cases where this will not be so, *e.g.*, where one is just sufficiently off tune on a simple double-hump band-pass to locate one side-band on each side of a hump. The question of whether in complicated cases, such as that of a television receiver, the phase distortion is likely still to be present when the circuits have been adjusted for flat frequency response, is a difficult one: however, the writer would hazard the opinion that no likely combination of circuits will give such distortion provided that they are all accurately aligned to the signal (or symmetrically staggered) and provided that the H.F. Frequency characteristic is exactly symmetrical, *e.g.*, Q values of each circuit of a two-circuit band-pass should be equal, etc.

Pye Radio Ltd., Cambridge. M. V. CALLENDAR.

[* Mr. Callendar's statement that "Johnstone's figure for this case is 16 per cent. for total R.M.S. harmonic (not 22 per cent. as stated in the Editorial)" is apparently based on some misunderstanding. The "nearly 22 per cent." was estimated from the original graph, whilst Fig. 2 on p. 535 gives exactly 20 per cent. We do not know how our correspondent obtains his figure of 16 per cent.—ED.]

"Cathode-Ray Oscillography"

To the Editor, The Wireless Engineer

SIR,—I cannot entirely agree with your Editorial Article in the August issue on "Cathode-Ray Oscillography," by MacGregor Morris and Henley. An examination of a copy of this book, while confirming the rest of your remarks, leads me to differ from your opinion that: "The authors are on safe ground so long as they are giving descriptive accounts of constructional details and experimental apparatus."

The diagrams of connections, in particular, are often misleading, and as inaccurate as the mathematical work which you so justly criticise. It is a fact too little appreciated by most writers, that the drawing of a really good diagram of connections is an art, worthy of even more care than the accompanying letter-press. Some may have a natural gift for it, most of us have to reach the desired end by sweat and blood, drawing and redrawing a circuit diagram in the endeavour to make it self-explanatory without the aid of the description in the text. That this result is attainable, may be seen by examining the diagrams in "The Cathode-Ray Oscillograph in Radio Research" (hereinafter referred to as "R.R.") Compare the lucidity of Fig. 15 of "R.R." with the authors' Fig. 93*b*, which is the same circuit with the omission of any representation of the cathode-ray tube itself. In spite of this omission, which does not help the reader, the resulting diagram looks more complicated. It would be very difficult to improve on any diagram in "R.R.," and it is certainly unwise to omit the connections to the tube entirely, as the authors do in many, if not all, cases, (*c.f.*, the time-bases of Figs. 93*a* and 96 with Figs. 13 and 12 of "R.R.").

Perhaps the worst error is in Fig. 127, purporting to be a "Circuit for hysteresis measurement." This is not a "mere slip," for in the adjacent text we read ". . . along the y — axis deflections proportional to the flux-density are obtained from the voltage developed across R_3 by the secondary current." Since R_3 is shown as a resistance connected directly across the secondary winding, it is manifest that the deflection will be proportional to dB/dt, *not* B, which is very different. The condenser-resistance arrangement necessary to obtain the time-integral of the secondary voltage, which *is* proportional to B, is entirely omitted; this is all the more dangerous in that if a circuit is connected up exactly as in the diagram given, the resulting curve of dB/dt against H does bear a superficial resemblance to a $B-H$ curve! The authors need have looked no further than the issue of this journal for April 1935 for a diagram of a circuit which will give a true $B-H$ curve.

In Fig. 123, "Circuit used for dielectric loss measurements," the deflector-plates marked a and c are shown with no "grid-leaks" or conductive path of any kind for D.C. to the anode. Yet we read on page 142 "The importance of tying down the potential of all plates of the oscillograph cannot be too strongly emphasised. Each plate should have a leakage path to the anode of not less than 5MΩ. Isolated plates drift in potential, causing random movements of the beam which are very puzzling." While agreeing most emphatically with what the authors mean, (but do not say), namely, that the resistance of the leakage path must not be *greater* than 5MΩ, or, if they prefer it, that the leak conductance must not be less than 0.2 micro-mho, one can only wonder, especially in view of the statement that "Many readers of this book will never have used an oscillograph or seen one in use," that the authors have not taken meticulous care to show this leakage path in all diagrams. But in fact we find that in most figures the anode is entirely omitted; where it is shown, as in Fig. 97, no connection whatever to the rest of the circuit is drawn. A comparison of Fig. 98 with Fig. 36 of "R.R." shows that not only is the anode of the cathode-ray tube omitted from the former, but also the vital earth connection to the cathodes of the four triodes.

The frequency comparison circuit of Fig. 125 is inferior in practical convenience to that described in the booklet sent out by Messrs. Cossor with each of their tubes.

The "Device for simultaneously recording two phenomena" (Fig. 137) is of much less general applicability than that described by Davidson in the *Journal of Scientific Instruments*, November 1934 (*W.E.* abstract 857 of 1935).

The "Typical gas-focused oscillograph supply circuit" of Fig. 76 is apparently for "battery H.T." yet we read in the accompanying text that "The supply circuit of a low-voltage oscillograph usually takes the form of a mains unit in a compact box. . . ." Instead of a diagram of connections of such a mains unit, we have only a very beautiful, but quite useless, photograph in Fig. 77, which shows nothing but the "compact box."

Cambridge. C. R. COSENS.

Television Technical Terms and Definitions

By E. J. G. Lewis. pp. 95 + 14 for notes. 13 Figs. Sir Isaac Pitman and Sons, Ltd., Parker Street, London, W.C.2. Price 5s.

It has been said that television embraces all the sciences: certainly, if it be regarded primarily as a branch of radio engineering, technicians in that field are suddenly finding themselves in need of a knowledge of optics, chemistry, and the practice of many arts based on them. Similarly those approaching from the optical standpoint may find themselves out of their depths in electrical fundamentals.

The form of the book under review is perhaps the most helpful for a rapid supplementing of one's technical knowledge in the less familiar corners of this multifarious art. Over 1,000 definitions are given by the author; and, recognising that there are many more to come, he has provided 14 blank pages under alphabetical heads for recording subsequent information.

The book is claimed to be "complete, up to date, and authoritative." Unfortunately, a perusal of it does not fully confirm this rather ambitious description.

Notable omissions are *Interference* (both electrical and optical varieties) and *Dipole*. It is true that *Doublet Aerial* is defined, but only as a *horizontal* aerial, whereas for effective reception of television signals in this country a doublet should be vertical.

Any up-to-date work on television should do better justice to the *Scophony System* (which, incidentally, is on p. 21 attributed to *Watson*). Of the important developments of the last two years, including the famous supersonic light relay, no mention is made.

Finally, as regards authoritativeness, it is particularly to be regretted that further publication is given to the popular confusion of what is commonly (but not altogether correctly) called the *Kerr Effect* with the *Faraday Effect*. The former, of more practical importance in television, is described as the rotation of a beam of polarised light by an electro-magnetic force. Actually, of course, it does not result in rotation of a beam of light, or even of the plane of polarisation thereof; and is not effected by electro-magnetic force.

The definition of *Johnson Effect* as "variations in the resistance of wire and components . . ." seems hardly adequate, for even although such variations are stated to be "extremely slight" there is no other suggestion that the term being defined might not cover any sort of variableness of resistance. Another example of looseness of description occurs under *Oscillator, Squegging*, where it is stated that "a grid leak and condenser discharges the accumulated charge on the grid"; while we wonder what idea the reader searching for information on the super-regenerative receiver (defined not under that title but under *Armstrong Circuit*) would gain from such a statement as "The circuit is kept just below oscillation point, and therefore in its most sensitive condition."

Such criticisms extend to a comparatively small proportion of the definitions, however; and if the reliability of the information were everywhere put beyond question by a little more precision here and there praise of the work would be unqualified. The illustrations cover the more important devices used in practical television, and explain their basic principles with admirable clearness. As a suggestion for the next edition, references to them might usefully be made under all headings that they illustrate. M. G. S.

An International Radio Vocabulary

THE International Electro-Technical Commission has issued an advance announcement regarding the forthcoming publication of the first edition of the International Electro-Technical Vocabulary, which is due to appear next year. This event marks the culmination of many years' work by delegates from all parts of the world.

Definitions are to be given in both English and French (the official languages of the I.E.C.) while a translation of the *terms* only will appear in German, Italian, Spanish and Esperanto.

Some at least of the language difficulties encountered in drafting internationally acceptable definitions in both English and French have been surmounted by the action of the British and American Committees, who agreed that the definitions should in the first place be drafted in French.

This first edition of the Vocabulary represents an attempt towards the unification of electrical nomenclature, and although only one of its 14 sections (that headed Radio-communication) deals exclusively with our subject, it will probably be accorded at least as warm a welcome in wireless circles as in other branches of electrical engineering.

The price of the Vocabulary has been tentatively fixed at approximately 10/- the exact price being dependent on the anticipated demand. Those interested are invited to request the British National Committee of the I.E.C. to reserve a copy or copies on publication; the address is c/o the British Standards Institution, 28, Victoria Street, London, S.W.1.

Valve Data

THE Valve Data number of *The Wireless World* was published on November 20th and contained a supplement in which is given a list of over 850 valves now on the market, including American types, with tabulated details of their properties. Special articles on new valve developments are also a feature of this issue.

Generalised Characteristics of Linear Networks*

By E. K. Sandeman

LINEAR electrical networks consisting of inductances, capacitances and resistances are employed among other purposes to supply characteristics which vary with the frequency of the applied electrical oscillations.

When the characteristics of interest are impedance, attenuation and phase shift, it is evident that by changing the absolute values of the elements of inductance capacitance and resistance, a given form of characteristic, expressed as a function of frequency, may within practical limits, be obtained in any required range of absolute frequency. Further, a network, designed for instance, to afford a variation of attenuation with frequency, may be arranged to give the same attenuation curve as a function of frequency, when operating between any values of impedance lying within a practical range of impedances determined by the limitations of design of the component elements and of the associated circuits. Evidently the performance of any specific embodiment of such a network is specifiable uniquely by a specific curve of attenuation against frequency, this curve being valid only for given absolute values of terminating impedance.

An infinite number of infinite series of curves are therefore required to specify the performance of the whole group of embodiments of a single form of attenuating network.

As will be seen by reference to any work on filter theory, it is well known that when the relations between the elements are fixed, by performing a simple process of disembodiment, the performance of the whole group of networks of a given form can be determined by a single curve for each characteristic required.

Although the present purpose is mainly to present a series of such characteristics, the use of these characteristics will be made clearer by a preliminary discussion of their method of derivation for the benefit of those to whom the method is not familiar.

* MS. accepted by the Editor, July, 1936.

Two-Terminal Networks

Such networks are usually employed to afford an impedance in which the resistive and reactive components vary with frequency in some prescribed way.

Fig. 1 shows six two-element networks for which generalised characteristics are given. Consider the network Type 1 of Fig. 1.

The impedance of this at any angular frequency ω is given by

$$Z = \frac{RLj\omega}{R + Lj\omega} = \frac{RLj\omega(R - Lj\omega)}{R^2 + L^2\omega^2}$$

$$\therefore Z = \frac{RL^2\omega^2}{R^2 + L^2\omega^2} + j\frac{R^2L\omega}{R^2 + L^2\omega^2} \quad \ldots \quad (1)$$

Define f_0 as the frequency at which the reactance of L in ohms equals the numerical value of R in ohms and make $\omega_0 = 2\pi f_0$, then $L\omega_0 = R$ and $L = \dfrac{R}{\omega_0}$ .. .. (2)

Substituting (2) in (1)

$$\therefore Z = \frac{R \cdot R^2\dfrac{\omega^2}{{\omega_0}^2}}{R^2 + R^2\dfrac{\omega^2}{{\omega_0}^2}} + j\frac{R^2R\dfrac{\omega}{\omega_0}}{R^2 + R^2\dfrac{\omega^2}{{\omega_0}^2}}$$

$$= R\left[\frac{\dfrac{\omega^2}{{\omega_0}^2}}{1 + \dfrac{\omega^2}{{\omega_0}^2}} + j\frac{\dfrac{\omega}{\omega_0}}{1 + \dfrac{\omega^2}{{\omega_0}^2}}\right] \quad \ldots \quad (3)$$

$$= R[Y + jX] \text{ say.}$$

Evidently the quantity within the brackets is characteristic of the impedance of all possible networks of this type. If the real and the imaginary parts of the quantity inside the brackets are plotted as a function of $\dfrac{\omega}{\omega_0}$ the resultant curves afford a rapid means of determining the impedance characteristics of any specific embodiment of the network, by assigning values to R and L and so to

$$\omega_0 = \frac{R}{L} \quad f_0 = \frac{R}{2\pi L}$$

Fig. 2 shows a plot of the functions Y and X dependent on $\frac{\omega}{\omega_0}$.

$$Y = \frac{\frac{\omega^2}{\omega_0^2}}{1 + \frac{\omega^2}{\omega_0^2}} \text{ and } X = \frac{\frac{\omega}{\omega_0}}{1 + \frac{\omega^2}{\omega_0^2}}$$

It may be noted that $\frac{\omega}{\omega_0} = \frac{2\pi f}{2\pi f_0} = \frac{f}{f_0}$

Example

Determine from Fig. 2 the impedance characteristic of a network of Type 1 in which $L = 1$ Henry and $R = 1{,}000$ ohms.

$$f_0 = \frac{R}{2\pi L} = \frac{1{,}000}{2\pi} = 159$$

Hence when $f = 159 \qquad \frac{f}{f_0} = \frac{\omega}{\omega_0} = 1$

when $f = 318 \qquad \frac{f}{f_0} = \frac{\omega}{\omega_0} = 2$

when $f = 79.5 \qquad \frac{f}{f_0} = \frac{\omega}{\omega_0} = 0.5$

and so on.

The absolute frequency scale can therefore be determined by writing 159 against $\frac{\omega}{\omega_0} = 1$ on Fig. 2, 318 against $\frac{\omega}{\omega_0} = 2$, 79.5 against $\frac{\omega}{\omega_0} = 0.5$, and so on.

In practice, of course, this cumbersome method is never adopted. Plotting is carried out to a logarithmic frequency scale as in Fig. 2, as a result of which it is only necessary to have a scale of absolute frequency plotted on the same kind of paper representing equal ratios of $\frac{\omega}{\omega_0}$ and of frequency by equal intervals, in which case, by sliding the two scales so that the frequency 159 comes opposite $\frac{\omega}{\omega_0} = 1$ the curves of Fig. 2 have direct application to the scale of absolute frequency.

The values of reactance and resistance are then given by 1,000 Y and 1,000 X. The values of reactance and resistance are given for three values of f in the table below.

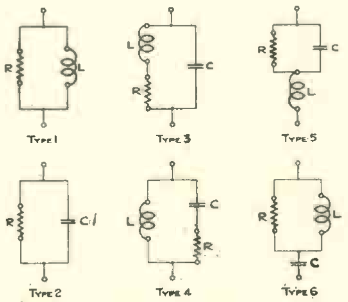


Fig. 1.—*(Above) Classification of six most simple networks containing a single resistance.*

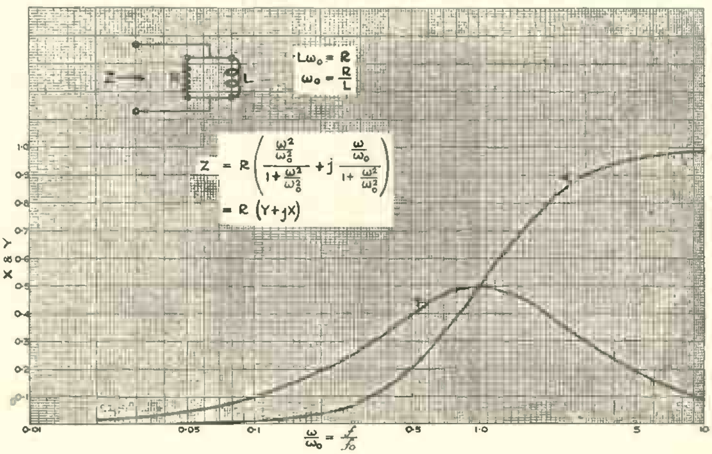


Fig. 2.—*(Right) Generalised resistance and reactance curves for network shown.*

f	Y	Network Resistance. Ohms.	X	Network Reactance. Ohms.
79.5	0.2	200	0.4	400
159	0.5	500	0.5	500
318	0.8	800	0.4	400

In the case of networks containing more than two elements, it is necessary to introduce extra parameters determining the relations between the elements, the number of parameters being always one less than the number of elements in the network.

Fig. 1 Type 3 shows a very common form of three-element two-terminal network.

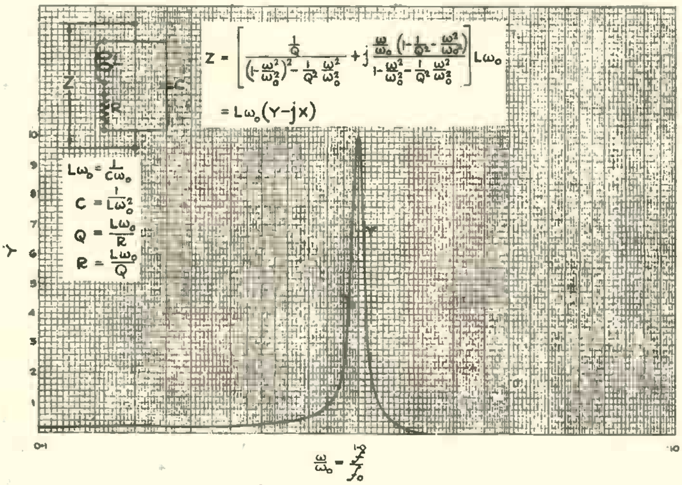


Fig. 3.—(*Above*) *Generalised resistance curve for network shown with* $Q = 10$.

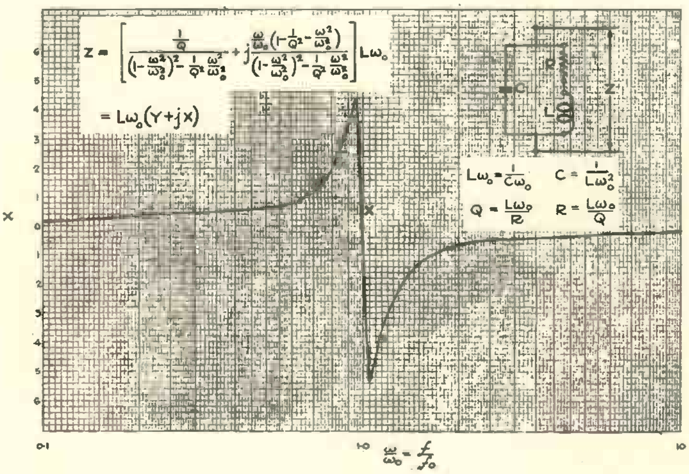


Fig. 4.—(*Right*) *Generalised reactance curve for network shown with* $Q = 10$.

pression may be carried out in a number of ways, by insertion of the parameters ω_0 and Q. It is evidently possible to eliminate any two of the element values, and to obtain resulting expressions which when multiplied by the value of the remaining element give the resultant resistance and reactance components of the impedance. Since, however, the purpose of a network of this kind is usually to obtain a high impedance at the resonant frequency, it is convenient to think of the condenser as a means of multiplying the reactance of the inductance by a calculable factor. For this reason the impedance is determined as a function of $L\omega_0$ the reactance of L at resonance.

Put $$L\omega_0 = \frac{1}{C\omega_0} \quad \ldots \quad (4)$$

and $$\frac{L\omega_0}{R} = Q \quad \ldots \quad \ldots \quad (5)$$

$$Z = \frac{1}{Cj\omega + \dfrac{1}{R + Lj\omega}}$$

$$= \frac{R + Lj\omega}{1 + Cj\omega(R + Lj\omega)}$$

$$= \frac{R + Lj\omega}{1 - LC\omega^2 + RCj\omega}$$

$$= \frac{(R + Lj\omega)(1 - LC\omega^2 - RCj\omega)}{(1 - LC\omega^2)^2 + R^2C^2\omega^2}$$

$$= \frac{R}{(1 - LC\omega^2)^2 + R^2C^2\omega^2} + j\frac{L\omega(1 - LC\omega^2) - R^2C\omega}{(1 - LC\omega^2)^2 + R^2C^2\omega^2} \quad \ldots \quad (6)$$

The process of generalisation of this ex-

From (4) and (5) therefore are obtained

$$C = \frac{1}{L\omega_0^{\,2}} \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

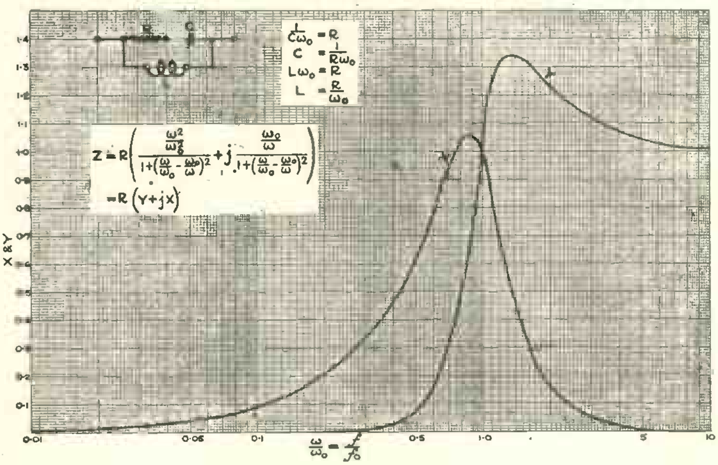
$$R = \frac{L\omega_0}{Q} \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

Substituting (7) and (8) in (6)

$$\therefore Z = \frac{\dfrac{L\omega_0}{Q}}{\left(1 - \dfrac{\omega^2}{\omega_0^2}\right)^2 + \dfrac{1}{Q^2}\dfrac{\omega^2}{\omega_0^2}} + j\frac{L\omega_0 \dfrac{\omega}{\omega_0}\left(1 - \dfrac{\omega^2}{\omega_0^2} - \dfrac{1}{Q^2}\right)}{\left(1 - \dfrac{\omega^2}{\omega_0^2}\right)^2 + \dfrac{1}{Q^2}\dfrac{\omega^2}{\omega_0^2}}$$

$$= L\omega_0\left[\frac{\dfrac{1}{Q}}{\left(1 - \dfrac{\omega^2}{\omega_0^2}\right)^2 + \dfrac{1}{Q^2}\dfrac{\omega^2}{\omega_0^2}} + j\frac{\dfrac{\omega}{\omega_0}\left(1 - \dfrac{1}{Q^2} - \dfrac{\omega^2}{\omega_0^2}\right)}{\left(1 - \dfrac{\omega^2}{\omega_0^2}\right)^2 + \dfrac{1}{Q^2}\dfrac{\omega^2}{\omega_0^2}}\right] \quad \ldots \quad (9)$$

It may be noted in passing that when $\dfrac{\omega}{\omega_0} = 1$ the real part of the impedance $= QL\omega_0$ and the imaginary part $= -L\omega_0$.



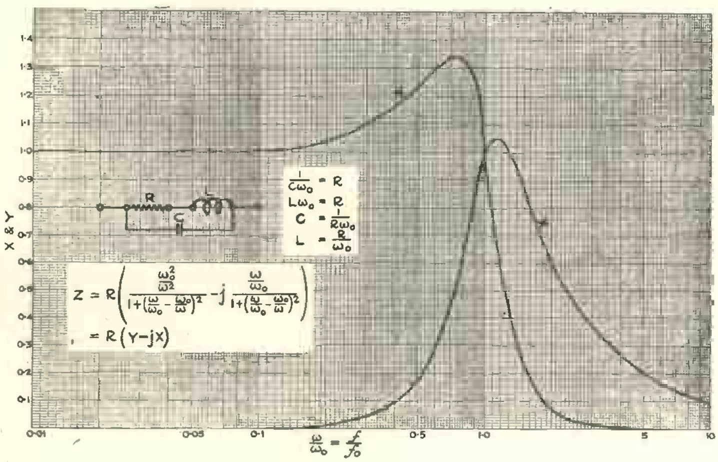


Fig. 5.—*(Left) and* Fig. 6 *(Above) Generalised resistance and reactance curves for networks shown.*

Evidently equation 9 represents a family of curves, the particular curve chosen being determined by the value of Q.

Fig. 3 shows the curve for the resistance term inside the bracket and Fig. 4 the curve for the reactance term when $Q = 10$.

To determine the absolute resistance and reactance curves it is only necessary to fix L and C and so determine ω_0, or to fix L and ω_0 and so determine C.

Fig. 5 shows generalised curves for a type 3 network derived in terms of R instead of $L\omega_0$ for the case where $Q = 1$, i.e., $L\omega_0 = \dfrac{1}{C\omega_0} = R$.

Fig. 6 shows curves for a type 4 network (which is similar to type 3 with the resistance transferred from the inductance arm to the capacitance arm) for the case where $L\omega_0 = \dfrac{1}{C\omega_0} = R$.

Comparing Figs. 5 and 6, it will be seen that one can be converted into the other by inverting one set of curves about the point $\dfrac{\omega}{\omega_0} = 1$. This is equivalent to substituting $\dfrac{\omega_0}{\omega}$ for $\dfrac{\omega}{\omega_0}$ on the frequency ratio scale. One other point has to be observed: in Fig. 5 the reactance is always negative, and in Fig. 6 the reactance is always positive.

The generalised characteristics of networks (5) and (6) are shown respectively on Figs. 7 and 8, and, as might be expected, are mutual inverts about the point $\dfrac{\omega}{\omega_0} = 1$.

There exists still another useful method of approach to the type 3 network, and that is applicable when the capacitance represents an unwanted shunt capacitance, such as the grid cathode capacitance of a valve, the resistance is an existing circuit element in parallel with this capacitance, such as the anode resistance of the previous valve, and the inductance is added to effect partial neutralisation of the capacitance. The net effect of adding the inductance is to increase the impedance effective across the condenser terminals over a limited range of frequency.

Assume such a value of neutralising inductance L is chosen that $R = nL\omega_0 = \dfrac{n}{C\omega_0}$

where $\omega_0 = \dfrac{1}{\sqrt{LC}}$.

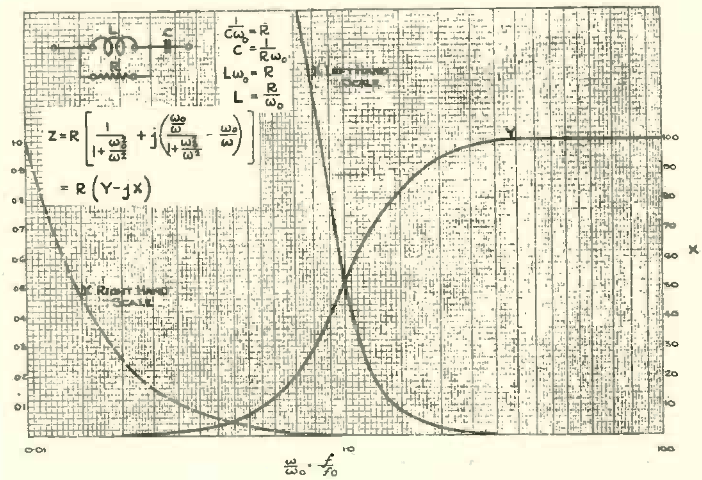
From equation (6) the reactive component of the resulting impedance

$$X = \frac{L\omega(1 - LC\omega^2) - R^2C\omega}{(1 - LC\omega^2)^2 + R^2C^2\omega^2}$$

Inserting $L = \dfrac{R}{n\omega_0}$ and $C = \dfrac{n}{R\omega_0}$

$$X = \frac{\dfrac{R}{n}\dfrac{\omega}{\omega_0}\left(1 - \dfrac{R}{n}\dfrac{n}{R}\dfrac{\omega^2}{\omega_0{}^2}\right) - R^2\dfrac{n}{R}\dfrac{\omega}{\omega_0}}{\left(1 - \dfrac{\omega^2}{\omega_0{}^2}\right)^2 + R^2\dfrac{n^2}{R^2\omega_0{}^2}\omega^2}$$

$$= R\left[\frac{\dfrac{\omega_0}{\omega}\left(\dfrac{1}{n} - n\right) - \dfrac{1}{n}\dfrac{\omega}{\omega_0}}{\left(\dfrac{\omega_0}{\omega} - \dfrac{\omega}{\omega_0}\right)^2 + n^2}\right] \quad .. \quad (10)$$



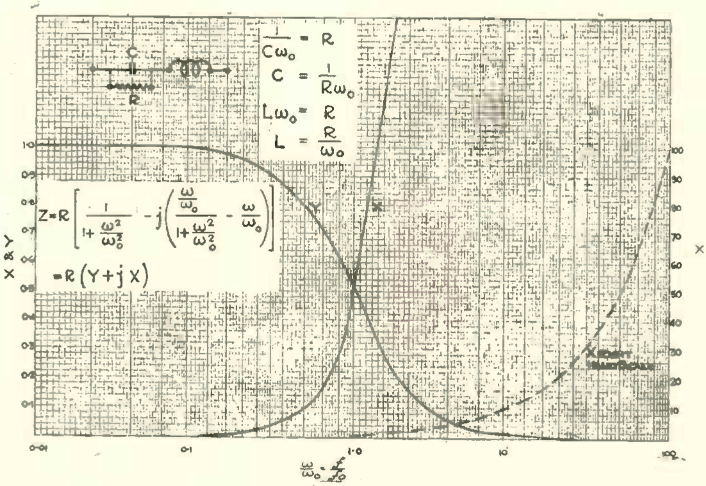


Fig. 7.—(*Left*) and Fig. 8 (*Above*) *Generalised resistance and reactance curves for networks shown.*

Evidently if the quantity $\dfrac{1}{n} - n$ is finite the reactance assumes finite positive values at low frequencies. For certain purposes this may be regarded as undesirable and it is customary to make $n = 1$ in which case

$$X = R\left[\frac{-\dfrac{\omega}{\omega_0}}{1 + \left(\dfrac{\omega}{\omega_0} - \dfrac{\omega_0}{\omega}\right)^2}\right] \quad .. \quad .. \quad (11)$$

$$\text{and } L = \frac{R}{\omega_0} = R\sqrt{LC} = R^2C \quad .. \quad (12)$$

The resistance and reactance characteristics of the resultant network are then given by Fig. 5.

Fig. 9 shows the resistance and reactance characteristics of a type 2 network which represents the condition before the addition of the neutralising inductance.

Some idea of the improvement realised by the addition of the neutralising inductance can be obtained by comparison of Figs. 5 and 9. For the purpose of comparison these figures can be directly superposed with the $\frac{\omega}{\omega_0} = 1$ points and the zeros of the "X and Y" scale coinciding. It will then be seen that the value of $Y + jX$ is appreciably larger over an interval of nearly two octaves, when the inductance is added; the fall in impedance is delayed to a much higher frequency.

Four-Terminal Networks

The characteristics here considered are the usual transmission characteristics which determine the effect of the network in modifying electrical oscillations traversing it.

Three characteristics will be defined.

(1) *The insertion loss proper* determines the change in the received energy and voltage consequent on the insertion of the network between two pieces of apparatus, a generator of alternating electromotive forces in a given range of frequency, and a receiver. The generator and the receiver each have a definite impedance; for many purposes this can be usefully considered to be a pure resistance, and often the impedance of generator and receiver are equal. When the generator is connected directly to the receiver, it delivers to the receiver energy and voltage of determined magnitude and phase. When the network is inserted the received energy and voltage is changed in magnitude and phase.

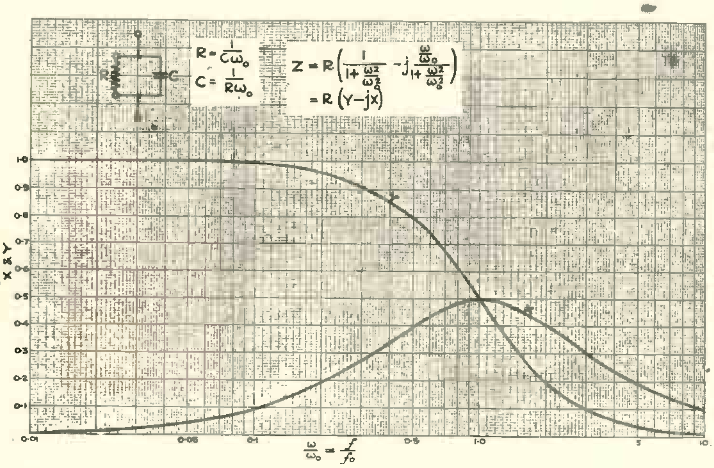


Fig. 9.—*Generalised resistance and reactance curves for network shown.*

The amount of this change determines the insertion loss proper.

Evidently the insertion loss proper has two characteristics—a power ratio and a phase angle at each frequency. For simplicity the term *insertion loss* will be used to refer to the power ratio and the term *insertion phase shift*, or simply phase shift, to refer to the angle.

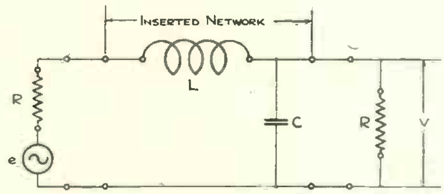


Fig. 10.

Evidently the power ratio may be expressed as the numerical ratio of P_1, the received power, when the network is out of circuit and P_2 the received power when the network is in circuit, more usually it is expressed in decibels or nepers. In the present case decibels are used, in which case the insertion loss is given by

$$L = 10 \log_{10} \frac{P_1}{P_2} \text{ decibels.}$$

If the generator impedance = the receiver impedance = a pure resistance, and if V_1 and V_2 are the received voltages corresponding respectively to P_1 and P_2, then

$$L = 20 \log_{10} \left| \frac{V_1}{V_2} \right| \text{ decibels.}$$

The vertical bars each side of the fraction $\frac{V_1}{V_2}$ indicate that the magnitudes of the voltage vectors are to be taken.

(2) *The voltage transfer constant* is the vector ratio between the vectors respectively describing the input voltage applied to the network and the voltage observed across the output terminal of the network: defined by $\frac{\text{output voltage}}{\text{input voltage}}$.

It is most conveniently determined by

two curves plotted against frequency, one of the magnitude of the voltage transfer constant, and the other of the angle of the voltage transfer constant.

(3) *The voltage transfer loss* is probably a new term and is expressed as the number of decibels which correspond to A, the magnitude of the voltage transfer constant.

The voltage transfer loss

$$L_T = 20 \log_{10} A$$

The use of this term appears to be logical and is justified by its convenience. In this case the phase shift may be defined as the transfer phase shift, but no ambiguity results if the term phase shift is used alone.

It is difficult to lay down rigid rules which determine when each type of characteristic should be used to describe the performance of a network. When the network operates between finite impedances there is no choice, the insertion loss proper must be used, described by the insertion loss and the insertion phase shift as defined above. When the network can be regarded as operating between a generator of zero impedance and open circuit (infinite receiving impedance) the convention adopted is largely a matter of personal preference. Since the received power is zero it is rather difficult to modify suitably the definition of the insertion loss, to describe the network characteristics under this condition, and the use of the voltage transfer constant, or the voltage transfer loss is preferable. The voltage transfer constant is evidently used if voltage ratios are easier to handle and the voltage transfer loss if losses in decibels are required. This case arises in considering coupling circuits between valves, when the anode impedance of the first valve and the grid filament capacitance of the second valve are treated as part of the network and the grid filament conductance of the second valve is low enough to be considered zero.

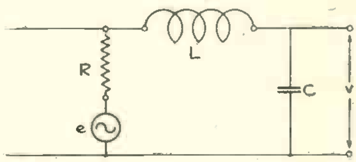


Fig. 12.

Example of Network Treated by Use of Insertion Loss

Fig. 10 shows the equivalent circuit for the high frequency resonance of a communication transformer working between image impedances. The method of deriving this circuit is described in *Electrical Communication*, April 1929, page 282, "Transformers as Band Pass Filters," by E. K. Sandeman.

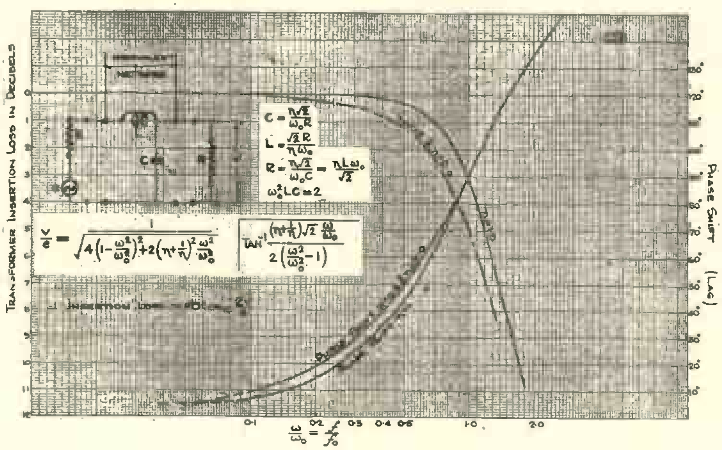


Fig. 11.—*High frequency insertion loss and phase shift of audio-frequency transformers.*

$$\text{The voltage } V = \frac{\dfrac{R}{1 + j\omega CR}}{R + j\omega L + \dfrac{R}{1 + j\omega CR}}\, e$$

$$= \frac{e}{\sqrt{(2 - \omega^2 CL)^2 + \left(\omega CR + \dfrac{\omega L}{R}\right)^2}} \left| \operatorname{Tan}^{-1} \frac{\omega CR^2 + \omega L}{\omega^2 RLC - 2R} \right. \quad .. \quad (13)$$

$$\text{Put} \quad \omega_0{}^2 = \frac{2}{LC}$$

$$\text{and } \frac{\omega_0 CR}{\sqrt{2}} = \frac{\sqrt{2}R}{\omega_0 L} = n$$

Then $C = \dfrac{n\sqrt{2}}{\omega_0 R}$ and $L = \dfrac{R\sqrt{2}}{n\omega_0}$ .. (14)

Substituting 14 in 13 :—

$$\frac{V}{e} = \frac{1}{\sqrt{4\left(1 - \dfrac{\omega^2}{\omega_0^{\,2}}\right)^2 + 2\left(n + \dfrac{1}{n}\right)^2 \dfrac{\omega^2}{\omega_0^{\,2}}}} \Bigg| \mathrm{Tan}^{-1} \frac{\left(n + \dfrac{1}{n}\right)\sqrt{2}\,\dfrac{\omega}{\omega_0}}{2\left(1 - \dfrac{\omega^2}{\omega_0^{\,2}}\right)} \quad .. \quad (15)$$

$$= A\,|\overline{\phi}$$

Fig. 11 for $n = 0.5$, 1.0 and 2. It is evident an ideal design of transformer results if n is made equal to unity, since the loss at high frequencies is then a minimum.

Example of Network Treated by Use of Voltage Transfer Loss

Fig. 12 shows the equivalent circuit for the high frequency resonance of a communication transformer working into open circuit. The method of deriving this circuit will be clear by reference to the above mentioned article.

The voltage transfer constant

$$\frac{V}{e} = \frac{1}{\sqrt{1 + (n^2 - 2)\dfrac{\omega^2}{\omega_0^{\,2}} + \dfrac{\omega^4}{\omega_0^{\,4}}}} \Bigg| \mathrm{Tan}^{-1} \frac{n\dfrac{\omega}{\omega_0}}{1 - \dfrac{\omega^2}{\omega_0^{\,2}}}$$

$$= A\,|\overline{\phi}\,.$$

Where $n = RC\omega_0 = \dfrac{R}{L\omega_0}$

$$\omega_0 = \frac{1}{\sqrt{LC}}$$

$$C = \frac{n}{\omega_0 R} \quad L = \frac{R}{n\omega_0}$$

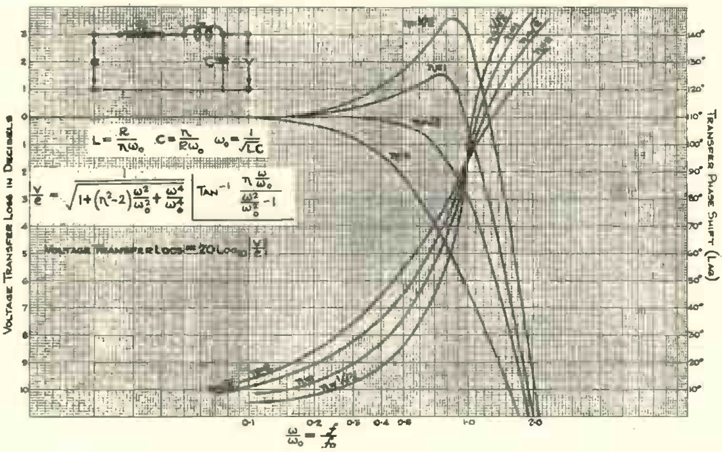
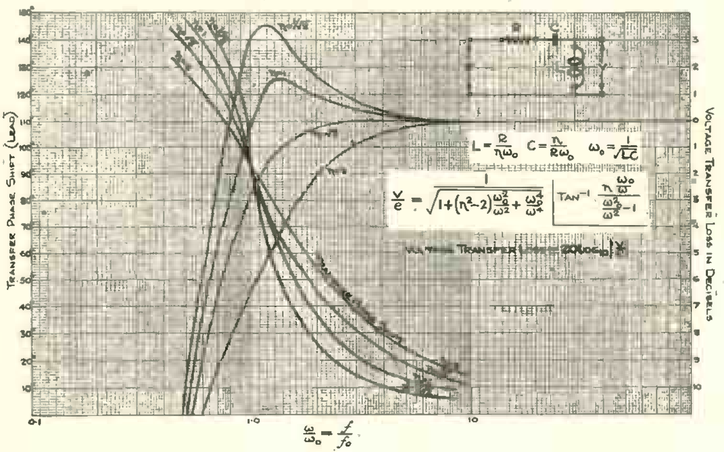


Fig. 13.—*(Above) High frequency response and phase shift for audio frequency transformers working into open circuit.*

Fig. 14.—*(Right) Transfer loss and phase shift of network shown.*



The negative value of the angle makes the sense indeterminate from the analysis, but it is evident that the angle always represents a lag.

Since in the absence of the network $V = \frac{1}{2}e$, the insertion loss proper is expressed by the ratio $2A\,|\overline{\phi}$ and the insertion loss is expressed in decibels by

$$L = 20 \log_{10} 2A$$

The insertion loss and the insertion phase shift of the network of Fig. 10 are shown in

The voltage transfer loss is given by

$$L_T = 20 \log_{10} \left|\frac{V}{e}\right| = 20 \log_{10} A$$

The voltage transfer loss and the phase shift of the network of Fig. 12 are shown on

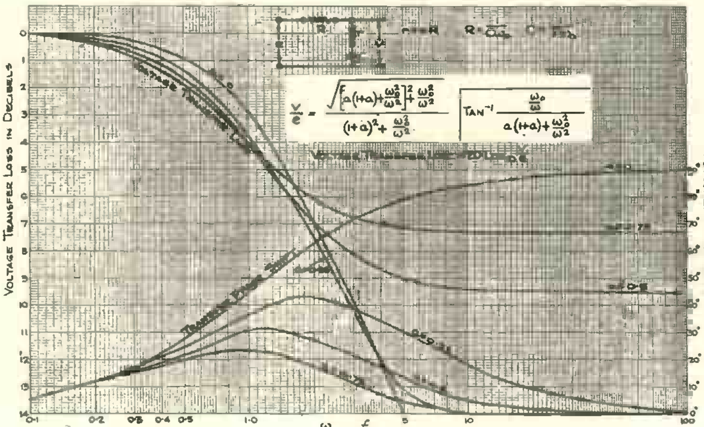
Fig. 13 for values of $n = \frac{1}{\sqrt{2}}$, 1, $\sqrt{2}$ and 2.

Figs. 14–18 show the characteristics of a number of networks in forms appropriate to the work to which they were originally applicable. They also show the method of deriving the parameters and the generalised formulae for the insertion loss, voltage transfer constant, voltage transfer loss, whichever is relevant, and for the phase shift.

For purposes of practical application each figure is complete in itself, containing all the necessary data to enable the absolute frequency scale for any particular embodiment to be aligned with the generalised scale of $\frac{\omega}{\omega_0} = \frac{f}{f_0}$. There are however a few points which appear worth comment.

An important point to notice is the principle of inversion by which the characteristics of one network can be derived from another by inversion about the axis $\frac{\omega}{\omega_0} = 1$. The formula for the second network is derived by substituting $\frac{\omega_0}{\omega}$ for $\frac{\omega}{\omega_0}$ in the formula for the first network, and by changing the sign of the reactance or phase shift.

The necessary and sufficient conditions which make such an inversion possible are: firstly, that the second network is derived from the first by changing every inductance into a capacity and vice versa: secondly, that the controlling parameters relating ω_0 and the various elements of the network shall be the same. If for instance a controlling parameter in the first network is given by $L\omega_0 = aR$, the corresponding parameter in the second network is given by $a'R = \frac{1}{C\omega_0}$ where



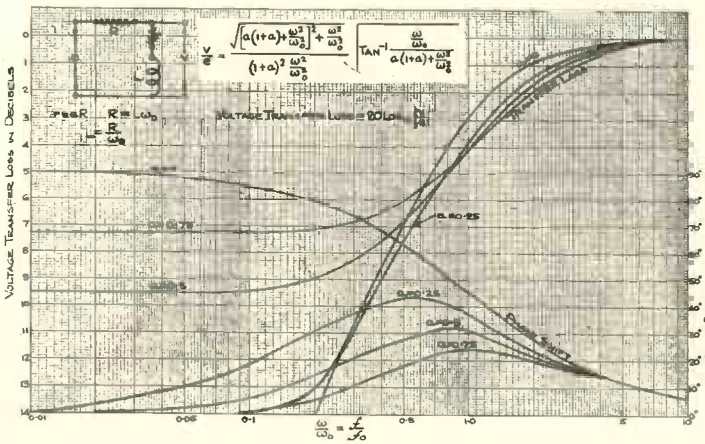
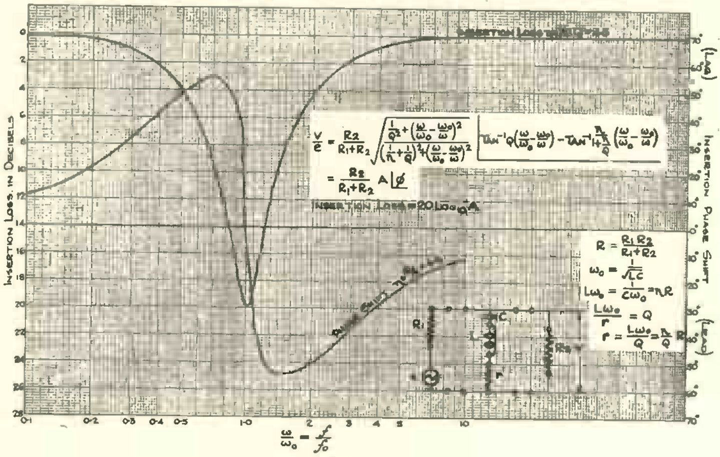


Fig. 15.—(*Above*) *and* Fig. 16 (*Right*) *Transfer loss and phase shift for networks shown.*

C replaces L, and a' must equal a for the condition of inversion to hold. As the labour of deriving the formulae and evaluating them is very considerable in the more complicated cases, the value of this principle will be easily appreciated. Examples of such inversion are Figs. 5 and 6, Figs. 2 and 9, Figs. 13 and 14 Figs. 15 and 16. In each case calculation was made for one case only and the formula and characteristics for the inverted network were derived by applying the principle of inversion. Care must be observed in applying this principle to see that appropriate changes are made in the signs of the reactance and phase shift.

It will be noted that in certain cases to save space, positive and negative reactances have been plotted in the same direction and distinguished by + and − signs. For the same reason on Fig. 18, the part of the response curve representing a gain has been inverted about the axis of zero loss.

In certain cases, for instance, where valves can be used as separating elements, it is possible to combine the curves of several networks by adding or subtracting ordinates directly. In this case the preliminary investigation can be made entirely in terms of the generalised curves, sliding them over one another until a composite curve of the required form is obtained. The values of the circuit elements of the component networks follow as soon as the position of the composite curve (and therefore of the component curves) in the frequency gamut is fixed. It is found in practice that the possession of a series of generalised curves such as the above enables studies, which would otherwise be quite impracticable, to be made comparatively without effort.



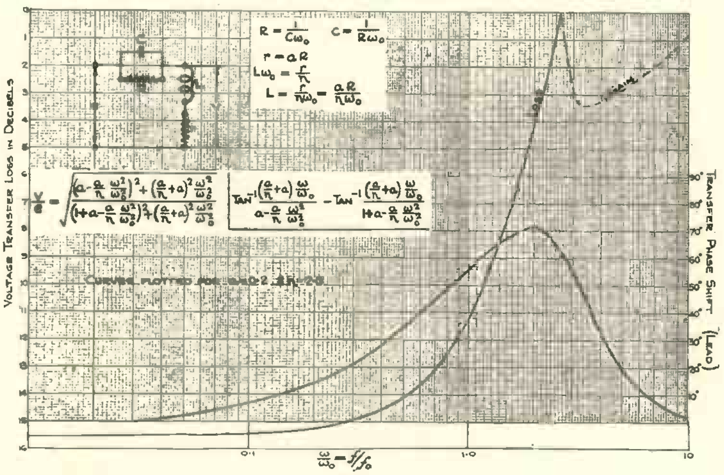


Fig. 17.—(*Left*) *Transfer loss and phase shift for network shown.*

Fig. 18.—(*Above*) *High-frequency insertion loss and phase shift of audio-frequency transformers.*

The collection of curves given is necessarily limited as the time involved in constructing them is appreciable.

Kelly's Engineering Trades Directory 1936

THE 22nd edition of this famous directory of the engineering and allied trades has made its appearance. It contains, among other things, a classified list of the many trades associated with the engineering industry, the names of the various firms being arranged alphabetically under their respective trades which are also in alphabetical order. The London Postal District has a section to itself.

In addition there is another section in which can be found the names and addresses of firms listed under their respective towns and villages, these latter being classified under their counties. The directory may be obtained from Kelly's Directories, Ltd., 186, Strand, London, W.C.2, price 50/-.

Some Recent Patents

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. A selection of abstracts from patents issued in the U.S.A. is also included, and these bear a seven-figure serial number.

AERIALS AND AERIAL SYSTEMS

450 714.—Wireless aerial for long and short wave working.

K. T. Hardman. Application date 23rd August, 1935.

2 029 015.—Circular or polygonal arrangement of aerials for radiating uniformly in the horizontal direction, but with a restricted vertical field.

O. Böhm (assignor to the Telefunken Co.).

TRANSMISSION CIRCUITS AND APPARATUS

448 043.—Preventing distortion in a frequency-modulating system.

Marconi's W.T. Co. (assignees of N. E. Lindenblad). Convention date (U.S.A.) 29th November, 1933.

451 568.—Short-wave oscillation-generator provided with a simplified metering system.

Marconi's W.T. Co. (assignees of N. E. Lindenblad). Convention date (U.S.A.) 16th February, 1935.

452 960.—Split-anode magnetron valve in which both anode voltage and the emission current are separately modulated in phase opposition.

Telefunken Co. Convention date (Germany) 30th November, 1934.

2 028 857.—System of relaying television or other ultra-short-wave signals from a "Beam" transmitter to a group of distant non-directional redistributing aerials.

V. K. Zworykin (assignor to Radio Corporation of America).

2 037 977.—Short-wave oscillation-generator of the split-anode magnetron type, arranged so as to be self-starting.

C. W. Hansell (assignor to Radio Corporation of America).

RECEPTION CIRCUITS AND APPARATUS

446 749.—Tuning dial arranged to have an open scale showing a large number of stations in a comparatively limited space.

B. Hesketh. Application date 1st November, 1934.

446 778.—Biasing scheme for a valve operating as a detector for short-wave signals of the order of 10 metres.

J. R. H. Forman and Baird Television. Application date 30th January, 1935.

446 809.—Inductive coupling provided with means to offset the usual double-hump in the resonance curve.

E. K. Cole. Convention date (Italy) 11th April, 1935.

447 104.—Permeability tuning for a superhet receiver.

Aladdin Radio Patents and H. C. Hebard. Application date 21st November, 1935.

447 125.—Tuning indicator for a wireless set in which the moving part is made to resemble the thread of mercury in a thermometer.

A. C. Cossor and D. G. Rennett. Application date 10th November, 1934.

447 148.—Suppressing the so-called "ground noise" in radio receivers.

E. K. Cole (communicated by AGA-Baltic Radio). Application date 15th January, 1935.

447 593.—Cutting-out static and similar impulsive interference in broadcast reception.

K. Pulvari-Pulvermacher. Application date 4th October, 1935.

448 045.—Screening device for a radio receiver designed for use on a motor car.

J. Y. Johnson. Application date 3rd December, 1934.

448 058.—Variable-selectivity circuit in which control is effected by varying a back-coupling element in accordance with the amplitude of the received signal.

Ideal Werke Akt für drahtlose Telephonie. Convention date (Germany) 15th February, 1934.

448 113.—Amplifier designed to handle a wide band of frequencies without discrimination between high and low.

Kolster-Brandes and C. W. Earp. Application date 30th November, 1934.

448 321.—Method of suppressing "image frequencies" in a superhet receiver.

Standard Telephones & Cables (assignees of Standard Villamossági Részvénytársaság). Convention date (Hungary) 4th April, 1935.

448 779.—Remote tuning control of a wireless receiver through a concentric feed-line or cable.

Marconi's W.T. Co.; G. M. Wright; and N. M. Rust. Application date 14th November, 1934.

449 110.—Multiple-section tuning-coil for all-wave receivers, comprising a series of windings graded both as regards inductance and sectional area.

Hazeltine Corporation (assignees of J. K. Johnson). Convention date (U.S.A.) 5th May, 1934.

VALVES AND THERMIONICS

446 956.—Construction of transmitter-valve designed to facilitate cooling.

Soc. Francaise Radio-Electrique. Convention date (France) 16th November, 1934.

446 975.—Clamps or supports for the electrode assembly of electron discharge devices.

Standard Telephones & Cables; D. H. Black; and W. T. Gibson. Application date 8th November, 1934.

448 418.—Valves with spherical electrodes for generating wavelengths of the order of millimetres. Circuits for radiating such wavelengths or for treating various substances with them.

E. E. W. Kassner. Convention date (Germany) 31st August, 1933.

449 727.—Carbon or graphite screens for separating the electrodes of gas-filled discharge tubes.

N. V. Philips' Co. Convention date (Germany) 9th October, 1934.

DIRECTIONAL WIRELESS

447 238.—Direction-finder in which a Neon lamp is used to give a visual indication of the bearings of a distant beacon station by a stroboscopic effect.

J. Marique and S.A. Internationale de T.S.F. Convention date (Belgium) 4th January, 1935.

447 273.—Direction-finder in which a pair of fixed frame aerials, set at an angle to each other, give a continual indication of the bearing of a distant beacon station.

Radio Navigational Instrument Corporation. Convention date (U.S.A.) 10th November, 1933.

450 975.—Blind landing system for aeroplanes in which television signals are transmitted non-directionally, and the glide-beam is swung up and down and modulated according to its angle of inclination.

Marconi's W.T. Co. and R. J. Kemp. Application date 25th January, 1935.

2 025 212.—Beacon for transmitting a rotating beam for air navigation purposes in which a centre fixed dipole is associated with two constantly-rotating dipoles which act as reflectors.

E. Kranar (assignor to C. Lorenz Akt.).

ACOUSTICS AND AUDIO FREQUENCY CIRCUITS AND APPARATUS

447 346.—Method for expanding the volume ratio of sounds reproduced by a gramophone or radio set.

F. Aigner. Convention date (Austria) 30th December, 1933.

449 874.—High-gain low-frequency amplifier free from background noises and valve " hiss."

L. H. Paddle. Application date 3rd January, 1935.

2 034 226.—Tone compensating circuits for low-frequency amplifiers.

E. F. Carter (assignor to United Research Corporation).

TELEVISION AND PHOTOTELEGRAPHY

447 046.—Preserving the high-frequency values in a short-wave television transmitter.

Radio Akt. D. S. Loewe. Convention date (Germany) 11th October, 1933.

447 312.—Gain control system for a television receiver in which there is no resulting unbalance between the high and low frequencies.

Radio Akt. D. S. Loewe. Convention date (Germany) 21st September, 1934.

448 111.—Applying symmetrical synchronising voltages to the deflecting electrodes of a cathode-ray tube from a saw-toothed oscillation-generator of unsymmetrical disposition.

A. C. Cossor and L. H. Bedford. Application date 28th November, 1934.

448 648.—Television system in which the picture and sound signals are separated from each other in the receiver by the so-called " aperture effect."

F. S. Turner. Application date 15th February, 1935.

449 177.—Time-base circuit for television, in which the storage condenser is located inside the cathode-ray receiver.

Marconi's W.T. Co.; L. M. Myers; and R. Cadzow. Application date 21st December, 1934.

450 303.—Synchronising in television systems where the picture " pick-up " apparatus is located at a distance from the radio transmitter.

Telefunken Co. Convention date (Germany) 26th March, 1935.

451 042.—Synchronising in television by using impulses of different amplitude.

Radio-Akt. D. S. Loewe. Convention date (Germany) 12th February, 1934.

451 786.—Phase-changing network suitable for synchronising a television receiver.

P. W. Willans and Baird Television. Application date 13th February, 1935.

451 980.—Portable television-receiver comprising a small cathode-ray tube which is mounted as an eye-piece for viewing the picture.

General Electric Co. and L. C. Jesty. Application date 15th May, 1935.

452 650.—Electrode arrangement for focusing the electron stream in a cathode-ray tube for television.

Radio-Akt. D. S. Loewe. Convention date (Germany) 27th November, 1933.

2 026 872.—Television receiver in which the Neon lamp fluctuations are first recorded upon a suitably-prepared photographic film from which the picture is projected in enlarged form.

L. de Forest (assignor to American Television Laboratories Inc.).

2 026 915.—Compensating for the negative-resistance characteristic, and the resulting tendency to self-oscillation, in a glow-discharge tube as used in television.

G. Schubert (assignor to Fernseh, A.G.).

SUBSIDIARY APPARATUS AND MATERIALS

450 686.—Piezo-electric crystal, coated with metal, and provided with line-gratings, to act as a " light " valve.

I. G. Farbenindustrie Akt. Convention date (Germany) 31st January, 1934.

451 664.—Moving-coil speaker in which a small central metal cone is provided to increase the high-note response.

British Thomson-Houston Co. and J. Moir. Application date 8th February, 1935.

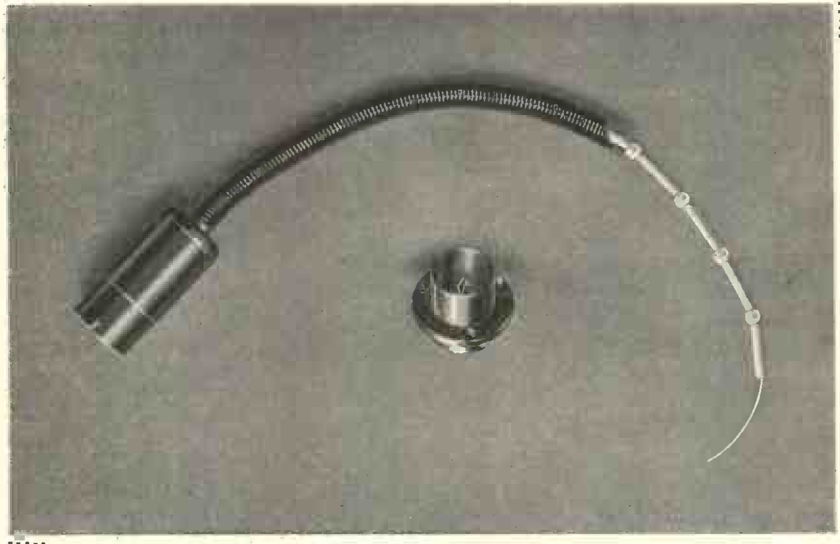
MISCELLANEOUS

446 804.—Tuning unit or " resonator " for very high-frequency currents.

Convention date (U.S.A.) 29th September, 1934.

2 027 526.—Holder for piezo-electric crystal-oscillator designed to permit the crystal being moved bodily to adjust its frequency.

R. E. Franklin (assignor to Radio Corporation of America).



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