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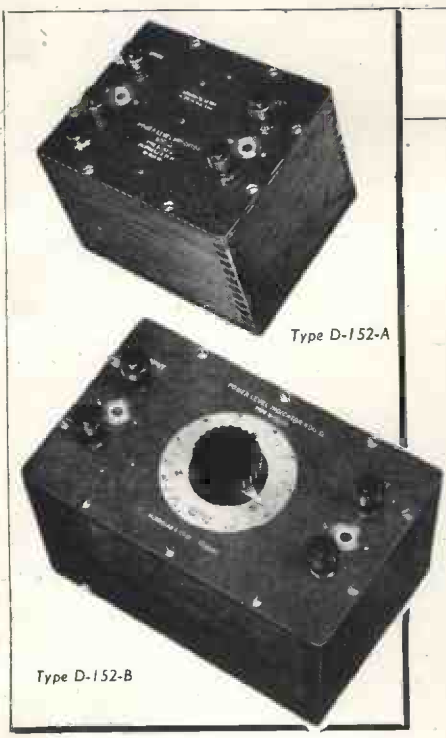

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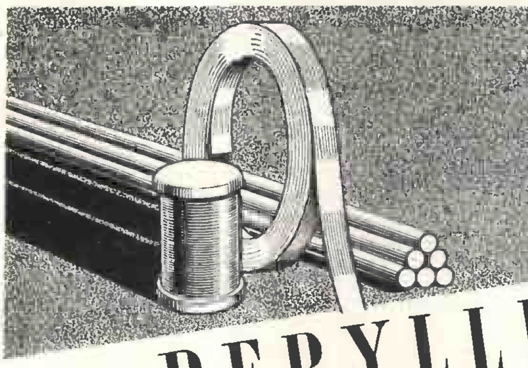

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WIRELESS ENGINEER

Editor HUGH S. POCOCK, M.I.E.E. Technical Editor Prof. G. W. O. HOWE, D.Sc., M.I.E.E.

VOL. XXI FEBRUARY, 1944 No. 245

Editorial

Coupled Circuits

WHEN explaining the meaning of coupling coefficient and the existence of two resonant frequencies in coupled circuits to junior students, it may be preferable to take a numerical example rather than to develop formulae with symbols. The following examples have been specially chosen for this purpose. The figures attached to the coils are the inductances in microhenries, those attached to the condensers are the *reciprocals* of the capacitances in microfarads, the frequencies are in megacycles per second. The advantage of using the reciprocals of the capacitances is that the values can be treated just like resistances or inductances when connected in series or parallel. One can put $S = 1/C$ and refer to it as the stiffness of the condensers; we then have $\omega^2 = S/L$. The method adopted is that first described in 1916*, in which the condensers are so charged with the circuits open that on simultaneously closing the circuits, there is no interchange of energy between the circuits, which simply oscillate as a single circuit. One can either neglect the losses or assume that both circuits have the same decrement; we shall neglect the losses. Fig. 1a shows two separate similar circuits with magnetic coupling. $S = 100$, $L = 10$, $M = 1$. If the two condensers are equally charged so that on discharge the inductances of the coils are naturally decreased, the effective inductance of each coil will be reduced from 10 to 9 and $\omega_1^2 = 100/9$; if the charge of one condenser is reversed, the effective inductance will be increased from 10 to 11 and $\omega_2^2 = 100/11$. For either circuit alone $\omega_0^2 = 100/10$. Defining the coupling coefficient k as the ratio M/L, $k = 1/10$ and we have

$$\frac{1}{\omega_1^2} : \frac{1}{\omega_0^2} : \frac{1}{\omega_2^2} = 9 : 10 : 11 = (1 - k) : 1 : (1 + k)$$

$\frac{1}{\omega_0^2}$ is thus the arithmetic mean of $\frac{1}{\omega_1^2}$ and $\frac{1}{\omega_2^2}$.

We also have $k = \frac{\omega_1^2 - \omega_2^2}{\omega_1^2 + \omega_2^2} = \frac{11 - 9}{11 + 9} = \frac{1}{10}$. This will be found a very useful way of defining k.

Fig. 1b is essentially the same as Fig. 1a, each circuit has a total inductance of 10, 1 of which is common to the two circuits. If the condensers are charged as shown in Fig. 1c, then on simultaneously closing the two circuits, the current will flow around the circuit as shown, the coupling coil acting like the galvanometer of a balanced

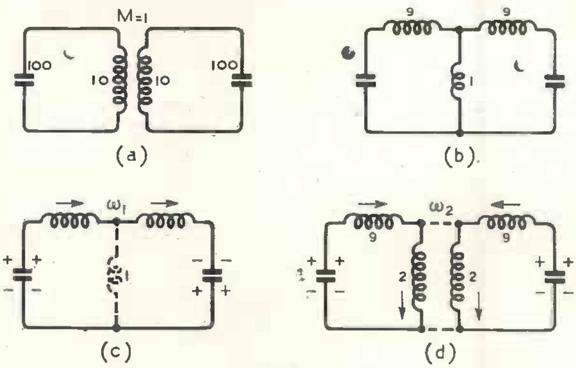


Fig. 1.

bridge and playing no part in the oscillation; $\omega_1^2 = 100/9$. If the condensers are charged as shown in Fig. 1d, then on discharge the current will flow as shown; we have pictured the coupling

* Howe: A new method of determining the frequencies and coupling coefficients in coupled oscillatory circuits. *Electrical World*, Vol. 68, p. 368.

coil replaced by two in parallel ; it is fairly obvious that $\omega_2^2 = 100/11$.

Figs. 2a, b, and c show the same procedure applied to a case of coupling by a condenser common to the two circuits. It must be remembered that values of S can be treated just like

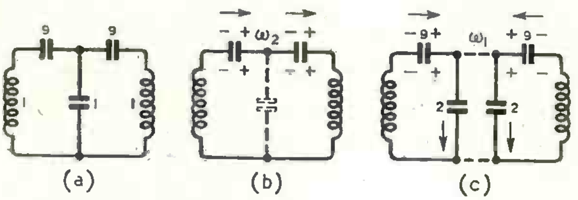


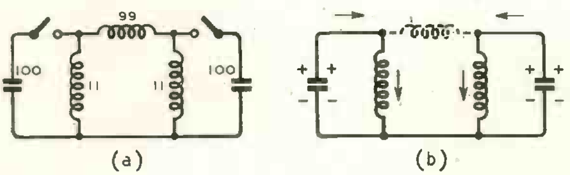
Fig. 2.

values of L and added when in series. With the values shown $\omega_0^2 = (9 + 1)/1 = 10$, $\omega_2^2 = 9/1$ and $\omega_1^2 = 11/1$ if we denote the higher frequency by the suffix 1. Hence in this case of capacitance coupling

$$\omega_2^2 : \omega_0^2 : \omega_1^2 = 9 : 10 : 11 = (1 - k) : 1 : (1 + k)$$

ω_0^2 is thus now the arithmetic mean of ω_1^2 and ω_2^2 ; but $k = \dfrac{\omega_1^2 - \omega_2^2}{\omega_1^2 + \omega_2^2} = \dfrac{11 - 9}{11 + 9} = \dfrac{1}{10}$ as before.

Fig. 3a shows another type of inductive coupling. It is important to note that to determine ω_0 for the circuit when uncoupled, it is necessary to open one of the switches shown and thus merely disconnect one the condensers, leaving all the three inductances in circuit. The three inductances must be regarded as common to the two circuits.



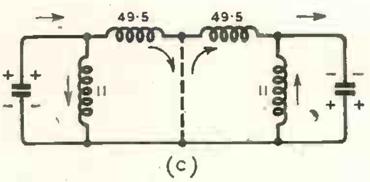


Fig. 3.

With the values shown $\omega_0^2 = 100/10 = 10$, since the inductance of 11 and 110 in parallel is 10. If the two condensers are charged as shown in Fig. 3b and then simultaneously discharged, we have $\omega_2^2 = 110/11$. If charged as shown in Fig. 3c and simultaneously discharged, the midpoint of the coupling inductance is at the same potential as the point opposite, and they could be connected as shown by the dotted line. Each condenser is thus discharging through two inductances of 11 and 49.5 in parallel, i.e. 9 and $\omega_1^2 = 100/9$. Hence the frequencies and the coupling coefficient are the same as in Fig. 1. The values of S and L were, of course, chosen to make this so, or, in other words, Fig. 3a was determined from Fig. 1b by the star-mesh transformation. If in Fig. 4a all the three elements are the same type, i.e. R, L or S then the star is equivalent to the mesh of Fig. 4b, in which the elements are of the same type as in the star and $A = \dfrac{ab + bc + ca}{a}$; to find B or C the denominator is put equal to b or c. In the present case $a = 1$, $b = c = 9$, hence $A = 99/1$ and $B = C = 99/9$, which are the values used in Fig. 3.

Fig. 5 shows an analogous capacitive coupling which is derived from Fig. 2a by star-mesh transformation. Here $\omega_0^2 = 10$, obtained by opening one of the switches in Fig. 5a ; $\omega_1^2 = 11/1$ from Fig. 5b and $\omega_2^2 = 9/1$ from Fig. 5c, since an S of 49.5 in parallel with one of 11 gives a resultant S of 9. Hence the frequencies and coupling coefficient are the same as in Fig. 2.

Dissimilar Tuned Circuits

In Fig. 6 the two coupled circuits are dissimilar, but are tuned to the same resonant frequency as

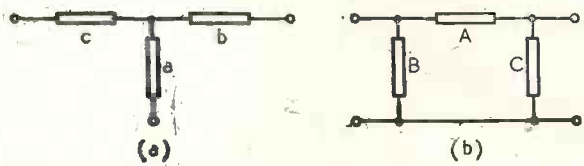


Fig. 4.

before, since in one circuit $\omega_0^2 = S/L = 40/4 = 10$ and in the other $\omega_0^2 = 250/25 = 10$. If $M = 1$ Fig. 6a can be replaced by Fig. 6b. The two resonant frequencies are not so obvious as they were in Fig. 1, since the circuits, excluding the coupling coil, are not now of the same resonant frequency, but the coupling coil can be replaced by two coils in parallel having inductances x_1 and x_2 such that their joint inductance is equal to that of the coupling coil, and such that the left-hand and right-hand circuits have the same resonant frequency. To find x_1 and x_2 we have

$$\frac{x_1 x_2}{x_1 + x_2} = 1 \qquad \frac{3 + x_1}{24 + x_2} = \frac{40}{250}$$

from which we find that

$x_1 = 3/5$ and $x_2 = -3/2$ or $x_1 = 7/5$ and $x_2 = 7/2$. These two cases are shown in Figs. 6c and 6d. If the condensers are charged to the correct amount and simultaneously discharged, there will be no P.D. between the points A and B, assuming equal decrements for the two circuits.

In Fig. 6c, we have

$$\omega_1^2 = 40/3\tfrac{3}{5} = 250/22\tfrac{1}{2} = 100/9$$

and in Fig. 6d

$$\omega_2^2 = 40/4\tfrac{2}{5} = 250/27\tfrac{1}{2} = 100/11.$$

Hence ω_0, ω_1, ω_2 and k are exactly the same as in Fig. 1. It will be noted that

$$k = \frac{\text{coupling or mutual inductance}}{\text{geometric mean of the total inductances}} = \frac{1}{\sqrt{4 \times 25}} = \frac{1}{10}.$$

It will also be noted that Figs. 1c and d are limiting

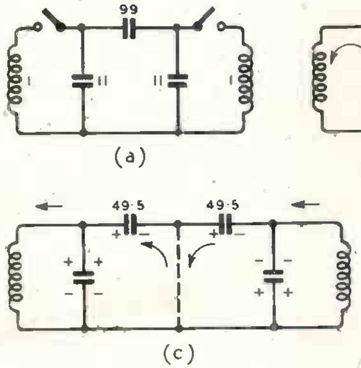


Fig. 5.

cases of Figs. 6c and d, for as the circuits approach similarity, x_1 and x_2 must either both approximate to 2 or they must both approximate to zero whilst preserving the condition $\frac{1}{x_1} + \frac{1}{x_2} = 1$, which means that the negative one must be the bigger of the two.

The negative inductance of $-3/2$ in Fig. 6c could be replaced by a condenser of capacitance 3/50, i.e. $S = 50/3$; then

$$\omega_1^2 = \frac{S}{L} = \frac{250 + 50/3}{24} = \frac{800/3}{24} = 100/9$$

as found above.

Fig. 7 illustrates the analogous case of capacitance coupling. Figs 7a, b, c correspond to Figs. 2a, b, c.

$$\omega_0^2 = \frac{S}{L} = \frac{4}{0.4} = \frac{25}{2.5} = 10$$

$$\omega_2^2 = \frac{3 + 3/5}{0.4} = \frac{24 - 1.5}{2.5} = 9$$

$$\omega_1^2 = \frac{3 + 7/5}{0.4} = \frac{24 + 3.5}{2.5} = 11$$

$$k = \frac{1}{\sqrt{4 \times 25}} = 1/10 \text{ as before.}$$

The negative S is equivalent to a positive L of 1/6; ω_2^2 would then be equal to $\frac{24}{2.5 + 1/6} = \frac{24}{8/3} = 9$ as above.

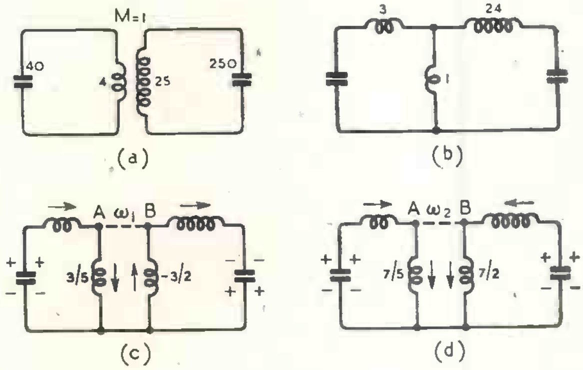


Fig. 6.

By means of the star-mesh transformation the star arrangement of Fig. 6b can be converted into the mesh arrangement of Fig. 8a. It is similar to Fig. 3a, except that the circuits are now dissimilar.

On opening the right-hand switch we have inductances of 132 and 33/8 in parallel which are equivalent to an inductance of 4; this gives $\omega_0^2 = 40/4 = 10$. Similarly on opening the left-hand switch we have inductances of 33 and $99 + 33/8$ in parallel which are equivalent to 25 giving $\omega_0^2 = 250/25 = 10$. As in Fig. 3c we must now divide the coupling inductance into two

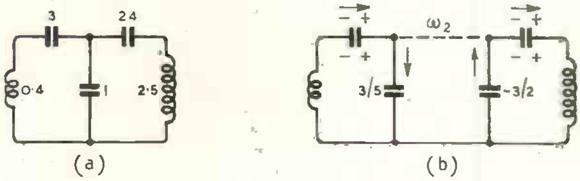
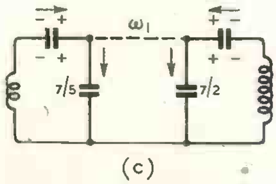


Fig. 7.



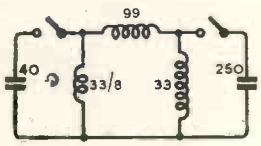
parts x_1 and x_2 such that $x_1 + x_2 = 99$ and the two halves of the system have the same frequency, the condition for which is that

$$\frac{x_1 \times 33/8}{x_1 + 33/8} = \frac{x_2 \times 33}{x_2 + 33} \times \frac{40}{250}$$

Solving these equations gives $x_1 = -66$ and $x_2 = 165$ or $x_1 = 198/7$ and $x_2 = 495/7$. These two solutions are represented in Figs. 8b and c. In each case the two circuits have the same natural frequency, and if the condensers are charged to the correct values and simultaneously discharged, the two circuits will oscillate without any interchange of energy. In Fig. 8b we have -66 in parallel with $33/8$ giving a joint L of $22/5$ and $\omega_2^2 = \frac{40}{22/5} = \frac{100}{11}$, or in the right-hand circuit, 165 in parallel with 33, giving a joint L of $55/2$ and $\omega_2^2 = \frac{250}{55/2} = \frac{100}{11}$.

In Fig. 8c, $198/7$ and $33/8$ have a joint L of $18/5$ and $\omega_1^2 = \frac{40}{18/5} = \frac{100}{9}$; similarly $495/7$ and 33 have a joint L of $45/2$ and $\omega_1^2 = \frac{250}{45/2} = \frac{100}{9}$.

Fig. 9 shows the result of applying the star-mesh transformation to Fig. 7a; it should be compared with Fig. 5a, which shows the symmetrical case.



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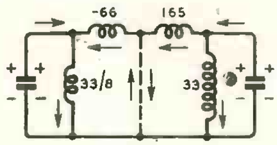
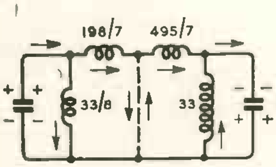


(b)

Fig. 8.



(c)

On opening the right-hand switch we have 132 and $33/8$ in parallel, giving a joint S of 4 and $\omega_0^2 = 4/0.4 = 10$. Similarly, on opening the left-hand switch we have 33 in parallel with $99 + 33/8$, giving a joint S of 25 and $\omega_0^2 = 25/2.5 = 10$. The coupling condenser is now divided into two in series as in the inductive coupling; the values of S are identical with those of L in Figs. 8b and c. In Fig. 9b we have

$$\omega_1^2 = \frac{22/5}{0.4} = \frac{55/2}{2.5} = 11$$

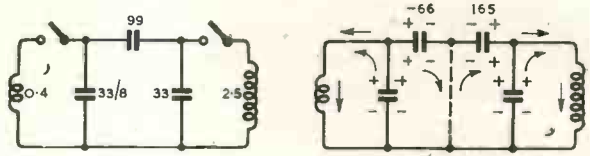
and in Fig. 9c $\omega_2^2 = \frac{18/5}{0.4} = \frac{45/2}{2.5} = 9$.

It should be noted that the relation between current and voltage in a negative L or S is in the opposite direction to that in a positive L or S. A negative inductance is equivalent to a positive capacitance and *vice versa*; hence the phase reversal.

In Figs. 8 and 9, by suitably adjusting the charges of the condensers before the simultaneous discharge, the two currents in the dotted connection can be made equal and opposite, and it is then unnecessary as it plays no part in the oscillations. Hence ω_1 and ω_2 are natural frequencies of the system.

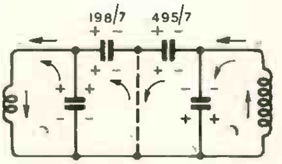
The Coupling Coefficient

In Fig. 4a the coupling coefficient is simply $\frac{a}{a+b}$ or $\frac{a}{a+c}$ when $b = c$; if b and c are not equal then k is the geometric mean of these two



(a) (b)

Fig. 9.



(c)

fractions, i.e.

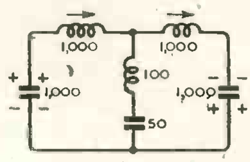
$$k = \frac{a}{\sqrt{(a+b)(a+c)}}.$$

In these expressions a, b and c represent either L or S.

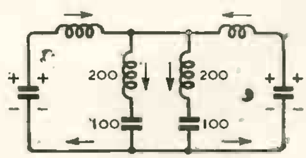
In Fig. 4b $k = \frac{B}{A+B} = \frac{C}{A+C}$ when B and C are equal; otherwise $k = \sqrt{\frac{BC}{(A+B)(A+C)}}$. This is obtained at once from the above by means of the star-mesh relationship $a = \frac{BC}{A+B+C}$ etc.

This definition of k can be applied to all the cases so far considered and gives 0.1 in every case for the values assumed for the components.

Fig. 10, however, shows a case of mixed coupling for which this definition of k is not applicable.



(a)



(b)

Fig. 10.

Its value can be found at once by determining ω_1^2 and ω_2^2 by the methods which we have described. By charging the condensers as shown in Fig. 10a the coupling arm plays no part and

$$\omega_1^2 = S/L = 1000/1000 = 1.$$

In Fig. 10b the coupling inductance of $100\mu H$ is replaced by two of 200 in parallel, and the coupling condenser of $0.02\mu F$ or $S = 50$ is replaced by two of $S = 100$ in parallel. On simultaneously discharging the condensers as shown we have

$$\omega_2^2 = \frac{S}{L} = \frac{1000 + 100}{1000 + 200} = \frac{11}{12}.$$

For the coupling coefficient we have

$$k = \frac{\omega_1^2 - \omega_2^2}{\omega_1^2 + \omega_2^2} = \frac{1 - 11/12}{1 + 11/12} = \frac{1}{23} = 0.0435$$

For either circuit alone in Fig. 10a we have

$$\omega_0^2 = \frac{1000 + 50}{1000 + 100} = 21/22$$

and

$$\omega_2^2 : \omega_0^2 : \omega_1^2 = 121 : 126 : 132$$

With capacitive coupling these formed an arithmetic progression, whilst with inductive coupling this was true of their reciprocals; with mixed coupling no such simple relationship is to be expected.

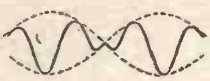
As stated at the beginning, this article is intended as a suggestion to those engaged in teaching the subject; it should give students a clear grasp of the origin of the two resonant frequencies and the resulting double hump in the resonance curve in all the cases of coupling likely to be met with in practice. It contains nothing that is not implicit in the 1916 article.

The next step would be the consideration of the effect of losses in the circuits; this we dealt with in the Editorial of June 1941.

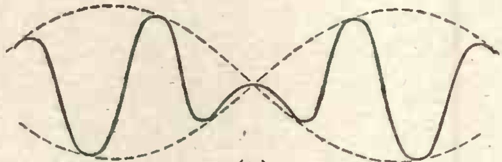
G. W. O. H.

Errata

Mr. H. L. Kirke of the B.B.C. has drawn attention to a detail of Fig. 3 in the December Editorial which is not correct. The centre part of the bottom curve is reproduced here (a); it shows the resultant as following the dotted envelope and coming to a sharp point as it passes through zero. It is not easy to show exactly what happens on this small scale, but it is shown on an enlarged scale in (b). Being the difference between two sine waves of different wavelength which both reach their equal maxima at the point under consideration, the resultant must pass horizontally through zero as shown.



(a)



(b)

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.

Ideal Filters

To the Editor, "Wireless Engineer"

Sir,—Ever since I read D. A. Bell's interesting article on "The Theory of Ideal Filters" in the July, 1943, issue of *Wireless Engineer* I have felt prompted to express my views on the subject. However, I had difficulties in visualising for the mere purpose of argument a realisable filter which would approach the requirements of idealised filters, until I came across M. Levy's paper on "The Impulse Response of Electrical Networks" in the December issue of the *Journ. I.E.E.* describing such an approach in the form of reflecting artificial lines, and giving examples of transients simulating those of idealised filters to an astonishing degree.

If I have understood Mr. Bell rightly he wants to demonstrate the inadequacy of the Fourier Integral for expressing the shock response of idealised filters by showing that these awkward ripples before the main rise of the response do not, after all, occur in reality. The author takes the example of a dissipationless low-pass filter consisting of *m* sections, and having as a response to unit function the integral of the Bessel function of the order *2m*. True enough this response does not show any oscillations before the main rise, none of those puzzling ripples resulting from the application of the Fourier integral, which apparently occur before unit function has been applied. But the quoted dissipationless low-pass filter does by far not fulfil the requirements of idealised filters, so that it is hardly to be expected that the solution (as a Bessel integral) for the filter discussed should agree with the result of analysing (as a Fourier integral) a circuit of entirely different properties.

Allow me, please, to put forth my views on this matter.

(1) The Fourier Integral Theorem is a mathematical conception which represents an arbitrary aperiodic function (with certain limitations) as the limit of a sum of a continuous spectrum of sine and cosine functions. This mathematical representation can thus be employed for the decomposition of an arbitrary function into a spectrum of componental sinoids, and vice versa, for the synthesis of a given spectrum, or certain parts of it, into an aperiodic function.

(2) If we so represent, for example, the well-known unit step function we obtain a hyperbolically distributed continuous spectrum of sine functions, including zero frequency of amplitude one half. It must be recalled that the unit step function, as well as its sinusoidal components, are defined along the abscissa in the interval minus infinity to plus infinity. If we now arbitrarily interpret the abscissa as time, the above analysis means that we have decomposed the unit function into an infinite number of frequencies. These sine waves (or frequencies) naturally have neither beginning nor end; they just exist in a certain phase and amplitude relationship to one another, and it would be futile to ask how they were introduced some undefinable time ago with a view to producing a unit step later on. At some instant between minus infinity and plus infinity, most conveniently in the finite region, the sinusoidal components add up to the unit step, and, at all other instants to the horizontal portions of the unit function. If, then, certain waves of

the spectrum are omitted—e.g., all frequencies higher than ω_0, and the rest of the spectrum is synthesised, the result can of course, no longer be the unit step, it actually is a "distorted unit step" known as $\frac{1}{2} + \frac{1}{\pi}$ Si $(\omega_0 t)$. The main rise of the Sine Integral function occurs at the same instant as the previous unit step.

(3) Now the telecommunication physicist steps in and postulates: this omission of a certain band of frequencies is the action of an idealised filter, hence $\frac{1}{2} + \frac{1}{\pi}$ Si $(\omega_0 t)$ is the response of this filter to unit function. And the dilemma is created at once: who produced the ripples before the unit step voltage was switched on? The reason for this dilemma can be found in the fact that the interpreting physicist did not take into account the fundamental relationship between the two components of the response of a network—*viz.*, amplitude and phase.

(4) An idealised filter must have infinitely steep flanks in order to cut off completely at one discrete frequency. This is generally recognised and accepted. But it is very often overlooked that such a filter axiomatically must also have a linear and infinitely steep phase characteristic, and hence, an infinite time delay. This can easily be seen when considering that such a filter is dissipationless, and consequently the phase change at resonance of one purely reactive filter element—there is an infinite number of them making up the filter—is sudden—i.e., steplike.

(5) The infinite time delay of the idealised filter fully explains the apparent dilemma—*viz.*, the response curve follows the impressed unit step after this infinite time delay. If the unit step occurs at, say, $t = 0$, the phases of the sine waves impressed at minus infinite time ago begin to arrive at the filter output at $t = 0$, and there produce infinitesimally small ripples—i.e., a horizontal line. As time progresses towards plus infinity these ripples of "frequency" ω_0 become larger until at time plus infinity—if one can at all talk about it—they become finite in amplitude, jump up by unity, and gradually decay again. The Fourier integral, however, does not tell anything about this infinite time delay; it merely sums up waves.

The fallacy in Mr. Bell's paper is the statement that "a filter with a large number of sections behaves like an ideal filter." Even an infinite number of sections of the filter chosen as an example could not form an idealised filter, for, neither is the phase curve linear in the pass range—it is a $\sin^{-1}$-curve; nor is the amplitude curve infinitely steep in the cut-off range—it is a $\cosh^{-1}$-curve; nor can such a filter be non-reflectively terminated in practice—the characteristic impedance is a pure resistance varying with frequency. The chosen filter is therefore certainly not an example against the practicability of the Fourier integral concept.

Furthermore, I cannot agree with Mr. Bell's statement that in an ideal filter "the amplitude of the oscillations in the response after the main rise increases with the number of sections, and with an infinite number of sections the oscillations would probably represent 100 per cent. modulation of the mean level." As a matter of fact, for an idealised filter, naturally comprising an infinite number of sections, the ripples decay according to a hyperbolic law as a very good approximation, and the first ripple does not overshoot the steady state value by more than nine per cent. of the total magnitude of the steady state step, which is proved by Gibbs' phenomenon (see any detailed book on Fourier's Theorem).

As explained in Mr. Levy's paper, filters can be constructed which, in their response to unit function, do contain any desired number of ripples before the main rise, and it is shown that the amplitude and phase characteristics of such filters really approach those of idealised filters to a very high degree. Besides, there seems to be no theoretical reason why this type of filter should not lend itself to the approach of idealised filters to any desired degree of accuracy. These reflection filters may be taken as a practical proof for the views laid down in paragraphs 1 to 5.

Quite generally it can be said that if one proceeds to such "unnatural" notions as idealisations one must be aware of, and courageously accept, "unnatural" consequences, such as, for instance, an infinite time delay.

Finally, I should like to mention that the derivation shown in the appendix to Mr. Bell's paper, though being fascinatingly short, is not correct. The expression

$$\phi(t) = \frac{2}{\pi} \sum_{m=0}^{\infty} \frac{\sin(2m+1)\,\omega_0 t}{(2m+1)\cdot\omega_0}$$

is wrong, and should read

$$\phi(t) = \frac{2}{\pi} \sum_{m=0}^{\infty} \frac{1}{2m+1} \sin(2m+1)\,\omega_0 t.$$

The erroneous ω_0 in the denominator comes from a wrong application of the general formula for the Fourier coefficients. If the ω_0 were correct it would mean that the amplitudes of the harmonic components $\frac{1}{\pi^2} \cdot \frac{T}{2m+1}$ would be proportional to the period T which evidently cannot be the case, for stretching the square wave of constant amplitude in the abscissa direction could not possibly alter the ordinates of the components for simple geometric reasons. Besides, the jump from equation (ia) to (ii) is unjustified, because the term $d\omega$ in the integral ought to find its counterpart as a $\Delta\omega$ somewhere in the expression for the series.

The derivation of the Fourier integral is not difficult, unless one goes to extremely rigorous mathematical reasoning. However, I do not believe that the derivation of the Fourier expression for unit function can be done along the lines suggested in Mr. Bell's paper—viz., by avoiding the Fourier integral proper.

London, W.2. G. L. HAMBURGER.

To the Editor "Wireless Engineer"

SIR,—I am glad Mr. Hamburger has taken up some of the mathematical points relating to the "ideal filter" concept, and on this ground I entirely agree with him; but we differ in our interpretation of the phrase "ideal filter." To Mr. Hamburger this means a filter which reproduces exactly the results obtained by the mathematical process explained in his paragraph (2); at the time of writing I was not aware that such a filter could be made, but it is now clear that there are at least two ways of making one. From the practical point of view, on the other hand, the "ideal filter" is the limiting case of a filter which could give the desired steady-state response if it were not for limitations on the pureness of the reactances employed and the number of sections; the practical value of any deductions from an ideal-filter theory depends upon their being an approximation to the results obtained with real filters. My purpose, then, was to point out that the postulate that "the omission of a certain band of frequencies is the action of an idealised filter" does not, in general, give us a correct approximation to the behaviour of real filters; this consideration at present applies to the vast majority of filters in practical use, but perhaps the realisation of filters which are ideal in the mathematical sense will bring into use filters having a better practical performance than those used hitherto.

London, N.21. D. A. BELL.

Valve Amplification Factor*

By H. Herne

SUMMARY—A new treatment of the electrostatic field of a triode is outlined, and it is shown to be valid for values of the grid wire diameter from zero to two-thirds of the grid pitch. A formula for the amplification factor of a planar triode is deduced, and it is shown to be the product of two factors, the first dependent on the grid geometry alone and the second on the anode to grid plane spacing; a graph is given for the rapid evaluation of the amplification factor of specific geometries. The extension of the method to other geometries and to multi-grid valves is indicated.

CONTENTS

1. Review of Previous Work

THE first attack on the general electrostatic problem of an infinite uniform wire grating seems to have been made by Maxwell.[1] He, himself, pointed out that his solution was limited to the case when the wire diameter was small compared to the pitch, and in the field map[2] he gives the wire diameter as less than 1/11 of the pitch. This method and these results were used by other workers, notably Abraham[3], *v.* Laue[4], King[5] and Miller[6] to deduce various approximations for the amplification factor of the triode, but all these results are limited to values of grid wire diameter less than about 1/10 of the grid pitch.

The above treatment was improved by Vogdes and Elder[7]. They used the same transformation as Maxwell, but made the perimeter of the transformed grid wire more accurately an equipotential by placing two line charges on the grid where Maxwell and the other workers mentioned above had used only one. By this device they were able to extend the range of usefulness of the transformation to values of the grid wire diameter from zero to about one-third of the grid pitch, but above this value the transformed grid wire becomes less of a circle in section and more kidney shaped. The two line charges, therefore, do not make the grid an equipotential in this case.

More recent work by Ollendorf[8] and Howland[9] has attacked the problem as a general one in potential theory, and by considering generalised boundary conditions the problem of any grid can be solved in infinite series, the physical significance of which is an acceptance of the transformation used by Maxwell with an infinite series of line charges to make the perimeters of the transformed electrodes equipotentials. It is tedious to obtain numerical results by this method as it necessitates a lengthy series of successive approximations to the result required.

* MS. accepted by the Editor, September, 1943.

2. Précis of Present Method and Results

The method of the present paper is to consider the shaded cell of the infinite uniform grid shown in Fig. 1. From this it is possible by an appeal to symmetry to build up the whole electrostatic field of the grid and this cell, therefore, is all that need be investigated in detail.

The transformation dealt with in detail in the following Section, 3, is intended to map the boundary $XACDX_1$ on the real axis of the transformed plane, i.e. on the real axis QR of the plane T in Fig. 2. The actual transformation used is shown to be reasonably accurate for values of grid wire diameter from zero to two-thirds of the grid pitch, whilst even when the grid wire diameter is three-quarters of the grid pitch, the maximum deviation is not more than 11 per cent. Fig. 2 is then a standard configuration and the field is determined by a conventional analysis for two cases, firstly when the grid is charged and, secondly, when the grid is uncharged but is in a uniform field. By superposition it is possible to build up any required case from these two basic ones.

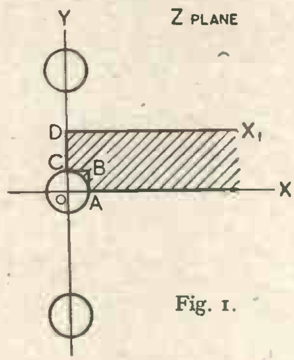


Fig. 1.

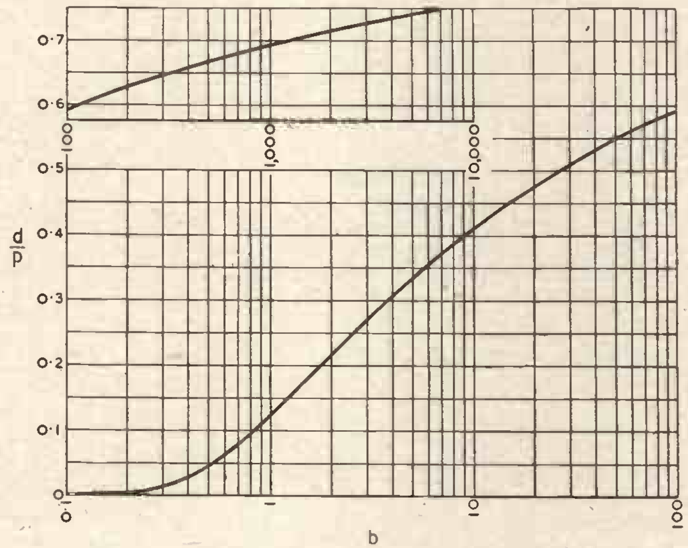
The amplification factor of a triode is defined in the conventional electrostatic manner as minus the change in anode voltage divided by the change in grid voltage to maintain a constant cathode field. It is assumed that a constant cathode field in the electrostatic case is equivalent to constant cathode current in the real valve. From the general equations for the field obtained above a value for the amplification factor may be deduced, and it is shown that this expression simplifies considerably to the product of two factors, one dependent on only the grid geometry and the other on the anode to grid plane spacing. The amplification factor (μ) may then be written thus

$$\mu = \frac{2\pi ab}{p}$$

where "a" is the anode to grid plane spacing, "p" is the grid pitch and "b" is a function of the grid geometry given by the curve of Graph 1. This expression for the amplification factor is shown to be valid when the cathode to grid plane spacing and the anode to grid plane spacing are both not less than the grid pitch. When the highest accuracy is required it is better to write the amplification factor as

$$\mu = \left(2\pi \frac{a}{p} - c\right)b$$

where "c" is a correction factor dependent only on the grid geometry and given by the curve of Graph 2; it is important only when the anode is close to a thick grid.



Graph 1.

3. Detailed Analysis of the Planar Triode

Consider a planar triode with the following dimensions:

a = anode to grid plane spacing
f = cathode to grid plane spacing
d = grid wire diameter
p = grid wire pitch

and let the grid be represented in Fig. 1, where $OD = p/2$ and DX' is parallel to OX.

The right angled semi-infinite polygon $XABCDX_1$ may be mapped on the upper half of the T plane by a standard Schwarz-Christoffel transformation. In Fig. 2 QR is the real axis of the T plane, and the point E is (1,0), F is (− 1,0) and G is (− k,0); then it is possible to make A transform to E, B to Q, C to F and D to G by the transformation

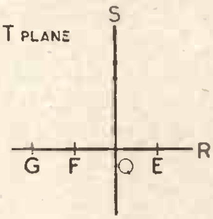


Fig. 2.

$$K\frac{dz}{dt} = (t-1)^{-1/2}\,(t)^{1/2}\,(t+1)^{-1/2}\,(t+k)^{-1/2} \qquad \ldots \quad \ldots \quad (1)$$

where K and k are constants determined by the boundary conditions.

Richmond[10] has shown that it is possible to round the corner ABC from a right angle to a smooth curve from A to C by substituting for $t^{1/2}$ in equation (1) the expression $(t-1)^{1/2} + n(t+1)^{1/2}$ where "n" is determined by the boundary conditions of the curve. Then equation (1) becomes

$$K\frac{dz}{dt} = (t+1)^{-1/2}\,(t+k)^{-1/2} + n(t-1)^{-1/2}\,(t+k)^{-1/2}$$

which integrates to

$$Kz = 2\tanh^{-1}\sqrt{\frac{t+1}{t+k}} + 2n\tanh^{-1}\sqrt{\frac{t-1}{t+k}} + C \qquad \ldots \quad \ldots \quad (2)$$

where C is the constant of integration.

To determine the constants K, k, n and C the values for the corresponding points in the Z and T planes are substituted in equation (2), giving

$$A \text{ to } E, \quad K\frac{d}{2} = 2\tanh^{-1}\sqrt{\frac{2}{k+1}} + C$$

$$C \text{ to } F, \quad K\frac{jd}{2} = 2jn\tan^{-1}\sqrt{\frac{2}{k-1}} + C$$

$$D \text{ to } G, \quad K\frac{jp}{2} = j\pi\,(n+1) + C$$

From these three equations we have

$$C = 0$$

$$K = 2\pi\,(n+1)/p \qquad \ldots \quad \ldots \quad (3)$$

$$k = \coth^2\left(\frac{\pi d(n+1)}{2p}\right) + \cot^2\left(\frac{\pi d(n+1)}{2np}\right) \qquad (4)$$

and n is given by the equation

$$\sin\left(\frac{\pi d\,(n+1)}{2np}\right) = \tanh\left(\frac{\pi d\,(n+1)}{2p}\right) \qquad (5)$$

This equation has been solved numerically for n, taking values of d/p from 0 to 3/4, and the result is given in Graph 3 and in Table I below as a matter of interest.

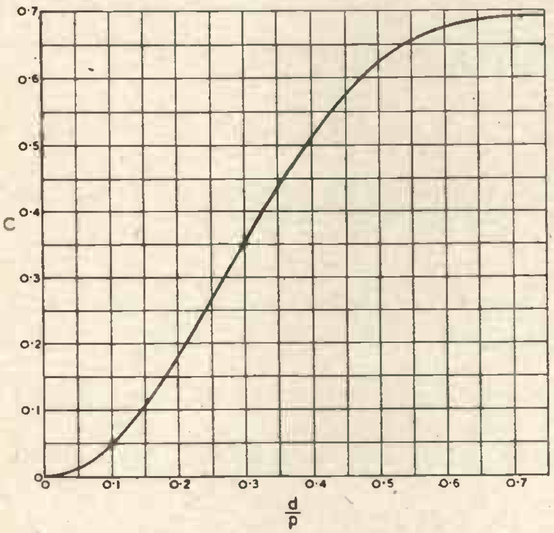
Thus it is possible to evaluate K, k and n for any grid geometry and therefore the transformation given by equation (2) is uniquely determined.

It is necessary to investigate how closely this transformation corresponds to the true circular

TABLE I

$\frac{d}{p}$	$\frac{\pi d}{2p}$	n	$\frac{d}{p}$	$\frac{\pi d}{2p}$	n
0	0	1.0000	0.4138	0.65	1.3313
0.0318	0.05	1.0016	0.4456	0.70	1.3963
0.0637	0.10	1,0067	0.4775	0.75	1.4710
0.0955	0.15	1.0149	0.5093	0.80	1.5572
0.1273	0.20	1.0264	0.5411	0.85	1.6580
0.1592	0.25	1.0426	0.5730	0.90	1.7759
0.1910	0.30	1.0619	0.6048	0.95	1.9165
0.2228	0.35	1.0850	0.6213	0.9759	2.0000
0.2546	0.40	1.1128	0.6366	1.00	2.0860
0.2865	0.45	1.1451	0.6685	1.05	2.2904
0.3183	0.50	1.1825	0.7003	1.10	2.553
0.3501	0.55	1.2254	0.7321	1.15	2.891
0.3820	0.60	1.2748	0.7639	1.20	3.35

grid wire. The simplest method is to use equation (2) to transform EF of the real axis of the T plane to a curve AC of the Z plane; if the distance of any point of AC from the origin O be denoted by R, the variation in R going along AC indicates how



Graph 2.

much AC deviates from a quadrant of a circle. The maximum variation in R for various values of d/p is given in Table II, and from this it is reasonable to deduce that the transformation is

TABLE II

$\frac{d}{p}$	$\frac{1}{10}$	$\frac{1}{5}$	$\frac{1}{3}$	$\frac{1}{2}$	$\frac{2}{3}$	$\frac{3}{4}$
Maximum value of R / Minimum value of R	1.0000	1.0004	1.004	1.018	1.063	1.11

satisfactory for values of d/p from 0 to 2/3 and may be used with discretion even to values of d/p equal to 3/4. As d/p increases beyond 3/4 the curve AC approaches the boundary ABC, and in the limit of d/p equal to 1 these two boundaries coincide.

The geometry of Fig. 2 can be treated by conventional methods. Consider first the electrical case when the grid is charged with a line charge of strength e per unit length of grid wire, and is in free space; then AC of Fig. 1 is an equipotential and AX and CDX are lines of force. The average surface charge densities induced on the anode and cathode are each $-e/2p$, and, from symmetry, πe lines of force are emergent from AC. Therefore, in Fig. 2 EF is an equipotential with πe lines of force emergent from it whilst the remainder of the real axis forms two boundary lines of force; Fig. 2 is then transformed to Fig. 3, so that the upper half of the T plane becomes the semi-infinite strip $V_I MNV_{II}$ with E transforming to M and F to N. The required transformation is another Schwarz-Christoffel given by

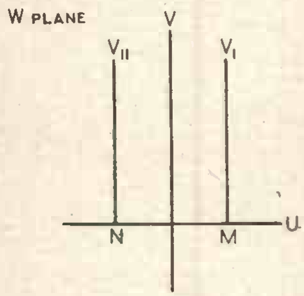


Fig. 3.

$$W = e \sin^{-1} T \quad \ldots \quad (6)$$

M is the point $(e/2,0)$ and N is $(-e/2,0)$. MN represents the grid wire surface, $V_I M$ the line of force AX and NV_{II}, the line of force CDX. Thus in Fig. 3 the lines "$V =$ constant" are equipotentials and the lines "$U =$ constant" are lines of force.

Suppose that the cathode of the valve transforms so that it cuts the real axis of the T plane at the point $(r_c, 0)$, and the anode at $(r_a, 0)$.

Equation (6) may be expanded to

$$r + js = \sin (U/e) \cosh (V/e) + j \cos (U/e) \sinh (V/e)$$

which gives the potential at any point $(r,0)$ of the real axis of the T plane as

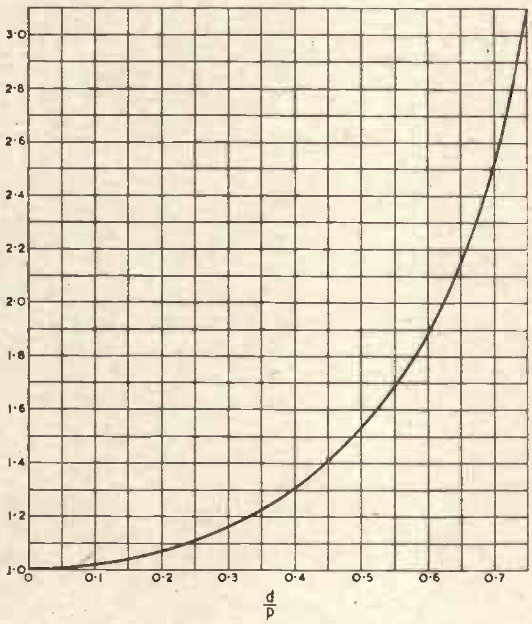
$$V_r = -e \cosh^{-1} r \quad \ldots \quad (7)$$

If V_g and V_a are the potentials of the grid and anode with respect to the cathode, then from equation (7)

$$V_g = e \cosh^{-1} r_c \quad \ldots \quad (8)$$

$$V_a = e \cosh^{-1} r_c - e \cosh^{-1} r_a \quad \ldots \quad (9)$$

The assumption made here is that the equipotential " V = constant " in Fig. 3 corresponds to an



Graph 3.

electrode surface " x = constant " in Fig. 1. This correlation has been investigated, and it is found that the transformation of an equipotential from Fig. 3 corresponds to such an electrode to better than 1 per cent. for all values of d/p from 0 to 3/4 so long as the spacing of the electrode from the grid plane is not less than the grid pitch.

The second electrical case to consider is when the grid is uncharged, but is in a uniform field due to a uniform surface charge density of $q/2p$ on the anode and $-q/2p$ on the cathode. In this case the line ACD of Fig. 1 is an equipotential and XA and DX_1 are lines of force, whilst πq lines of force pass through the boundary ACD. A of Fig. 1 must now transform to M of Fig. 3, and D to N; to consider the intermediary of Fig. 2 this means that E transforms to M and G to N. By considering a shift of the origin of the T co-ordinates, it can be seen that the transformation from Fig. 2 to Fig. 3 is given by

$$W = q \sin^{-1}\left(\frac{2T + k - 1}{k + 1}\right) \quad \ldots \quad (10)$$

which leads, as above, to expressions for the potential difference between electrodes as

$$V_g = q \cosh^{-1}\left(\frac{2r_c + k - 1}{k + 1}\right)$$

$$V_a = q \cosh^{-1}\left(\frac{2r_a + k - 1}{k + 1}\right) + q \cosh^{-1}\left(\frac{2r_c + k - 1}{k + 1}\right)$$

Superposing this case on the one above given by equations (8) and (9)

$$V_g = e \cosh^{-1} r_c + q \cosh^{-1}\left(\frac{2r_c + k - 1}{k + 1}\right) \quad \ldots \quad \ldots \quad (11)$$

$$V_a = e\,(\cosh^{-1} r_c - \cosh^{-1} r_a) + q\left\{\cosh^{-1}\left(\frac{2r_a + k - 1}{k + 1}\right) + \cosh^{-1}\left(\frac{2r_c + k - 1}{k + 1}\right)\right\} \quad \ldots \quad \ldots \quad (12)$$

and the surface charge density on the cathode is $-(e + q)/2p$. It is possible to rewrite equations (11) and (12) as

$$V_g = Ae + Bq$$
$$V_a = Ce + Dq$$

or

$$V_g = (A - B)\,e + B\,(e + q)$$
$$V_a = (C - D)\,e + D\,(e + q)$$

Solving for $(e + q)$,

$$(e + q)\begin{vmatrix} (A - B) & B \\ (C - D) & D \end{vmatrix} = \begin{vmatrix} (A - B) & V_g \\ (C - D) & V_a \end{vmatrix}$$

Differentiating with respect to V_g and keeping the cathode charge constant

$$0 = (A - B)\frac{dV_a}{dV_g} - (C - D)$$

But the amplification factor is given by $-\dfrac{dV_a}{dV_g}$ under this condition, hence

$$\mu = \frac{D - C}{A - B}$$

Substituting the values for A, B, C, D from equations (11) and (12)

$$\mu = \frac{\cosh^{-1}\left(\frac{2r_a + k - 1}{k + 1}\right) + \cosh^{-1}\left(\frac{2r_c + k - 1}{k + 1}\right) - \cosh^{-1} r_c + \cosh^{-1} r_a}{\cosh^{-1} r_c - \cosh^{-1}\left(\frac{2r_c + k - 1}{k + 1}\right)}$$

which approximates to

$$\mu = \frac{\log\left(\frac{2(2r_a + k - 1)2(2r_c + k - 1)\,2r_a}{(k + 1)\,(k + 1)2r_c}\right)}{\log\left(\frac{2r_c(k + 1)}{2(2r_c + k - 1)}\right)}$$

When a/p and f/p are greater than 1, r_c and r_a are both large compared to k, and so the above formula reduces further to

$$\mu = \frac{2 \log \dfrac{4r_a}{k+1}}{\log \dfrac{k+1}{2}} \quad \ldots \quad \ldots \quad (13)$$

To evaluate this expression in terms of the valve geometry, it is necessary to write equation (2) as

$$\frac{2\pi a(n+1)}{p} = 2 \tanh^{-1}\sqrt{\frac{r_a+1}{r_a+k}} + 2n \tanh^{-1}\sqrt{\frac{r_a-1}{r_a+k}}$$

i.e.,

$$\frac{2\pi a(n+1)}{p} = \log\left\{\left[\frac{1+\sqrt{\dfrac{r_a+1}{r_a+k}}}{1-\sqrt{\dfrac{r_a+1}{r_a+k}}}\right]\left[\frac{1+\sqrt{\dfrac{r_a-1}{r_a+k}}}{1-\sqrt{\dfrac{r_a-1}{r_a+k}}}\right]^n\right\}$$

Then approximately

$$\frac{2\pi a(n+1)}{p} = \log\left(\frac{4r_a+3k}{k}\right)^{n+1}$$

and $$\log\frac{4r_a}{k+1} = \frac{2\pi a}{p} - \log\frac{k+1}{k} \quad \ldots \quad (14)$$

It follows from equation (4) that

$$\frac{k+1}{2} = \left\{\coth^2\left(\frac{\pi d(n+1)}{2p}\right) + \cot^2\left(\frac{\pi d(n+1)}{2np}\right) + 1\right\}/2$$

By equation (5) this simplifies to

$$\frac{k+1}{2} = \operatorname{cosec}^2\left(\frac{\pi d(n+1)}{2np}\right)$$

whence

$$\log\frac{k+1}{2} = 2 \log \operatorname{cosec}\left(\frac{\pi d(n+1)}{2np}\right) \quad (15)$$

If the expression " $\log \operatorname{cosec}\left(\dfrac{\pi d(n+1)}{2np}\right)$ " be written as " $1/b$ ", and the expression " $\log\dfrac{k+1}{k}$ " as " c," then from equations (13), (14) and (15) the amplification factor may be written as

$$\mu = \left(\frac{2\pi a}{p} - c\right) b$$

" b " has been calculated for various values of d/p from 0 to 3/4 and the results are given in Table III and in Graph 1.

" c " also has been calculated and is given in Table IV and in Graph 2.

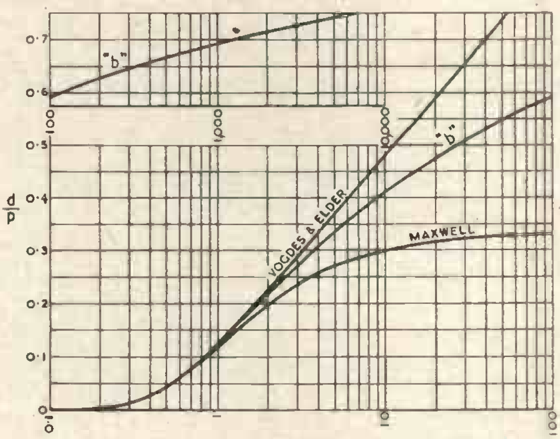
TABLE III

d/p	b	d/p	b	d/p	b
0	0	0.2865	3.42	0.5730	74.0
0.0318	0.435	0.3183	4.42	0.6048	127.5
0.0637	0.618	0.3501	5.76	0.6366	239.1
0.0955	0.815	0.3820	7.65	0.6685	514
0.1273	1.046	0.4138	10.35	0.7003	1254
0.1592	1.331	0.4456	14.31	0.7321	3874
0.1910	1.674	0.4775	20.35	0.7639	12600
0.2228	2.11	0.5093	29.94		
0.2546	2.68	0.5411	45.7		

TABLE IV

d/p	c	d/p	c	d/p	c
0	0	0.2865	0.3267	0.5730	0.6668
0.0318	0.0050	0.3183	0.3826	0.6048	0.6777
0.0637	0.0198	0.3501	0.4362	0.6366	0.6848
0.0955	0.0437	0.3820	0.4861	0.6685	0.6892
0.1273	0.0768	0.4138	0.5312	0.7003	0.6915
0.1592	0.1172	0.4456	0.5704	0.7321	0.6926
0.1910	0.1641	0.4775	0.6036	0.7639	0.6930
0.2228	0.2157	0.5093	0.6305		
0.2546	0.2705	0.5411	0.6514		

It will be noticed that as " a " is never less than " p ", the correction factor " c " is never more than 10 per cent., and may be ignored when a/p is large. The amplification factor may then be written as $\mu = \dfrac{2\pi ab}{p}$.



Graph 4.

If d/p is small a further approximation will reduce this formula to that given by Vogdes and Elder[7] mentioned above, and hence to that deduced from Maxwell's work ; in addition, the correction " log cosh " term in the Vogdes and

Elder formula does not differ from " c " above by more than 10 per cent. for values of d/p from 0 to 1/3, but for values greater than this the " log cosh " term increases indefinitely whilst " c " approaches an asymptote of $\log_e 2$. Graph 4 has been drawn to compare the result derived above for the amplification factor with those derived from Maxwell's work and by Vogdes and Elder; in each case only the factor corresponding to " b " above has been drawn, as this eliminates the effect of the anode.

4. Extension of the Method to Other Cases

The method outlined above may be applied to other geometries by altering the boundary conditions. For example, if the point C of Fig. 1 is taken to be $(0, g)$ the transformation given by equation (2) corresponds to an elliptic grid wire with axes of d and $2g$; thus any elliptic grid wire may be dealt with, and, in the limit, a strip grid either facing, or edge on to, the cathode. But in the limit it is obviously easier to use an unmodified Schwarz-Christoffel transformation for a semi-infinite rectangular box, as, for example, that given by equation (6).

The method may further be applied to cylindrical configurations by a preliminary transformation from the cylindrical case to that represented by Fig. 1. If the radius of the circle passing through the grid wire centres is R, and the angle subtended at the centre of symmetry by two adjacent grid wires is θ, then the required preliminary transformation is

$$\log \frac{W}{R} = \frac{\theta}{p} Z.$$

If the radius of the anode is R_a and the grid wire pitch and diameter are p and d respectively as above, then the substitution of $\frac{1}{\theta} \log \frac{R_a}{R}$ for $\frac{a}{p}$ gives the formula for the amplification factor in the cylindrical case immediately as

$$\mu = \left(\frac{2\pi}{\theta} \log \frac{R_a}{R} - c\right) b$$

where " b " and " c " are given by Graphs 1 and 2 respectively, as above. When θ is small it is often more convenient to write $\frac{p}{R}$ for θ in this formula, when the accurate expression for the amplification factor becomes

$$\mu = \left(\frac{2\pi R}{p} \log \frac{R_a}{R} - c\right) b$$

and the approximate formula becomes

$$\mu = \frac{2\pi R b}{p} \log \frac{R_a}{R}.$$

In practical units, using common logarithms, these are, accurately

$$\mu = (14.5 \frac{R}{p} \log \frac{R_a}{R} - c) b$$

and approximately

$$\mu = 14.5 \frac{Rb}{p} \log \frac{R_a}{R}.$$

Elliptic grid wires and strip grids in the cylindrical configuration obviously transform into elliptic and strip shapes in Fig. 1 by this logarithmic relation, and the above suggestions for treating these shapes in the planar case apply equally for the cylindrical one.

The method of adapting the results of a triode to multi-grid valves has been explained by Dow[11].

The author would be pleased to hear from any workers who have the opportunity to check the above theory with practical results for large values of d/p.

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February Meetings

THREE meetings of the Wireless Section of the Institution of Electrical Engineers have been arranged for February. The first is on Wednesday, 2nd, when Professor Willis Jackson, D.Sc., D.Phil., and L. G. Huxley, Ph.D., will give a paper on " The Solution of Transmission Line Problems by use of the Circle Diagram of Impedance." On Tuesday, 15th, a discussion on " Recording and Reproduction of Sound " will be opened by G. F. Dutton, Ph.D., B.Sc. " A Survey of the Problems of Post-War Television " will be given by B. J. Edwards at the meeting of the Section on Wednesday, 23rd. All the meetings commence at 5.30.

Professor J. D. Cockroft, F.R.S., chairman of the Electronics Group of the Institute of Physics, will lecture on the cyclotron and betatron at a meeting of the Group to be held at the Royal Society, Burlington House, Piccadilly, London, W.1, on Thursday, February 10th, at 5.30.

" A Review of Wide-Band Frequency-Modulation " is the subject of the paper to be given by C. E. Tibbs at the meeting of the British Institution of Radio Engineers at 6.30 on February 24th, at the Institution of Structural Engineers, 11, Upper Belgrave Street, London, S.W.1.

Temperature Coefficient of Capacitance*

Its Measurement in Small Radio Condensers

By W. Schick

SUMMARY.—The paper discusses the requirements for the measurement of temperature coefficient of capacitance of small radio condensers, and analyses possible errors with a view to defining conditions for the achievement of a certain accuracy. Various electrical and thermal methods are then reviewed as a preliminary to the description of an apparatus which was specially designed for the purpose.

Introduction

ORIGINALLY the temperature coefficient of capacitance was mainly a matter of concern in precision measuring apparatus, but as the question of frequency stability in radio engineering has received ever-increasing attention, there has arisen a need for components with improved stability, particularly with respect to temperature changes. The necessity for the accurate measurement of temperature coefficient of capacitance is, therefore, apparent. In addition, the introduction of ceramic condensers, consisting of rutile compositions which show decreasing permittivity with rise in temperature, has led to the adoption of temperature-compensated circuits and quantitative investigation of the behaviour of such condensers under varying temperature conditions has become important.

For this purpose special apparatus is required, because the usual methods for measuring capacitance provide neither sufficient sensitivity nor stability. In developing such measuring equipment, special consideration must be given to the mass production aspect of radio condensers, which require that a great number of tests shall be made in the shortest possible time.

General Requirements

It is general practice to define the temperature coefficient of capacitance of a condenser in parts per million, per unit capacitance, per degree centigrade. Thus :—

$$TC = \frac{\Delta C}{C \times \Delta t} \qquad \qquad (1)$$

Reference is also made sometimes to the temperature coefficient of permittivity of the dielectric which, in conjunction with the various coefficients of expansion involved, determines the temperature coefficient of capacitance in a given condenser.

It must be stressed that the temperature coefficient of a condenser cannot be completely defined unless the extent to which its behaviour is cyclic over short periods is determined and specified. The linearity of the relationship between capacitance change and temperature within a given temperature range is another factor which must be investigated and defined. It is on the assumption of cyclicity and linearity that the satisfactory functioning of temperature compensation is based.

The capacitance values in high frequency circuits where the temperature coefficient is of special significance can broadly be stated to range from 10 pF to 1000 pF. The temperatures under which these condensers are required to work depend on specific conditions but can be between —30 and + 70° C. To visualise clearly the requirements which have to be met by apparatus for the measurement of temperature coefficient, the table shown below gives the change of capacitance to be expected in extreme cases.

TABLE I

Type of Condenser	Capacitance pF	Temperature Coefficient $\times 10^{-6}$	Capacitance change for 30° C. change of temperature pF
Silvered Mica	100	+ 18	0.054
Rutile Ceramic	10	− 750	0.225
Clinoenstatite Ceramic	50	+ 120	0.180

The smallness of capacitance change will sufficiently illustrate the difficulties of measurement likely to be encountered, and it will be clear that a complete review of all factors introducing errors is necessary. The accuracy of measurement required depends very much on the purpose, but an error of 1 or 2×10^{-6} in the temperature coefficient will generally be permissible. On this assumption, the capacitance change alone of the examples cited in Table I must be accurately determined to within 0.006 pF, 0.0025 pF and 0.003 pF respectively, without allowing for any other errors. It follows from this that the apparatus

* MS. accepted by the Editor, July, 1943.

should yield reliable results to the nearest 0.001 pF.

Table II shows the possible causes of errors, including those having a thermal origin, deduced from a consideration of equation (1).

TABLE II

ΔC	Δt	C
1. Drift or instability of indicator during time for heating and cooling of condenser.	1. Temperature gradient in condenser.	1. Measurement of capacitance, especially low values.
2. Reading error determined by sensitivity of indicator.	2. Measurement of actual temperature of condenser.	
3. Calibration of capacitance change indicator.		

When reviewing the causes of errors, it must be remembered in the first place that temperature coefficient measurements involve the element of time. Whereas the readings can be taken within a few seconds, the time required for raising and lowering the temperature of the condenser, as well as ensuring complete thermal equilibrium at each reference level of temperature, may vary from a few minutes to one hour. The length of this period depends on the size of the object and also on the design of the thermal system. During this time the measuring apparatus may undergo changes due to extraneous causes such as variations in room temperature. If drift is suspected a stability check can be introduced before any readings are taken, but this method is rather cumbersome. It is better to eliminate any drift and check the stability by extensive observation under varying conditions before the apparatus is put into use.

The general rule applies that the sensitivity should be better than the calibration accuracy. If this is achieved the reading error will be negligible and it follows that a capacitance change of 0.001 pF will be shown up by the difference indicator.

The calibration accuracy of capacitance change depends on the electrical methods employed. The difference measurement will either be accomplished by a variable condenser which allows the reading of small capacitance increments, or by a calibrated audiofrequency oscillator. If first grade standards are used for calibration purposes, an accuracy considerably better than 1 per cent. can be achieved.

Error due to temperature measurement will depend on the temperature increment for which the capacitance change is ascertained. In such cases where linearity of the capacitance temperature relationship is established, large increments of 30°C to 40°C can be chosen, and if the temperature is determined to the nearest 0.1°C the resulting error will be smaller than 1 per cent. Investigations of linearity are, however, frequently required, and for this purpose increments of 5°C constitute an absolute minimum. A further problem arises in the location of the temperature indicator. Very accurate results can be obtained by attaching a thermocouple of suitable shape to the surface of the condenser. There are, however, disadvantages to this, mainly due to the fact that the thermocouple wires cause additional stray capacitances which undergo changes with temperature. If air is the medium surrounding the condenser and no temperature gradient exists in the chamber, the indicator can be placed at a distance from the test object. In such a case the use of a thermometer is convenient provided it is finely graduated and calibrated. Thermometers with graduation of 0.1°C and calibration to 0.05°C are very suitable.

Another important consideration is the possibility of the existence of a temperature gradient in the condenser. Its cause can be a gradient in the surrounding air which can be found by making temperature measurements at various points. Even if the temperature along the surface of the test object is uniform everywhere, there may still exist a gradient inside which will gradually disappear; therefore, until a state of thermal equilibrium is reached a capacitance drift will occur. This drift may be very slow on poor thermal conductors such as ceramics, and if readings are taken too early, the temperature coefficient measured will be too low.

The accuracy of the measurement of the capacitance C is related to the quality of the standards used for calibration. The limitation lies in the stray capacitances varying according to the shape of the condensers to be measured, and their position relative to the terminals. These variations amount to 0.1 to 0.3 pF and the lower the capacitance, the higher the percentage error will be. Above 100pF an accuracy of measurement of 0.2 per cent. presents no great difficulty.

Electrical Methods

To meet the requirements outlined, methods by which the small changes of capacitances are measured or indicated by changes of frequency offer many advantages[1]. However, other methods have been suggested which employ the resonance method and a valve voltmeter for indication[2].

The present paper describes modifications of the former method where the condenser under test controls a valve oscillator, the frequency change being determined by beating it against another oscillator, the frequency of which is kept constant during the measurement. These two oscillators will be referred to as "fixed" and "variable."

If the change of capacitance is small compared with the circuit capacitance, the change of frequency is expressed by the following equation:

$$\Delta f = -\Delta C \frac{f}{2C} \qquad \ldots \quad \ldots \quad \ldots \quad (2)$$

C being the total capacitance of the circuit. This expression naturally holds good only if the inductance of the circuit remains constant while the capacitance varies. In circuits for very high frequencies, mainly above 20 Mc/s, the inductive reactance of small radio condensers constitutes an appreciable part of the total circuit inductance. This inductive component also undergoes a change with temperature, and the relations become rather complex. For this reason capacitance measurements at radio frequencies are usually carried out around 1 Mc/s and the scope of this paper is limited to such frequencies.

Once the frequency change Δf has been detected by beating, it can be measured either by means of an audiofrequency bridge or by beating it against a calibrated audiofrequency oscillator. The comparison can best be performed in the latter case with a cathode ray oscillograph, where the Lissajous figure indicates either equality of the two audio frequencies or a certain ratio[3]. The computation of ΔC is then made by means of equations after the frequency f has been measured.

A method which dispenses with frequency measurement and computation employs a variable condenser of small capacitance and high reading accuracy to balance out the change in capacitance of the condenser under test and to restore the original frequency. In this case, an oscillograph or more simply a tuning indicator can be used for indicating zero beat between the variable and fixed oscillators. Precautions must, however, be taken to avoid pull-in between the two oscillators. To avoid this difficulty it is possible to set up a constant frequency difference between the two oscillators, and compare this beat frequency with an audio-oscillator of fixed frequency.[4]

Whatever modification of the beat frequency method is chosen, there is no difficulty in achieving sufficient sensitivity. For example, using the procedure last mentioned, a change of 0.0001pF can be indicated by a loud speaker. The actual problem lies rather in obtaining sufficient stability. A great deal has been published about the stability of valve oscillators. In this specific application the relevant factors contributing to drift are :–

Temperature ; Supply voltage ; Moisture ;

always providing that the mechanical construction is rigid and slow-ageing components are employed. One answer to the stability problem is to incorporate the two oscillators in a thermostatically controlled chamber, draw the supply voltages from batteries, and instal the equipment in a relatively dry atmosphere. The fixed oscillator can be of the quartz-controlled type,[4] or the transmission from a well monitored broadcasting station,[5] can be utilised. If drift is suspected in the variable oscillator, a check can be made by placing a high-grade variable condenser set to have identical capacitance with the condenser under test in the controlled chamber at the beginning of the measurement. While the temperature change is affected, the stability of the variable oscillator is observed, and then with a suitable switch, the condenser under test is brought back into the circuit to ascertain its capacitance change. Such a switch must be completely screened to exclude any residual capacitance between the two condensers.

It is not advisable to cover a wide range of capacitances with one frequency and one coil. If the test condenser is small, the capacitance of the circuit must be made up by a variable condenser. Any change in the capacitance of this condenser, or in the inductance of the coil for that matter, due to slight temperature variations, will be relatively large compared with the capacitance of the condenser under test. If the total circuit capacitance is 1000 pF, a change in the coil or the variable condenser of, say, 10 parts in a million will appear to be 1000 parts in a million on a 10 pF condenser. To overcome this difficulty, a number of coils or a variometer should be used to cover the capacitance range. It is, however, simpler to make the test frequency dependent mainly on the test condenser, and to adjust the fixed oscillator accordingly by a variable condenser.

Thermal Conditions

The thermal problems involved are of the same importance as the electrical ones, and unless very thorough consideration is given to them, errors will result.

One of the main difficulties is that the condenser to be subjected to temperature variation is connected to an oscillator, which must be kept at constant temperature. The electrical connection must, therefore, allow of only negligible heat transfer, and at the same time be stable in its self-capacitance. Moving objects in its locality, such as the operator, must not cause any change of capacitance in the line, and this requirement can only be met by screening. If the line is greater than a certain length, its inductance affects the accuracy of the measurement and corrections must be applied.

The heat-imparting medium is usually the air in a chamber made to surround the condenser. The air may be static, and its temperature variation can be achieved by altering the temperature of the chamber walls which may have a heater winding, or be immersed in a liquid the temperature of which is varied. Alternatively, the air in the chamber can either flow or circulate and be heated or cooled outside. The latter method is definitely preferable because it allows for speedy heat transfer to and from the test object and reduces the danger of temperature gradients. Small gradients can occur with an air stream, however, especially when air pockets are created by eddies, but baffles or suitable inlet and outlet arrangements will overcome this trouble. The function of the chamber should also be to exert a stabilising influence on the air current which may otherwise be excessively heated or cooled on its way through, thereby introducing gradients. If the chamber is built of thick copper sheet and is made no bigger than is required in relation to the size of the condenser, it will meet this requirement by acting as a stabilising circuit on account of its high thermal conductivity.

In actual routine work, temperature will mostly be measured on one point of the chamber, but in its design a check by means of thermocouples on temperatures at various points in the chamber and on the effect of using various shapes of condensers, is necessary. At the same time it must be remembered that the walls of the chamber and the terminals will expand and contract as well as the condenser itself. Steps must, therefore, be taken to ensure that the influence of varying stray capacitances due to these factors on the overall capacitance change should be negligible. This will again be critical when low capacitances are being considered.

Another factor to be considered is the thermal capacity of the system as a whole. When dealing with very small objects, which in themselves can be heated and cooled in a very short time, the total thermal capacity should be kept at a minimum to allow for quick changes and easy regulation of temperature. In a circulating system the blower fan is itself an object of considerable size, and it is, therefore, preferable to use a flow system supplied from an air compressor. High velocity air as well as a high level of heating and cooling energy make for great speed of operation.

It is well known that moisture films on insulators cause an increase of capacitance between the electrodes mounted on them, and these films can bring about very serious errors either by forming or drying during the measurement. The formation of the films is dependent on the relative humidity of the air, and providing this is not excessive in the atmosphere from which the pump draws, it can be reduced still further by elevating the air temperature. If, however, air with a certain moisture content is blown on a condenser which has been cooled down previously to low temperatures, a water film will be formed. The chemical drying of a rapid air stream is not always successful, and the freezing out of moisture presents the safest method of removal. Where the condenser bears a moisture film due to storage, it can be quickly dried up at the start of the experiment before readings are taken, for if the drying takes place during the test cycle, the condensers may wrongly appear to be non cyclic. This presents another advantage of the air current over still air.

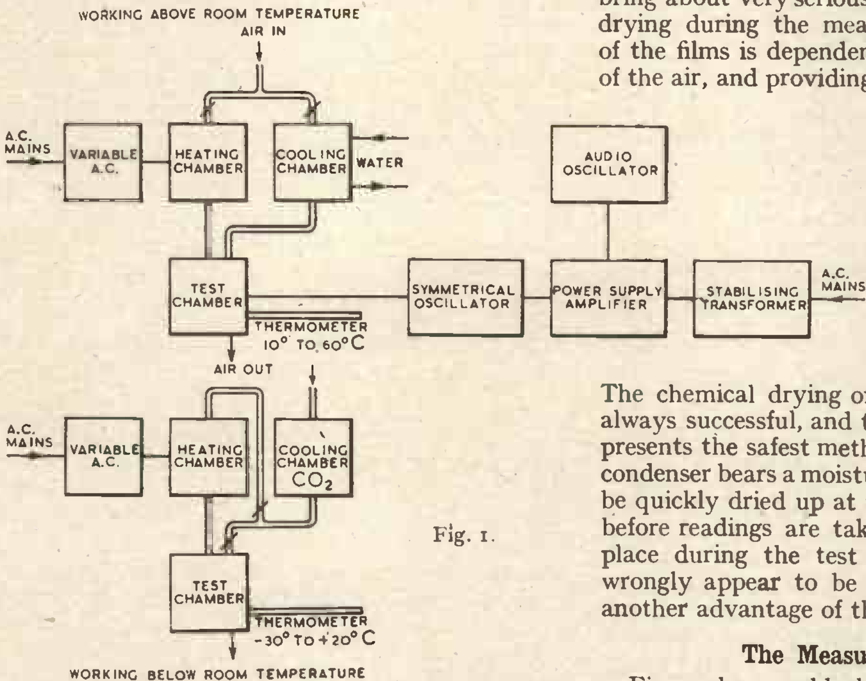


Fig. 1.

The Measuring Apparatus

Fig. 1 shows a block diagram of the apparatus which was designed for the measurement of temperature coefficient of capacitance of small radio condensers of the ceramic and silvered mica type. It covers a capacitance range from 10 pF to 1000 pF and allows temperature variations from $-20°$ to $+70°$ C. A capacitance change from

0.001 pF to 7.5 pF can be measured, and the test frequency varies between 0.9 and 2.8 Mc/s according to the test capacitance. A micrometer condenser

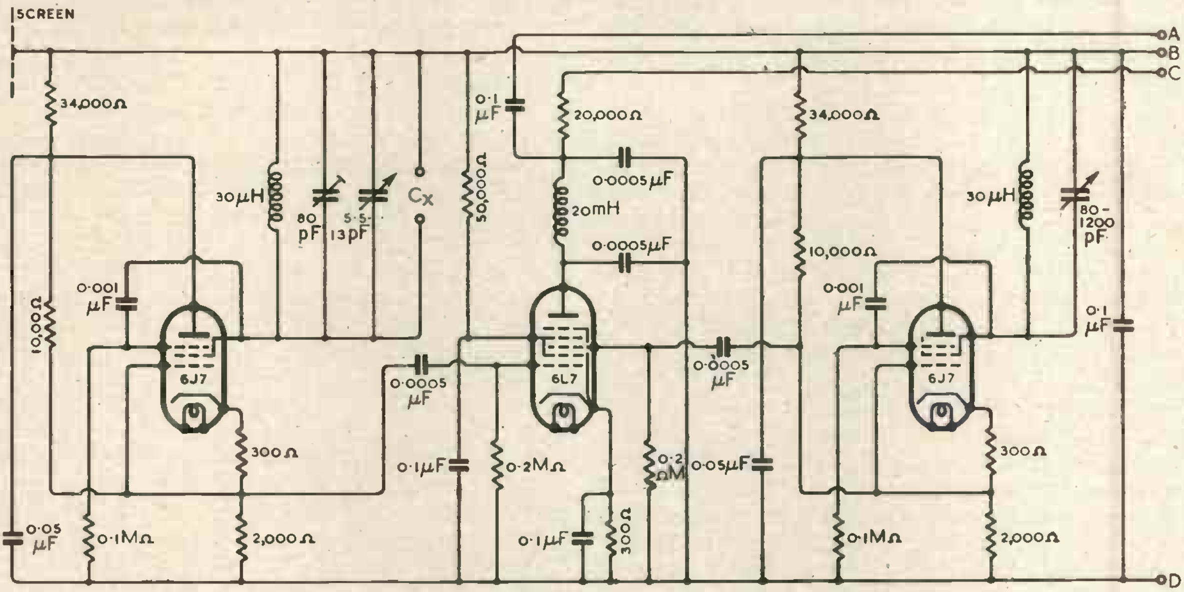
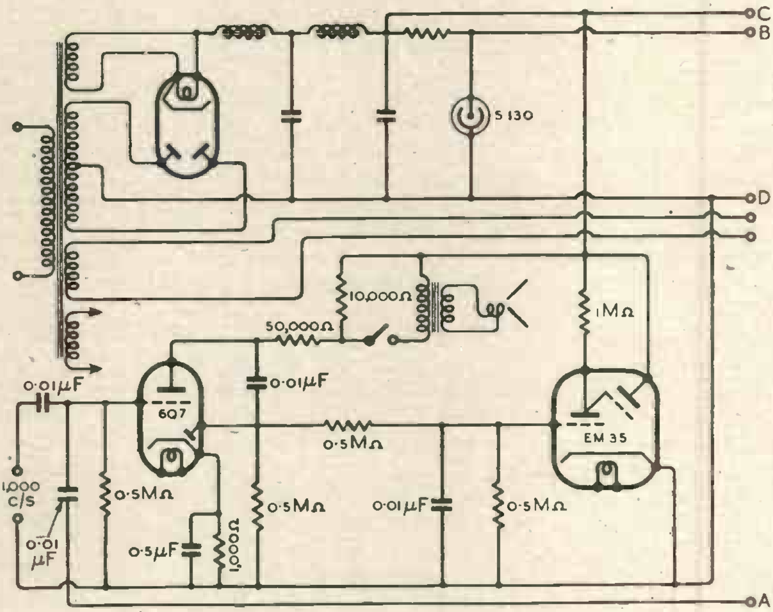


Fig. 2.

is used to balance out the capacitance change of the test condenser, and the double beat method is used for indicating the frequency change. The guiding principle in the design of the apparatus was simplicity and ease of operation; thermostating and battery supply were, therefore, ruled out though the accuracy referred to in previous sections was achieved.



To obtain a satisfactory stability both oscillators were constructed with identical components and supplied from one voltage source[6]. The object of this was to ensure that any frequency drift due to extraneous causes, i.e., room temperature and supply voltage, should not affect the difference frequencies of the two oscillators, since both are likely to drift in the same way. The advantages gained by such a symmetrical system can easily be demonstrated. In equation (2) ΔC is determined by Δf and f; assuming that the fixed oscillator remains absolutely constant during the measurement and that a stability of 0.001 pF is required, the variable oscillator must remain stable within 10 c/s if its frequency is 2 Mc/s and the circuit capacitance is 100 pF. If the frequencies of the two oscillators alter without influencing the beat note, a 2000 c/s drift, i.e. 0.1 per cent. of the total frequency, is permissible and the accuracy of the measurement is hardly affected. If zero beat were used for indication of capacitance balance, no error through drift would arise at all. In the apparatus described, the two oscillators are set at the beginning of the measurement at a frequency difference of 1000 c/s, and this beat note is checked by means of a stable audiofrequency oscillator. If under these conditions the two oscillators react equally to temperature and voltage change, a negligible error results as long as the frequency drift remains well within 1 per cent., which does not present great difficulties. It need hardly be stressed that the two oscillators must not only react equally to room temperature change, but also synchronously.

It has been maintained earlier that it is advantageous to make the frequency dependent mainly on the test condenser. This can easily be achieved with a symmetrical arrangement by providing the fixed oscillator with a variable condenser so that whatever the value of the test capacitance a 1000 c/s beat frequency is obtained. It follows that symmetry is not entirely complete, because the test condenser will be kept at controlled temperature whereas the variable one will be subject to room temperature fluctuations. It is well known that the temperature coefficient of a variable condenser with air dielectric depends on the insulating material used in its construction, and on the expansion coefficients of its various metal parts. At the minimum setting the former, and at the maximum setting the latter, will play a greater part. With a second grade variable condenser the insulating material will usually cause the temperature coefficient to vary inversely with the capacitance. In the apparatus described both oscillators have the same type of variable condenser, the one in the variable oscillator being set and fixed at its minimum of about 80 pF. This also gives stable electrical working conditions for the oscillator circuit when low capacitances of less than 100 pF are tested. Thus symmetry is improved and the remaining asymmetry depends only on the expansion coefficients of the variable condenser which does not cause any disturbing amount of drift.

The two oscillators and the mixing valve form one unit for it is clear that if the two oscillators are to react synchronously no temperature gradient must exist between them. They are enclosed in a heavy metal case which is lagged by Celotex to increase the thermal capacity of the unit. Changes in room temperature will, therefore, only slowly affect the oscillator components.

The circuit diagram is shown in Fig. 2. Transitron circuits are employed which have been dealt with elsewhere[7]. Otherwise the circuit is straightforward and calls for little comment. The two oscillator valves are carefully selected so that they react equally to changes of anode, and more particularly heater voltage.

The micrometer condenser C is of N.P.L. design, and has a capacitance swing of 7.5 pF. It consists of a metal cylinder in which a plunger is movable axially. One division on the micrometer head is equivalent to 0.0034 pF capacitance change. The calibration accuracy of the capacitance slope is better than 1 per cent. and linearity is also of this order. If backlash is allowed for 0.001 pF can easily be estimated. The variable condenser and coils are of the second-grade type.

Rectifier and amplifier form a separate unit. A tuning indicator allows beats lower than 1 c/s to be conveniently observed. Voltage fluctuations of the mains are reduced by two saturated transformers in cascade and a Neon stabiliser. Such variations as remain cause equal frequency changes in the two oscillators and so balance out. The 1000 c/s oscillator is of the mechanical type.

To enable a number of condensers to be measured during one heat cycle, arrangements have been described which consist of a rotatable switch incorporated in a chamber, which accommodates up to a dozen condensers[5]. It is clear that these condensers will all differ more or less in capacitance, and a beat note setting must be taken on each of them with the aid of a variable condenser having sufficient reading accuracy. The switch must also provide complete screening between the connected and unconnected contacts, otherwise serious errors may result. It is also necessary to take into account the temperature coefficient of the switch which may quite feasibly vary from position to position. A certain distance must also be allowed between the condensers so that expansion can cause no appreciable change in stray capacitances. Though these problems can be solved, it was decided not to use such an arrangement for a laboratory apparatus, particularly in view of the necessity for the accurate measurement of low capacitances, and to concentrate on reducing the time of the thermal cycle.

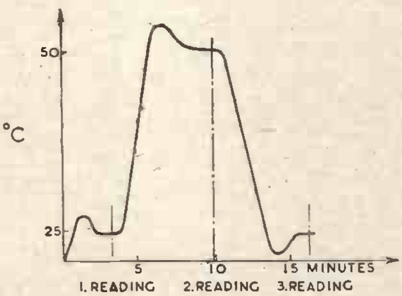


Fig. 3.

As shown in Fig. 1 an air flow system was employed, which draws air from a compressor and passes it through a heating chamber consisting of a wire cage with a consumption of 300 watts, or through a cooling chamber consisting of a copper tube helix immersed in flowing water or CO_2. When working at low temperature with CO_2 the thermal arrangement is altered somewhat by freezing out the mositure to avoid condensation on reheating the cooled condenser. By mixing or switching from one air current to another a fast temperature regulation can be affected. Heating from 25° C to 50° C for instance takes only two minutes. The temperature is indicated by a thermometer graduated in 1/10 of a degree C. The chamber is made of thick copper sheet 3in. × 3in. × 3in. and the gradient within was found to be less than 0.15°C between any two points.

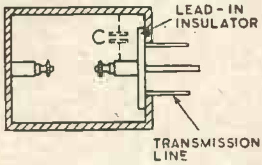


Fig. 4.

Fig. 3 shows a typical temperature/time curve obtained when testing a ceramic condenser. It can be seen that for fast work the reference levels of temperature are deliberately overshot either way to overcome the thermal gradient in the condenser which is a poor conductor. Thermal equilibrium in the object is indicated by stability of capacitance being reached. As can be seen in the diagram a whole cycle can be completed in 15 minutes.

The connection between chamber and oscillator consists of a rigid concentric transmission line of 13in. length and 1¼in. outer diameter. Because of its length and the perforation of the outer tube which acts as a complete electrical screen, no heat transfer of any consequence takes place. The inductance of the line is 0.12 μH and for the condensers above 200 pF capacitance, a correction is applied to take this into account.

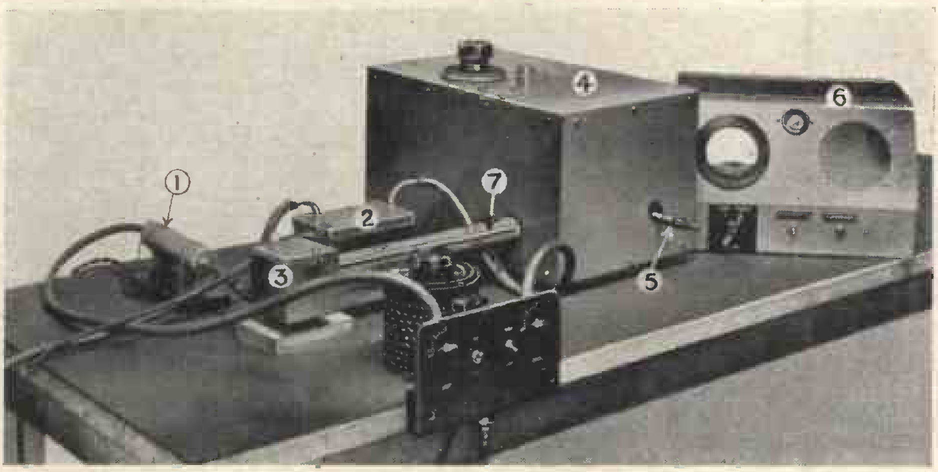


Fig. 5.—*The measuring apparatus described.* (1) *heating chamber;* (2) *cooling chamber;* (3) *test chamber;* (4) *oscillators;* (5) *micrometer condenser;* (6) *power supply and amplifier;* (7) *thermometer.*

Fig. 4 shows a cross section of the test chamber and from this it is obvious that the lead-in arrangement has a capacitance which will vary with temperature. The cause lies partly in the expansion of metal parts, and partly in the permittivity variation of the lead-in insulator. It was essential. therefore, to determine the amount of capacitance change per degree Centigrade and the cyclic behaviour. This was done by setting the two oscillators to beat by their internal capacitance without a condenser in the chamber. A heat cycle was produced and the capacitance change recorded. The apparatus showed a change of 0.003 pF per 10°C which is applied as a correction to each measurement. The influence of moisture films has been mentioned earlier and these form also on the lead-in insulator. Such films are likely to occur on unprotected ceramic material under damp conditions. A check on the cyclic behaviour of the chamber shows whether any instability due to a drying or condensed moisture film exists. Fig. 5 shows a photograph of the measuring apparatus. In front of the test chamber is the thermal control board with a Variac transformer which together allow heat regulation and control of the air current.

Conclusion

After using the apparatus for some time it was found that its stability was within 1/2 division of the micrometer condenser, this being equivalent to 0.0017 pF. The double beat method could most likely be replaced by the zero beat method, if decoupling between the two oscillators were improved.

The author desires to express his thanks to the United Insulator Company, for permission to publish this paper.

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The Phase Discriminator*

Its Use as Frequency-Amplitude Converter for F.M. Reception

By K. R. Sturley, Ph.D., M.I.E.E.

(Marconi School of Wireless Communication)

SUMMARY.—Since the phase discriminator (originally employed as the frequency error detector in an automatic frequency correction system) is able to convert a frequency change into a voltage amplitude change, it is particularly suited to act as the frequency to amplitude converter in a frequency-modulated receiver. When functioning as the latter, linearity of output voltage amplitude with change of frequency is all important, and maximum rate of change of amplitude with change of frequency (the requirement for automatic frequency correction) must be sacrificed for this.

Examination of the fundamental relationship between the primary and half secondary voltages of the phase discriminator shows that the primary voltage must increase as the applied frequency departs from the resonant or mid frequency, f_m, of primary and secondary circuits, if a linear characteristic is to be obtained over a range of off-tune frequencies. This entails a coupling coefficient greater than critical between primary and secondary.

Greatest range of linearity is obtained when $E_2/E_1 = 2$, $L_2/L_1 = 1$, for a coupling coefficient $k = 2/Q$, where Q is the magnification of the primary and secondary circuits. The linear range of off-tune frequency is from $\Delta f = 0$ to $\pm 0.5\frac{f_m}{Q}$.

Maximum frequency to amplitude conversion efficiency is obtained for couplings less than critical, i.e., $k < 1/Q$, and optimum design would appear to be obtained when $E_2/E_1 = 2$, $L_2/L_1 = 1.77$, $k = \frac{1.5}{Q}$, in spite of the reduction in the linear range of off-tune frequencies to $\Delta f = 0$ to $\pm 0.4\frac{f_m}{Q}$. A numerical example is given to illustrate the features of this optimum design condition.

Linearity of characteristic can be obtained at values of $E_2/E_1 > 2$ (but not for $E_2/E_1 < 2$) by reducing k, but the off-tune frequency linear range is progressively reduced as E_2/E_1 is increased.

1. Introduction

THE phase discriminator was initially developed as one unit of an automatic frequency correction (A.F.C.) system for correcting errors in the intermediate frequency carrier arising from inaccurate tuning or frequency drift of the local oscillator in a superheterodyne receiver. It acts as an error detector, converting a frequency error into a D.C. voltage (as shown by the curve *ABODE* in Fig. 1), which is subsequently used to control the bias of a variable reactance valve correcting the local oscillator frequency. Since this device converts a frequency change into an amplitude change it can be employed in a frequency-modulated (F.M.) receiver to convert a frequency modulated signal into an amplitude modulated one before application to a diode detector.

A typical circuit is shown in Fig. 2a. The secondary coil is centre tapped, the centre tap being connected to the high potential (anode) side of the primary by the coupling capacitance C_5. Two diodes D_1 and D_2 are used across each half of the secondary; the D.C. return path from the secondary centre tap to the junction of the two load resistances R_3 and R_4 is provided by the R.F. choke L_3. The primary voltage is developed across the latter, for the shunt capacitance C_4 across R_4 effectively earths point F with respect to radio frequencies. The total R.F. voltage applied to each diode therefore consists of the vector sum of the primary and one half secondary voltage. The equivalent circuit for the two diodes is that of Fig. 2b, where E_1 is the primary voltage across $L_1C_1R_1$ and E_2 the total secondary voltage across $L_2C_2R_2$.

The R.F. choke L_3 is not essential to the circuit and its chief value is in reducing damping on the primary circuit. This is important in an A.F.C. system, which requires a very small frequency separation (about 4 kc/s) between the positive and negative peaks (*D* and *B* in Fig. 1) of its frequency-voltage characteristic. For high-fidelity large frequency-deviation ($\pm$ 75 kc/s) F.M. signals, the frequency separation of *B* and *D* is of the order of 200 kc/s, and the circuits must be heavily damped

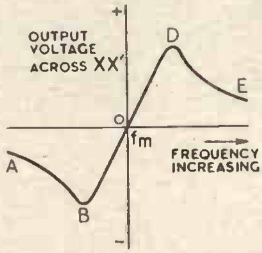

Fig. 1. *Typical voltage-frequency characteristic for a phase discriminator.*

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in order to achieve this. Under these conditions L_3 can be omitted and the secondary centre tap connected to the junction of R_3 and R_4. The two capacitances C_3 and C_4 are replaced by a single

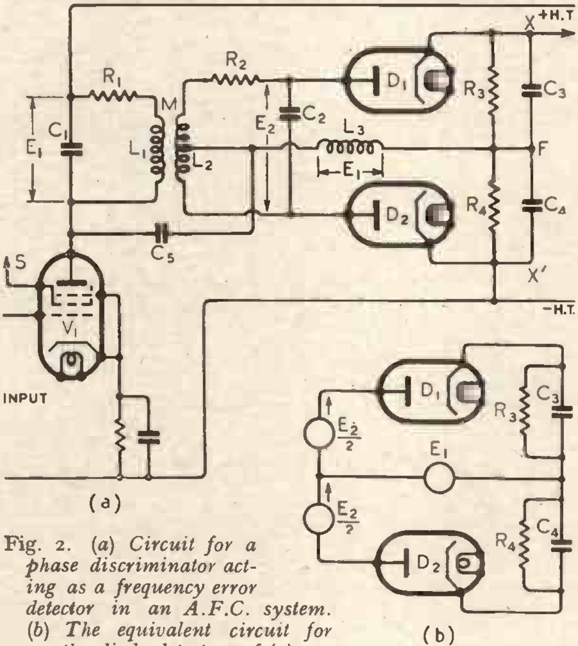

Fig. 2. *(a) Circuit for a phase discriminator acting as a frequency error detector in an A.F.C. system. (b) The equivalent circuit for the diode detectors of (a).*

capacitance C_3 joining the two diode cathodes as shown in Fig. 3; this change is essential, otherwise the primary voltage is short-circuited to earth. The primary voltage is developed across R_3 for diode D_1 and across R_4 for diode D_2, and the two resistances are now in parallel across the primary circuit, thus adding to its damping.

The characteristics required of a phase discriminator when it is to function as a frequency-amplitude converter are not exactly the same as those needed for A.F.C. purposes. Linearity of output voltage amplitude with change of frequency is all important, and maximum rate of change of amplitude at f_m, point O in Fig. 1 (the requirement for A.F.C.) must be sacrificed for this. Theoretical examination of discriminator performance shows that a coupling coefficient greater than that giving a maximum rate of change of amplitude is required.

2. Theory

To determine the shape of the discriminator characteristic and to estimate the conditions for linearity of conversion over a desired frequency range, it is necessary to develop the fundamental expressions for the primary E_1, and secondary E_2 voltages. This is done in Appendix Ia, where it is shown that the primary voltage may be expressed as

$$E_1 = g_m E_g R_{D1} \frac{1 + jQ_2F}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2k^2} \quad (1)$$

and in Appendix Ib, where the secondary voltage is given as

$$E_2 = g_m E_g R_{D1} \frac{- jQ_2k\sqrt{L_2/L_1}}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2k^2} \quad (2)$$

where g_m = mutual conductance of the amplifier valve V_1 before the phase discriminator.

E_g = input voltage to the valve V_1.

R_{D1} = resonant impedance of the primary of the phase discriminator $= \omega_m L_1 Q_1$.

Q_1 = magnification of the primary circuit including damping from the diodes, R_1, and slope resistance of V_1, but excluding that due to coupling to the secondary.

Q_2 = as for Q_1 applied to secondary.

$k = \dfrac{M}{\sqrt{L_1L_2}}$ = coupling coefficient.

$F = \dfrac{2\Delta f}{f_m}$

Δf = off-tune frequency from the mid frequency f_m.

The total voltage applied to diode D_1 is

$$E_{AF} = E_1 + E_2/2$$

and to the other D_2 is

$$E_{BF} = E_1 - E_2/2$$

and the output voltage of the discriminator across points XX' in Fig. 3 is the numerical difference

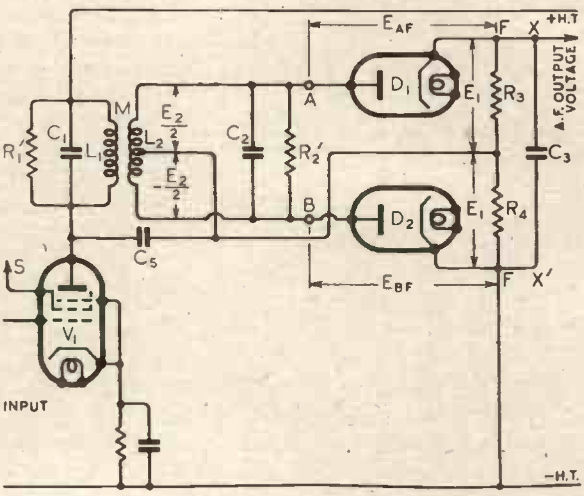

Fig. 3. *The phase discriminator acting as a frequency-amplitude converter in a frequency-modulated receiver.*

between the amplitudes of E_{AF} and E_{BF}, multiplied by the voltage detection efficiency, η_d, of the diodes. Therefore

$$E_{XX} = \eta_d(|E_{AF}| - |E_{BF}|)$$
$$= \eta_d(|E_1 + E_2/2| - |E_1 - E_2/2|) \quad \ldots \quad \ldots \quad (3)$$

The secondary voltage E_2 can be rewritten in terms of E_1 as

$$E_2 = E_1 \frac{-jQ_2k\sqrt{L_2/L_1}}{1 + jQ_2F} \quad \ldots \quad \ldots \quad (4)$$

so that $E_{AF} = E_1\left(1 - \dfrac{jQ_2k\sqrt{L_2/L_1}}{2(1 + jQ_2F)}\right)$

or $$\frac{E_{AF}}{E_1} = 1 + \frac{\alpha Q_2F}{2(1 + Q_2^2F^2)} - \frac{j\alpha}{2(1 + Q_2^2F^2)} \quad \ldots \quad \ldots \quad (5)$$

where $\alpha = Q_2k\sqrt{L_2/L_1} = E_2/E_1$ at $F = 0$, i.e., at the mid frequency f_m

$$\frac{|E_{AF}|}{|E_1|} = \sqrt{\left[1 + \frac{\alpha Q_2F}{2(1+Q_2^2F^2)}\right]^2 + \left[\frac{\alpha}{2(1+Q_2^2F^2)}\right]^2} \quad \ldots \quad \ldots \quad (6a)$$

Similarly

$$\frac{|E_{BF}|}{|E_1|} = \sqrt{\left[1 - \frac{\alpha Q_2F}{2(1+Q_2^2F^2)}\right]^2 + \left[\frac{\alpha}{2(1+Q_2^2F^2)}\right]^2} \quad \ldots \quad \ldots \quad (6b)$$

The vector relationship represented by expression (5) is shown in Fig. 4 for $E_2/E_1 = 2$, and the variation in length of $|E_{AF}|/|E_1|$ can be measured from this figure for different values of Q_2F and selected values of α, or alternatively it may be calculated from (6a). $|E_{BF}|/|E_1|$ can be found in a similar manner and subtracting it from $|E_{AF}|/|E_1|$ gives the relative voltage output from the discriminator if E_1 is a constant amplitude. It is plotted against Q_2F from 0 to +2 for different values of α (E_2/E_1) in Fig. 5. The other half of the curve from $Q_2F = 0$ to -2 is the same shape inverted. Two points arise from an examination of the curves: maximum output voltage occurs at a larger value of Q_2F as E_2/E_1 is increased, and none of the curves is truly linear. A second set of curves is drawn in Fig. 6 to show the departure from linearity, which is expressed in the form of ten times the ratio of the relative voltage output at a particular value of Q_2F to the voltage output at $Q_2F = 0.1$, multiplied by the value of Q_2F being considered. The general tendency is for the curve to become more linear as E_2/E_1 is increased, though there is not much difference between the curves for $E_2/E_1 = 4$ and 6.

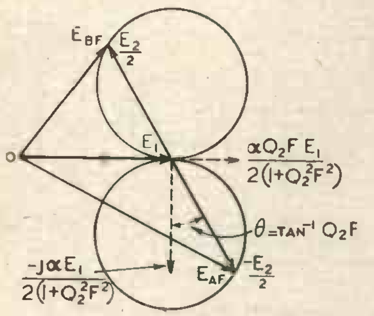

Fig. 4. *Vector relationship for the voltages in a phase discriminator.*

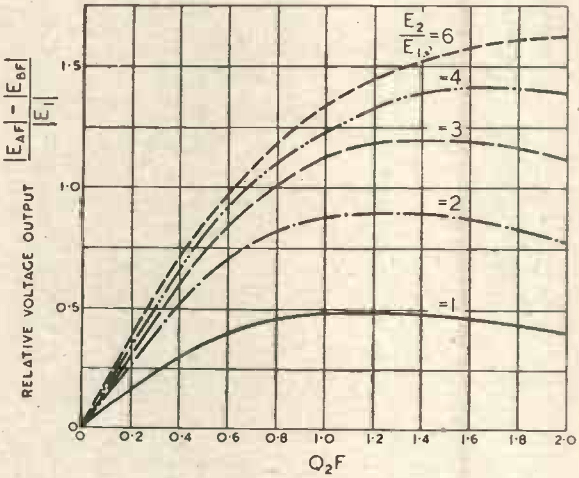

Fig. 5. *Relative voltage output from a phase discriminator for different off-tune frequencies.*

The divergence of these curves from the straight line can be offset if E_1 can be made to increase with increase of Q_2F, and this is possible by a suitable choice of coupling between secondary and primary. If it is assumed that $Q_1 = Q_2$, a series of curves of the ratio increase of E_1 from its value at f_m ($F = 0$) can be plotted against QF for different values of coupling factor Qk. The problem is greatly complicated if Q_1 and Q_2 are not equal because a separate set of curves must be drawn for each value of Q_1 and Q_2, and there is seldom any practical advantage to be gained by making them unequal. The ratio increase of E_1 from its value at f_m is from expression (1).

$$\frac{|E_1|}{|E_1|_{fm}} = \frac{(\sqrt{1 + Q^2F^2})(1 + Q^2k^2)}{\sqrt{[1 + Q^2(k^2 - F^2)]^2 + 4Q^2F^2}} \qquad (7)$$

and it is plotted in Fig. 7 as a series of curves for selected values of Qk. To find the most suitable value of Qk for compensating the ratio decrease in Fig. 6, the curves of Fig. 7 are drawn on transparent paper and placed on top of Fig. 6. Any two curves which coincide from $QF = 0$ give the conditions for a linear characteristic over the range of QF for which they are coincident. Greatest range of QF over which a linear characteristic is obtained is from 0 to approximately 1 with $E_2/E_1 = 2$ and $Q_2k = 2$. The resultant characteristic is linear up to $QF = 0.8$, and falls away slightly at $QF = 1$ where the output is about 2 per cent. low. A slightly lower value of Qk could be used with some reduction of the linear range of QF, and this has the advantage of giving a higher frequency-amplitude conversion efficiency. Thus for $E_2/E_1 = 2$, $Qk = 1.5$, the linear range is reduced to $QF = 0$ to 0.8, but conversion efficiency (see expression 8b) is raised in the ratio.

$$\frac{1 + Q^2k^2}{1 + Q_1^2k_1^2} = \frac{5}{3.25} = 1.538.$$

Fig. 8 shows the discriminator characteristic for the two values of Qk, 1.5 and 2 when $E_2/E_1 = 2$. It is seen that the value of QF at which maximum output is obtained is increased from 1.5 to 1.9 when Qk is increased from 1.5 to 2. It is interesting to note that $Qk = 2$ gives the maximum possible correction over the useful range of QF, and a linear characteristic cannot be obtained for values of E_2/E_1 less than 2. Correction is possible for values of E_2/E_1 exceeding 2; lower values of Qk are required and the linear range of QF is reduced, e.g., $E_2/E_1 = 3$ requires Qk to be 1.25 and the linear range of QF is from 0 to 0.85.

The frequency-amplitude conversion efficiency is measured by the slope of the characteristic at f_m, and it is obtained by differentiating (3) with respect to Δf, and equating all terms containing Δf to zero.

Thus conversion efficiency at f_m is

$$\frac{dE_{xx}'}{d\Delta f} = \frac{d}{d\Delta f}\left[g_m E_g R_{D1}\eta_d \frac{\left[1 + Q_2^2\left(\frac{2\Delta f}{f_m} + \frac{k}{2}\sqrt{\frac{L_2}{L_1}}\right)^2\right]^{\frac{1}{2}} - \left[1 + Q_2^2\left(\frac{2\Delta f}{f_m} - \frac{k}{2}\sqrt{\frac{L_2}{L_1}}\right)^2\right]}{\left[\left[\left(1 + Q_1Q_2\left(k^2 - \left[\frac{2\Delta f}{f_m}\right]^2\right)\right)\right]^2 + (Q_1 + Q_2)^2\left[\frac{2\Delta f}{f_m}\right]^2\right]^{\frac{1}{2}}}\right]$$

$$= \frac{2g_m E_g R_{D1}\eta_d}{f_m} \frac{Q_2^2 k\sqrt{L_2/L_1}}{(1 + Q_1Q_2k^2)\left(1 + \frac{Q_2^2k^2L_2}{4L_1}\right)^{\frac{1}{2}}} \quad \text{.. .. (8a)}$$

The condition for the maximum value of efficiency with variation of k is found by differentiating (8a) with respect to k and equating to 0 when

$$\sqrt{Q_1Q_2}\,k = \sqrt{\frac{\sqrt{Q_1^2Q_2^2 + 2Q_1Q_2^3\,L_2/L_1} - Q_1Q_2}{Q_2^2\,L_2/L_1}} \quad \text{.. .. (9a)}$$

or when $Q_1 = Q_2$

$$Qk = \sqrt{\frac{\sqrt{1 + 2L_2/L_1} - 1}{L_2/L_1}} \quad \text{.. (9b)}$$

which is always less than 1 whatever the value of L_2/L_1. However, it has already been shown that for linearity of characteristic Qk must be greater than 1, so that expression (9b) indicates that the smallest possible value of Qk should be used commensurate with adequate linear range of QF.

Expression (8a) may be rearranged in terms of α, the ratio of E_2/E_1 at f_m, as follows

$$\frac{dE_{xx}'}{d\Delta f} = 2g_m EgR_{D1}\eta_d \frac{Q_2(E_2/E_1)_{fm}}{f_m(1 + Q_1Q_2k^2)[1 + \frac{1}{4}(E_2/E_1)^2_{fm}]^{\frac{1}{2}}} \quad \text{.. .. (8b)}$$

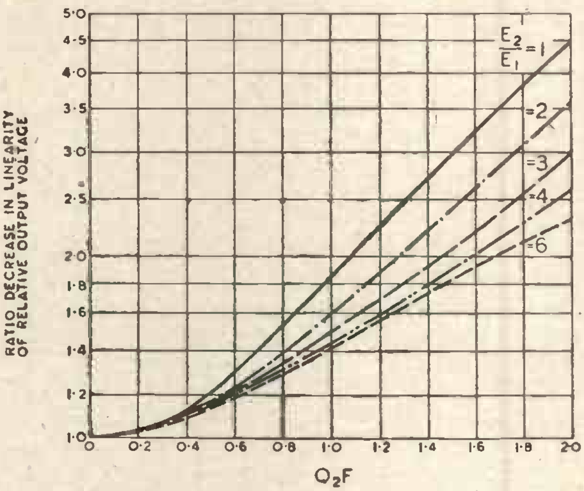

Fig. 6. *Ratio loss of linearity of relative output voltage for a phase discriminator at different off-tune frequencies.*

and it may be noted that the maximum value of conversion efficiency is obtained with E_2/E_1 as large as possible for given values of Q_1, Q_2, and k. A large value of E_2/E_1 calls for a large ratio of L_2/L_1 and in practice it is not feasible to make L_2/L_1 greater than about 2. Optimum design, taking all factors into consideration, would appear to be

$$E_2/E_1 = 2, \quad Qk = 1.5, \quad L_2/L_1 = 1.77$$

3. Application to a Particular Design

To illustrate the design of the frequency-amplitude phase discriminator converter, the following constants are assumed: $f_m = 4.5$ Mc/s, carrier frequency deviation $= \pm 75$ kc/s, $g_m = 2$mA/V, $R_a = 1\text{M}\Omega$, $\eta_d = 0.85$, $E_g = 1$ volt, $R_3 = R_4 = 0.1\ \text{M}\Omega$, $E_2/E_1 = 2$, $Qk = 1.5$, $L_2/L_1 = 1.77$, $QF = 1$ at $\Delta f = 100$ kc/s, or $Q = \frac{f_m}{2\Delta f} = 22.5$, $k = \frac{1.5}{Q} = 0.066$

Maximum conversion efficiency requires the highest possible value of R_{D1}, which means the highest value of L_1; but L_1 is limited by the maximum possible value of L_2, which in turn is decided by the minimum value of tuning capacitance C_2. Assuming the latter to be 50 $\mu\mu$F,

$$L_2 = 25\ \mu\text{H},\quad L_1 = 14.1\ \mu\text{H},\quad C_1 = 88.5\ \mu\mu\text{F},$$
$$M = k\sqrt{L_1 L_2} = 1.252\ \mu\text{H}.$$

This larger value for C_1 than for C_2 is an advantage because the primary has greater stray capacitance, e.g., from R_3 and R_4, etc.

$$R_{D1} = \omega_m L_1 Q = 9000\Omega$$

Conversion efficiency

$$= \frac{2 \times 2 \times 10^{-3} \times 1 \times 9000 \times 0.85 \times 22.5 \times 2}{4.5 \times 10^3 \times 3.25 \times \sqrt{2}}$$

$= 0.0665$ volts per kc/s per 1 volt peak input.

The relative voltage output scale in Fig. 8 can be converted to an actual voltage output for curve 1 by multiplying by 4.75, i.e., using scale A, and the QF scale can be replaced by the frequency scale as shown.

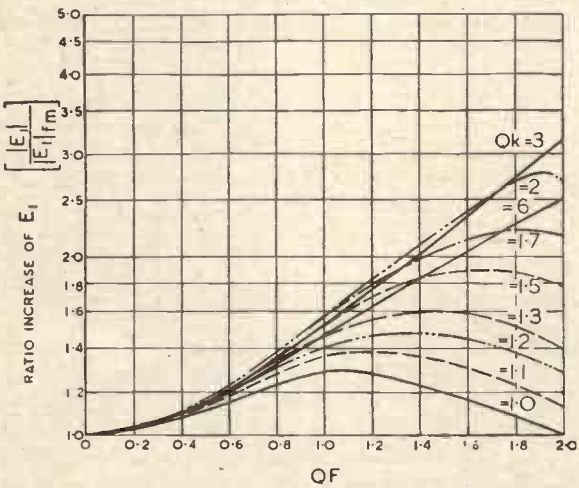

Fig. 7. *Ratio increase of primary voltage E_1 at different off-tune frequencies.*

The damping resistances R'_1 and R'_2 are calculated as follows: let the initial Q_o of the circuits be 150, then the initial primary resonant impedance,

$$R_{D10} = \omega_m L_1 Q_o = 60{,}000\ \Omega,$$

and the total damping resistance required across the primary to make $R_{D1} = 9000\Omega$ is

$$R'_T = \frac{60{,}000 \times 9000}{51{,}000} = 10{,}600\Omega$$

Part of this (see Appendix II) is provided by R_3 and R_4, which are in parallel across the primary, i.e., $50{,}000\Omega$, and the diode conduction damping resistances ($\frac{1}{2}R_3$ and $\frac{1}{2}R_4$) also in parallel making $25{,}000\Omega$. Damping from the slope resistance R_a of V_1 is so small that it can be neglected. Thus the

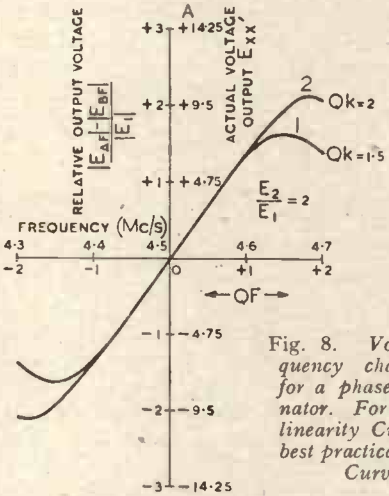

Fig. 8. *Voltage-frequency characteristic for a phase discriminator. For maximum linearity Curve 2 and best practical linearity Curve 1.*

detector circuit provides a damping resistance of $16{,}666\Omega$ and

$$R'_1 = \frac{16{,}666 \times 10{,}600}{6066} = 29{,}200\Omega$$

For the secondary

$$R_{D2} = \omega_m L_2 Q = 15{,}950\Omega$$
$$R_{D20} = 106{,}200\Omega$$
$$R''_T = \frac{106{,}200 \times 15{,}950}{90{,}250} = 18{,}800\Omega$$

The detector damping is equivalent to $\frac{1}{2}R_3$ in series with $\frac{1}{2}R_4$, i.e., $100{,}000\Omega$ so that

$$R'_2 = \frac{100{,}000 \times 18{,}800}{81{,}200} = 23{,}100\Omega$$

4. Adjustment of Converter

Correct tuning and adjustment of the phase discriminator converter is not difficult to carry out when the effect of the various factors on the characteristic is understood. Primary tuning mainly affects the symmetry of the positive and negative halves of the characteristic; a primary resonant frequency less than f_m makes the lower (negative in Fig. 1) peak, B, greater than the higher frequency peak, D. Secondary tuning controls the frequency of zero voltage (O in Fig. 1); a secondary resonant frequency less than f_m reducing the "zero" frequency below f_m. Mutual inductance coupling controls the upper and lower peak frequencies, moving them away from f_m as M is increased. Alignment procedure is therefore best carried out in the following manner. The coupling capacitor C_5 is disconnected and, with mutual inductance coupling less than critical, the primary and secondary are tuned for maximum voltage across either R_3 or R_4 when the input frequency is 4.5 Mc/s. C_5 is now connected and the primary retuned for equal positive and negative maxima across XX' at approximately equal off-tune frequencies on either side of 4.5 Mc/s. The secondary is next adjusted to give zero volts across XX' at f_m. Finally the mutual inductance coupling is increased until equal positive and negative maxima across XX' are obtained at 4.65 and 4.35 Mc/s (this corresponds to $QF = \pm 1.5$). The required linear characteristic should then be obtained.

The effect of variation of mutual inductance coupling (Qk) causes the characteristic to pass through the phases illustrated by curves 1, 2 and 3 in Fig. 9. Curve 1 is obtained when Qk is less than 1, the linear range of characteristic is restricted, and the peaks are close to off-tune frequencies corresponding to $QF = 1$. Curve 2 illustrates the correct value of Qk whilst curve 3 shows how linearity is lost when the optimum value of Qk is exceeded, a double S-shaped characteristic being obtained. As Qk is increased, the positive and negative peaks continue to increase in amplitude.

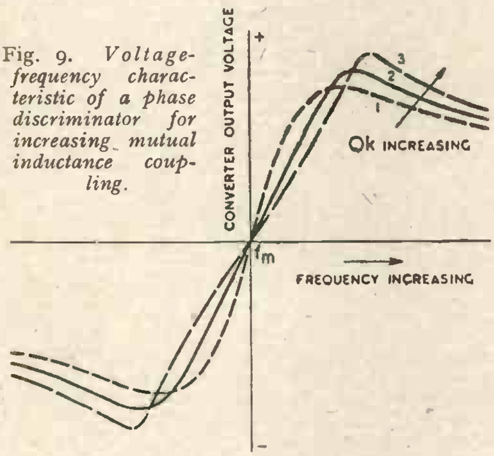

Fig. 9. *Voltage-frequency characteristic of a phase discriminator for increasing mutual inductance coupling.*

5. Conclusions

When the phase discriminator is used as a frequency-amplitude converter, couplings greater than critical ($Qk = 1$) are necessary in order to obtain a linear characteristic over a sufficiently large range of off-tune frequencies. The greatest range is from $QF = 0$ to $\pm 1 \left(\Delta f = 0 \text{ to } \pm 0.5 \frac{f_m}{Q}\right)$ with $Qk = 2$, $E_2/E_1 = 2$ and $L_2/L_1 = 1$, but greater conversion efficiency at a slight sacrifice of linearity (the QF range is from 0 to ± 0.8) is obtained with $Qk = 1.5$, $E_2/E_1 = 2$ and $L_2/L_1 = 1.77$ and these values appear to give the optimum practical design. Linearity of characteristic can be obtained for values of E_2/E_1 exceeding 2 by reducing Qk, but the off-tune frequency range is progressively decreased as E_2/E_1 is increased. For values of E_2/E_1 less than about 2, linearity of characteristic cannot be obtained because the rate of increase of E_1 with increase of off-tune frequency from zero is a maximum at $Qk = 2$.

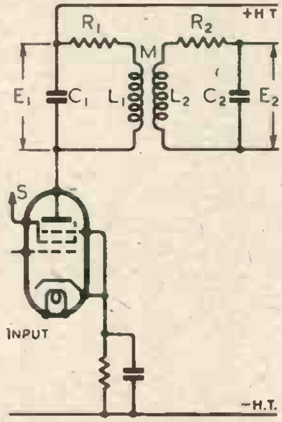

Fig. 10. *Tuned circuits with mutual inductance coupling.*

The author makes grateful acknowledgment to Marconi's Wireless Telegraph Company for permission to publish.

APPENDIX I

Expressions for the Voltages across the Primary and Secondary of Two Tuned Circuits Coupled by Mutual Inductance

The circuit diagram of an R.F. amplifier stage having two tuned circuits coupled by mutual inductance as its anode load is shown in Fig. 10, and its equivalent in Fig. 11. For the latter the valve is represented as a constant current generator of $I = g_m E_g$, and its slope resistance R_a, which normally would be in parallel with the primary tuned circuit, is included in the total series resistance component, R_{T1}, so that $R_{T1} = R_1 + \frac{\omega^2 L_1^2}{R_a}$ where R_1 = series resistance of the coil L_1.

Similarly any damping resistance external to the secondary circuit is included as part of R_{T2}.

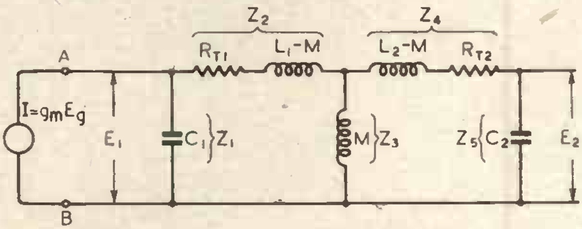

Fig. 11. *The equivalent of Fig. 10.*

(*a*) *The Expression for the Primary Voltage E_1*

The primary voltage $E_1 = g_m E_g Z_{AB}$ where Z_{AB} = impedance looking into the primary at points AB.

$$Z_{AB} = \frac{Z_1\left[Z_2 + \dfrac{Z_3(Z_4 + Z_5)}{Z_3 + Z_4 + Z_5}\right]}{Z_1 + Z_2 + \dfrac{Z_3(Z_4 + Z_5)}{Z_3 + Z_4 + Z_5}}$$

$$= \frac{Z_1[(Z_2 + Z_3)(Z_3 + Z_4 + Z_5) - Z_3^2]}{(Z_1 + Z_2 + Z_3)(Z_3 + Z_4 + Z_5) - Z_3^2}$$

$$Z_1 + Z_2 + Z_3 = \frac{1}{j\omega C_1} + R_{T1} + j\omega L_1$$

$$= R_{T1}\left[1 + j\frac{\omega_m L_1}{R_{T1}}\left(\frac{\omega}{\omega_m} - \frac{\omega_m}{\omega}\right)\right]$$

$$= R_{T1}\left[1 + jQ_1\left(\frac{\omega}{\omega_m} - \frac{\omega_m}{\omega}\right)\right]$$

where $\omega_m = \dfrac{1}{\sqrt{L_1C_1}} = \dfrac{1}{\sqrt{L_2C_2}}$

and $Q_1 = \dfrac{\omega_m L_1}{R_{T1}} = \dfrac{1}{\omega_m C_1 R_{T1}}$

$$\frac{\omega}{\omega_m} - \frac{\omega_m}{\omega} = \frac{(f - f_m)(f + f_m)}{f_m \cdot f}$$

Replacing f by $f_m + \Delta f$, where Δf = the frequency off-tune from f_m.

$$\frac{\omega}{\omega_m} - \frac{\omega_m}{\omega} = \frac{(\Delta f)(2f_m + \Delta f)}{f_m(f_m + \Delta f)}$$

$$\simeq \frac{2\Delta f}{f_m} = F.$$

Hence $Z_1 + Z_2 + Z_3 = R_{T1}(1 + jQ_1F)$

Similarly $Z_3 + Z_4 + Z_5 = R_{T2}(1 + jQ_2F)$

$$Z_3^2 = \omega^2M^2 = \left(\frac{\omega}{\omega_m}\right)^2 R_{T1} \,.\, R_{T2} \,.\, \frac{\omega_m L_1}{R_{T1}} \,.\, \frac{\omega_m L_2}{R_{T2}} \,.\, \frac{M^2}{L_1L_2}$$

$$= \left(\frac{\omega}{\omega_m}\right)^2 R_{T1} \,.\, R_{T2} \,.\, Q_1Q_2\,k^2$$

$$\simeq R_{T1} \,.\, R_{T2} \,.\, Q_1Q_2\,k^2$$

where $k = \dfrac{M}{\sqrt{L_1L_2}}$ = coupling coefficient.

$$Z_2 + Z_3 = R_{T1} + j\omega L_1 = R_{T1}\left(1 + jQ_1\frac{\omega}{\omega_m}\right) \simeq R_{T1}(1 + jQ_1)$$

$$\therefore\ Z_{AB} = \frac{1}{j\omega C_1}\,\frac{(1 + jQ_1)(1 + jQ_2F) + Q_1Q_2\,k^2}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2\,k^2}$$

Since $Q_1 \gg 1$ and $Q_1Q_2k^2 \ll Q_1$

$$Z_{AB} \simeq \frac{Q_1}{\omega C_1}\,\frac{(1 + jQ_2F)}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2\,k^2}$$

$$\simeq R_{D1} \,.\, \frac{(1 + jQ_2F)}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2\,k^2}$$

where

R_{D1} = resonant impedance of the primary

$$= Q_1\omega_m L_1 = \frac{Q_1}{\omega_m C_1}$$

$$\therefore\quad E_1 = g_m E_g R_{D1} \,.\, \frac{(1 + jQ_2F)}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2\,k^2}$$

(b) The Expression for the Secondary Voltage E_2

The secondary voltage
$$E_2 = \frac{E_1 \,.\, \dfrac{Z_3(Z_4 + Z_5)}{Z_3 + Z_4 + Z_5}}{Z_2 + \dfrac{Z_3(Z_4 + Z_5)}{Z_3 + Z_4 + Z_5}} \,.\, \frac{Z_5}{Z_4 + Z_5}$$

$$= g_m E_g \,.\, Z_{AB}\,\frac{Z_3Z_5}{Z_2(Z_3 + Z_4 + Z_5) + Z_3(Z_4 + Z_5)}$$

$$= g_m E_g \,.\, \frac{Z_1Z_3Z_5}{(Z_1 + Z_2 + Z_3)(Z_3 + Z_4 + Z_5) - Z_3^2}$$

$$Z_1Z_3Z_5 = \frac{1}{j\omega C_1} \,.\, \frac{M}{C_2}$$

$$= -j\frac{\omega_m}{\omega} R_{T1} \,.\, R_{T2} \,.\, \frac{1}{\omega_m C_1 R_{T1}} \,.\, \frac{1}{\omega_m C_2 R_{T2}} \,.\, \frac{M}{\sqrt{L_1L_2}} \,.\, \omega_m\sqrt{L_1L_2}$$

$$= -j\frac{\omega_m}{\omega} R_{T1} \,.\, R_{T2}\, Q_1\, Q_2\, k \,.\, \omega_m\sqrt{L_1L_2}$$

$$= -j\frac{\omega_m}{\omega} R_{T1} \,.\, R_{T2} \,.\, Q_1\, \omega_m L_1 \,.\, Q_2\, k\sqrt{L_2/L_1}$$

$$\simeq -j\, R_{T1} \,.\, R_{T2}\, R_{D1} \,.\, Q_2\, k\sqrt{L_2/L_1}$$

$$\therefore\quad E_2 = g_m E_g \,.\, R_{D1} \,.\, \frac{-jQ_2\,k\sqrt{L_2/L_1}}{(1 + jQ_1F)(1 + jQ_2F) + Q_1Q_2\,k^2}$$

APPENDIX II

Conduction Current Damping due to the Diode Detectors

The equivalent circuit is that of Fig. 12. It has been shown* that conduction current damping due to a diode detector approximates to a parallel damping resistance of half the D.C. load resistance when detection efficiency approaches unity. Under normal operating conditions

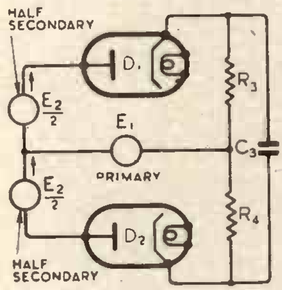

Fig. 12. *The equivalent detector circuit.*

detection efficiency has a value of 0.85 or greater, so we may assume that the damping resistance across one half secondary due to diode D_1 is $\frac{1}{2}R_3$; this is stepped up four times by the centre tap to $2R_3$ across the whole secondary. Similarly that from D_2 is $2R_4$ across the secondary and the total conduction resistance is

$$R_{d2} = \frac{2R_3 \,.\, 2R_4}{2R_3 + 2R_4} = \frac{2R_3R_4}{R_3 + R_4} = R'$$

if $R_3 = R_4 = R'$

The primary conduction current damping resistance is that due to D_1 and D_2 in parallel, i.e.,

$$R_{d1} = \frac{\frac{1}{2}R_3R_4}{R_3 + R_4} = \tfrac{1}{4}R'$$

In the circuit of Fig. 3, R_3 and R_4 are themselves in parallel with the primary, thus providing an additional damping resistance of $\frac{1}{2}R'$. Thus the total damping resistance across the primary is

$$R_{dt1} = \frac{\dfrac{R'}{2} \,.\, \dfrac{R'}{4}}{\dfrac{R'}{2} + \dfrac{R'}{4}} = \frac{R'}{6}$$

* Diode Detection Analysis. C. E. Kilgour and J. M. Glessner. *Proc. I.R.E.*, July 1933, p. 930.

Wireless Patents

A Summary of Recently Accepted Specifications

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

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

555 688.—Arrangement of gas-filled relay for controlling a two-wire low-frequency line.

Standard Telephones and Cables (assignees of J. B. Johnson). Convention date (U.S.A.) 26th March, 1941.

555 721.—Valve relay circuit for connecting a common signal line to any selected one of a number of differently-tuned circuits.

Standard Telephones and Cables (assignees of N. I. Hall). Convention dates (U.S.A.) 21st and 26th November, 1941.

DIRECTIONAL WIRELESS

555 803.—Eliminating the 180° ambiquity in a radio compass comprising a constantly-rotated frame and a cathode-ray indicating tube.

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

555 834.—Means for increasing the effective beam power used to illuminate the cathode-ray indicator of a radio compass having a rotating aerial.

Standard Telephones and Cables (assignees of H. G. Busignies). Convention date (U.S.A.) 6th March, 1941.

RECEIVING CIRCUITS AND APPARATUS

(See also under Television)

555 618.—Superheterodyne type of receiver with phase-detector and local-oscillator control for receiving frequency-modulated signals.

Philco Radio and Television Corpn. (assignees of C. Travis). Convention date (U.S.A.) 27th February, 1941.

555 732.—Construction and arrangement of the magnetic cores in a permeability-tuned circuit capable of covering a wide frequency range.

Marconi's W.T. Co. (assignees of R. L. Harvey and C. Wentworth). Convention date (U.S.A.) 1st March, 1941.

555 771.—Oscillation generator for frequency-modulated signals with a feed-back network serving to stabilize the amplitude of the oscillations.

F. G. Clifford. Application date 28th April, 1942.

555 857.—Mixing valve and associated circuit for converting a frequency-modulated signal into an equivalent amplitude modulated signal, with the aid of local oscillations.

Marconi's W.T. Co. (assignees of S. Hunt). Convention date 15th May, 1941.

TELEVISION CIRCUITS AND APPARATUS

For Transmission and Reception

555 773.—Direct pick-up television system wherein the rate at which successive areas of the mosaic screen are illuminated is greater than the rate at which they are scanned.

Standard Telephones and Cables (assignees of C. E. Huffman). Convention date (U.S.A.) 6th June, 1941.

555 840.—Television system in which the signal currents are deliberately distorted initially for the purpose of increasing the signal-to-noise ratio.

Standard Telephones and Cables (assignees of M. W. Baldwin, Junr.). Convention date (U.S.A.) 1st November, 1940.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

555 619.—Multiplex signalling system utilising intermittent pulses of short duration, each channel being distinguished by different repetition frequencies.

Standard Telephones and Cables; P. K. Chatterjea and L. W. Houghton. Application date 27th February, 1942.

555 863.—Cathode-ray tube in which the electron stream is subjected to velocity modulation in order to generate frequency-modulated dot-dash signals.

The British Thomson-Houston Co. Convention date (U.S.A.), 30th November, 1939.

555 864.—Cathode-ray tube in which the electron stream is velocity modulated in order to utilise the device for the simultaneous reception of signals of different frequencies.

The British Thomson-Houston Co. Convention date (U.S.A.) 30th November, 1939.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

555 656.—Cathode-ray tube in which the modulation control of the electron stream is made independent of the focusing and beam-deflection effects.

Standard Telephones and Cables (assignees of E. Bruce). Convention date (U.S.A.) 14th February, 1941.

555 790.—Cathode-ray tube with means for deflecting the electron stream in order to generate ultra-high frequencies.

Philips Lamps (communicated by N. V. Philips' Gloeilampenfabrieken). Application date 23rd October, 1941.

SUBSIDIARY APPARATUS AND MATERIALS

555 400.—Method of cutting and mounting a piezo-electric crystal so that it will vibrate at a number of independently controllable frequencies.

Standard Telephones and Cables (assignees of W. P. Mason). Convention date (U.S.A.), 25th July, 1941.

555 563.—Means for reducing heat losses in a "thermistor" or element having a non-ohmic temperature coefficient.

Standard Telephones and Cables (assignees of G. L. Pearson). Convention date (U.S.A.), 30th August, 1941.

555 615.—Mounting and arranging selenium dry-contact rectifiers to protect them from dampness.

Standard Telephones and Cables and E. A. Richards. Application date 27th February, 1942.

555 631.—Device for measuring the characteristics of balanced cables designed to carry ultra-high frequencies.

Telegraph Construction and Maintenance Co. and R. Sear. Application date 20th July, 1942.

555 726.—Gas-filled valve relay with means for reducing "arcing," particularly in converting systems.

The British Thomson-Houston Co. Convention date (U.S.A.) 3rd February, 1941.

Abstracts and References

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For the information of new readers it is pointed out that the length of an abstract is not necessarily an indication of the importance attached to the work concerned. An important paper in English, in a journal likely to be readily accessible, may be dealt with by a square-bracketed addition to the title, while a paper of similar importance in German or Russian may be given a long abstract. In addition to these factors of difficulty of language and accessibility, the nature of the work has, of course, a great influence on the useful length of its abstract.

PROPAGATION OF WAVES

393. CORRECTIONS TO "ON THE ATTENUATION OF ELECTROMAGNETIC WAVES IN TUBES" AND TO "CALCULATION OF SKIN-EFFECT BY THE METHOD OF PERTURBATIONS."—S. M. Rytov. (*Journ. of Phys.* [of USSR], No. 3, Vol. 4, 1941, p. 287: in English.) *See* 2860 & 4191 of 1940.

394. RANGES OF 112 Mc/s SIGNALS: DO AURORAS AND SPORADIC-E LAYER AFFECT THEM AS THEY DO 56 Mc/s SIGNALS?—E. P. Tilton. (*QST*, Oct. 1943, Vol. 27, No. 10, p. 34.) A rumour is mentioned that 2½ metre WERS signals in Detroit were heard in New York and confirmed by long-distance telephone.

395. PROPAGATION OF RADIATION IN A SCATTERING AND ABSORBING MEDIUM [Derivation of Equations for Parallel-Ray & Diffuse Radiation: Application to Optics of Atmosphere & Sea].—E. O. Hulburt. (*Journ. Opt. Soc. Am.*, Jan. 1943, Vol. 33, No. 1, pp. 42–45.) For a correction *see* Sept. issue, No. 9, p. 505.

396. SIMPLIFIED FORMULAE FOR SCATTERED AND RE-SCATTERED SUNLIGHT [Hammad & Chapman's Formulae (3456 of 1939) made Easily Applicable by Obvious Reductions & the Discarding of Unimportant Terms].—L. Silberstein. (*Journ. Opt. Soc. Am.*, Sept. 1943, Vol. 33, No. 9, pp. 526–532.)

397. THE EXTENSION OF THE DOPPLER PRINCIPLE TO DISCONTINUOUS PERIODIC EFFECTS [such as Time between Hits of Bullets fired by Machine Gun on Aeroplane in Motion].—L. Fleischmann. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 103–105.)

398. ULTRA-VIOLET RADIATION: VARIATION OVER THE SUNSPOT CYCLE [Hypothesis of the Shielding Effects associated with Atmospheric Ionisation].—J. R. Ashworth. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, p. 354.) A note by "T.W.B." on Ashworth's letter dealt with in 3255 of 1943.

399. "REPORTS ON PROGRESS IN PHYSICS: VOL. 9 (1942/3)" [Book Review].—W. B. Mann (Edited by). (*Proc. Phys. Soc.*, 1st Nov. 1943, Vol. 55, Part 6, No. 312, p. 516.)

List of contents only: they include ten reports by members of the Gassiot Committee of the Royal Society (among them Chapman's "Atmospheric physics and chemistry" and "Photochemistry of atmospheric oxygen", Bamford's "Photochemical processes in an oxygen–nitrogen atmosphere", Sayers's "Ionisation phenomena in the earth's atmosphere", and Cowling's "Absorption of water vapour in the far infra-red") together with ten other reports, including Bruckshaw's "Physics and the search for oil" and Jenkins's "Field emission of electrons". For a review of Vol. 8 *see* 2536 of 1942.

400. A STUDY OF ABSORPTION LINES OF POTASSIUM VAPOUR UNDER VARYING CONDITIONS OF TEMPERATURE AND PRESSURE [in connection with Saha's Hypothesis on the Reversing Layers of Stars: Number of Absorption Lines increase with Increase in Partial Vapour Pressure].—D. K. Bhattacharya & S. P. Sinha. (*Indian Journ. of Phys.*, June 1943, Vol. 17, Part 3, pp. 131–134 and Plates.)

PROPERTIES OF CIRCUITS

401. A PROBLEM IN THE SUMMATION OF SERIES [in connection with Electromagnetic Theory].—L. S. Goddard. (*Proc. Cambridge Phil. Soc.*, Oct. 1943, Vol. 39, Part 3, pp. 200–202.)

The sums in question were

$$U_n(\alpha) = \alpha^2 n^2 \sum_{m=1}^{\infty} \sin^2(\pi m/\alpha)/m(m^2 - \alpha^2 n^2)$$

and $$V_n(\alpha) = \alpha^2 n^2 \sum_{m=1}^{\infty} m \sin^2(\pi m/\alpha)/(m^2 - \alpha^2 n^2)^2,$$

where n is a positive integer and $\alpha > 1$. They have been obtained in some special cases by Hahn (1336 of 1941, in connection with cavity resonators), but not very conveniently. "It is our purpose in this note to obtain expressions in a closed form for $U_n(\alpha)$ and $V_n(\alpha)$ in the case when α is an integer. These are very convenient for purposes of calculation, and approximate values . . . for α not an integer may be obtained by interpolation."

402. Oscillator and Amplifier Systems with a Single Tuned Circuit: the Effect of the Tuned-Circuit Characteristic on the Stability and Amplification Factor Respectively [for Short & Ultra-Short Waves].—G. T. Shitikov. (*Izvestiya Elektroprom. Slab. Toka*, No. 12, 1940, pp. 24–31.)

Consideration is given to facilitating the choice of L and C in the design of a high-frequency amplifier or a Hartley oscillator, using a single tuned circuit. The discussion is confined to operation on short and ultra-short waves. For the case of an h.f. amplifier coupled to a preceding valve stage through a single tuned circuit (Fig. 1), a formula (7) determining the permissible amplification factor k is derived for the condition that for a given frequency and variation in valve inter-electrode capacity, k shall not decrease more than $\sqrt{2}$ times, if both valves are changed. It appears from this formula that k is independent both of Q and of the choice of values of L and C. It is further shown that for a Hartley oscillator the stability is also independent of L and C but is directly related to Q. Methods are therefore indicated, and have been verified experimentally, for determining the optimum value of Q. A number of practical suggestions are added.

403. The Active-Power Losses in Linear Quadripoles, and Their Dependence on the Value of the "Transformed" Impedance.—A. Weissfloch. (*E.N.T.*, Dec. 1942, Vol. 19, No. 12, pp. 259–265.)

This is the paper dealt with in 3287 of 1943. Author's summary:—"If a linear quadripole fulfils the requirements of the so-called 'inversion law,' it is possible to construct, on the basis of three measurements, an equivalent circuit (Fig. 2) which divides it into a loss-component and a wattless component. Also, on the basis of these measurements, a [circle-family] diagram (Fig. 3) can be drawn from which it is easy to obtain for any load the 'relative' active-power loss N_v/N_{ges}, where N_v is the active-power loss in the quadripole itself and N_{ges} is the active power applied to the quadripole." See also 404, below.

404. A Simple Method of determining the Active-Power Losses in Receiving Diodes at Very High Frequencies.—Weissfloch. (*See* 460.)

405. The "Triple Tie-Line" Device and Its Use for the Matching of Bolometer Resistances.—Meinke: Fränz. (In paper dealt with in 524, below.)

406. Transmission-Line Matching Simplified: Easier Methods of determining Stub Dimensions.—Garretson. (*See* 454.)

407. A Generalisation of Foster's Reactance Theorem, extending It to Arbitrary Impedances [with Special Reference to the Matching of Aerials & Feeders].—K. Fränz. (*E.N.T.*, May 1943, Vol. 20, No. 5, pp. 113–115.)

In matching an aerial to a cable the problem is to develop a matching circuit which will transform, without loss, the characteristic impedance of the cable to the conjugate complex of the aerial impedance. For a fixed frequency this problem can, in principle, always be solved, but up to the present the known circuits will give a wide-band matching only to a limited extent: for instance, for an aerial of natural damping d they will allow matching over a frequency band of relative width approximately $(\omega_2 - \omega_1)/(\omega_2 + \omega_1) \approx d$. "This would seem to suggest that theoretical limits must exist for wide-band matching: such a limit will be determined in the following pages. It will be seen that it bears a close relation to the well-known Foster's reactance theorem" [for other work on this theorem *see*, for example, Maa, 3302 of 1943 and back reference]. But Foster's theorem is too restricted to solve the present problem: Cauer's proof (3017 of 1941) that for every circuit composed of coils, condensers, and ohmic resistances there exists an equivalent circuit, with only a single ohmic resistance, which has the same impedance for all frequencies, indicates that what is required is a generalised analogy to the theorem.

"We shall now show that for a passive but otherwise arbitrary circuit the knowledge merely of the effective resistance over a finite frequency range will give a lower limit for the slope of the decrease of the reactance with frequency, which will explain the above-mentioned relation between aerial damping and maximum matching interval. In the limit, for vanishing aerial damping, this interval shrinks to zero according to Foster's theorem." The generalised analogy to that theorem states that "the slope of the reactance cannot become more strongly negative than corresponds to that reference function whose real component vanishes except in the given frequency range." In the case where the real component has a simple characteristic course, this leads to the statement that approximate compensation is possible so long as $(\omega_2 - \omega_1)/\omega_0 < d$, where $\omega_0 = \frac{1}{2}(\omega_1 + \omega_2)$. The above generalisation of Foster's theorem is valid not only for circuits with "lumped" components but also for wave-guides, line sections, etc.

408. The Calculation of the Characteristic Impedances of Symmetrical Feeders.—Tatarinov. (*See* 458.)

409. Matching an Output Valve to a Low-Impedance Line [Advantages of Cathode-Follower Circuit].—S. N. Ray. (*See* 529.)

410. Grid Control, Cathode Control, and Cathode Amplifier [Cathode-Follower Circuit].—W. Kleen. (*E.N.T.*, June 1943, Vol. 20, No. 6, pp. 140–144.)

In addition to the classic type of amplifier circuit in which the input circuit is connected between grid and cathode and the output circuit between anode and cathode ("grid control"), there are two other types of circuit, distinguished by the mode of connection of the external circuits to the three primary electrodes. The fundamental properties of all three types are examined in this paper: the treatment applies equally well whether they are used for transmission, reception, or frequency multiplication.

Fig. 1 shows the classic connection. This is habitually referred to as "grid control": actually, both input and output circuits are connected to the cathode. Figs. 2a & b show the "cathode amplifier" connection (*Anglicè* "cathode follower") in two forms: actually, both input and output circuits are connected to the anode, while in the "cathode control" circuit of Fig. 3 they are connected to the grid. The writer therefore urges the adoption of Steimel's suggestion that the three types of circuit should be designated according to the electrode common to both input and output circuits, so that the "grid control" circuit would be known as the "cathode-basis connection," the "cathode follower" as the "anode-basis connection," and the "cathode-control" circuit as the "grid-basis connection": for he points out that all three types involve "grid control" equally, so that the present designations are really meaningless. Worse, they have already led to misunderstandings, such as the assumption that the "cathode follower" is identical with the really quite different "cathode control." There are, in addition, three other possible types of circuit, obtained by the simple reversal of the input and output : technical applications of such " reversed " systems are imaginable, but they have so far attained no practical importance, and are dealt with no further here.

For each of the three main types, the writer obtains the equivalent quadripole diagram and equations, and on the basis of these discusses their input and output admittances and their amplifications. The characteristic property of the cathode follower, making it so useful for wide-band amplification, is brought out, and the main difference in properties between this circuit and the " cathode-control " circuit is explained. The results are tabulated at the end of the paper, where it is also mentioned that the same relations hold good for multi-grid valves, provided that account is taken of the fact that so far as a.c. processes are concerned every auxiliary grid is equivalent to that electrode of the quadripole with which it is connected in an impedance-free manner.

411. ON THE MODE OF ACTION OF A " CATHODE " AMPLIFIER [Cathode-Follower Stage, for Television, etc.].—K. Müller-Lübeck. (*E.N.T.*, Dec. 1942, Vol. 19, No. 12, pp. 253–259.)

The only references given to previous work are to the Wunderlich paper dealt with in 36 of 1943 and the Költer paper there mentioned. Author's summary :—" For a 'cathode' amplifier with an ohmic cathode-circuit resistance the geometrical construction of the working characteristic is given for arbitrary grid-voltage values, together with the voltage and current equivalent-circuit diagrams for small grid-voltage variations. Further, the curvature of the working characteristic for moderately large grid-voltage variations is discussed. The results are then applied to an example of a compensated d.c. cathode stage [*i.e.* a d.c. stage with compensated steady cathode-voltage, as used by the Fernseh A.G. : Fig. 5], and finally to the so-called 'anode–cathode' stage [for providing one voltage with its sign reversed and another without reversal, or for straightening the characteristic of an " anode " stage : Fig. 7]. If this latter arrangement is taken as an 'anode' stage with a counter-coupling cathode-resistance, the linearising action of the latter on the anode-voltage characteristic can be calculated".

412. CORRESPONDENCE ON " NEGATIVE FEEDBACK : SOME PITFALLS IN APPLYING IT TO QUALITY AMPLIFIERS " [2326 of 1943].—J. T. Terry. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 380–381.)

" Rookie " criticises various points in Terry's paper : the author replies.

413. A NEW TEST METHOD FOR AMPLIFIERS AND COMPONENTS OF COMMUNICATION ENGINEERING.—Knapp & Gerrmann. (*See* 527.)

414. HISTORIC FIRSTS : THERMAL NOISE.—J. B. Johnson. (*Bell Lab. Record*, Oct. 1943, Vol. 22, No. 2, p. 59.)

415. DISTRIBUTION IN TIME OF SPONTANEOUS FLUCTUATION VOLTAGE [due to Thermal Agitation, Shot Effect, and Flicker & Resistance-Fluctuation Effects].—M. Surdin. (*Phil. Mag.*, Oct. 1943, Vol. 34, No. 237, pp. 716–722.)

" During the last two decades extensive experimental and theoretical work was carried out and, chiefly because of the study of the first two phenomena, led to formulae giving the mean-square of the fluctuation voltage or current. However, for the better understanding of these phenomena and for certain practical applications, a knowledge of the distribution in time of the fluctuation voltage is desirable. It is proposed to consider this problem from the thermodynamical point of view, since it has the greatest general application and in the case of voltage fluctuations avoids the use of the somewhat difficult concept of 'random phase.' The derivation given here, of the probability of distribution of a variable describing a closed system, is somewhat different from the one given by Einstein in his general treatment of fluctuations, and follows closely Landau & Lifschitz's.

" Consider a closed system in which x is some physical quantity characteristic of the state of the system. Then the problem is to determine the probability that this quantity has a value x different from the mean value x_0 which it has in the state of equilibrium". Eqn. 14 is found for dp, the probability that x has the value in the given range dx, and is applied to an amplifier which, with its supply voltages, is thermally insulated : then eqn. 17 gives the probability of finding v, the instantaneous output voltage due to fluctuations, at any time in the range dv, while eqns. 18 & 19 give, respectively, the probabilities that the fluctuation voltage will lie between zero and v, or will exceed v, at any given instant. The derivation of eqn. 17 involved no assumption as to the ratio of the inverse of the " time of correlation " to the highest frequency passed by the amplifier, so that this equation is valid, when the spectrum density of v^2 is a function of frequency (in particular in the case of shot effect), for frequencies of the order of, or higher than, the inverse of the time of flight of the electrons : *cf.* Landon (1103 of 1943 and back references) who proved a similar law for fluctuations when the inverse of the " time of correlation " was large compared to the highest frequency passed by the amplifier. " Inversely, it may be proved (Appendix I) that if the probability of finding v in the range dv follows the 'normal law', it is valid for all types of fluctuation, whatever the ratio of the inverse of the 'time of correlation' to the highest frequency passed by the amplifier".

The mean absolute value of v is given by eqn. 20. " An interesting distribution is that of v^2. It can be seen that it does not follow the 'normal law' . . . Eqn. 23 justifies experimental methods using a square-law detector or thermocouple for the measurement of the mean-square value of v . . . The responses of a linear and a square-law detector may now be compared. The probability of distribution of fluctuation voltages, after linear rectification (no smoothing), will remain a 'normal law', whereas after a square-law rectification the distribution will be given by eqn. 22. It will be seen that a linear detector is more advantageous ; the figure of merit, if taken as the mean fluctuation after rectification, is 0·8 for a full-wave linear rectifier and 1 for a full-wave square-law rectifier". Appendix II deals with the distribution of amplitude envelope of fluctuation voltage, often of great interest when narrow-band amplifiers are considered. For other recent work *see* 2420 of 1941 and 1427 & 2090 of 1943.

416. TUNED TRANSFORMERS : DESIGN OF THESE ELECTRONIC UNITS SIMPLIFIED BY MEANS OF UNIVERSAL PERFORMANCE CURVES [Derivation of Formulae for the Curves dealt with in 2091 of 1937].—J. E. Maynard. (*Gen. Elec. Review*, Oct. 1943, Vol. 46, No. 10, pp. 559–561 : to be contd.)

417. TRANSFORMERS FOR TELECOMMUNICATIONS [including Sections on Shunt Loss, Leakage Reactance & Series Loss, Transmission Loss : Core Types : Transformers to carry Polarising D.C., and to pass 17 c/s A.C : Capacity Balance & Electrostatic Screening : Resonance : Modern Core Materials : Testing : etc.].—E. V. D. Glazier. (*P.O. Elec. Eng. Printed Paper No. 176*, pp. 1–14.)

418. TRANSIENTS IN COUPLING CIRCUITS : THE TRANSIENT BASIS FOR THE NECESSITY OF A BAND-WIDTH FOR

PASSING DOTS IS DISCUSSED, *and* OSCILLOGRAMS OF COUPLING-CIRCUIT TRANSIENTS.—G. B. Hoadley & W. A. Lynch. (*Communications*, June 1943, Vol. 23, No. 6, pp. 32–34 and 36, 38 : July, No. 7, pp. 22–24 and 26, 28.)

419. SOME ASPECTS OF CROSSTALK BETWEEN UNBALANCED CIRCUITS [carrying R.F. Currents : with Particular Reference to Methods of Terminating Coaxial Cables used for Bay & Panel Wiring, and to the Influence of Earth Connections].—H. Stanesby & E. W. Ayers. (*P.O. Elec. Eng. Journ.*, April 1943, Vol. 36, Part 1, pp. 14–17.)

420. SIMPLIFICATION OF CABLE CAPACITANCE NETWORKS [Star/Mesh Conversion Formulae obtained by Simple Method avoiding Use of Determinants].—V. R. Pettitt. (*P.O. Elec. Eng. Journ.*, Oct. 1943, Vol. 36, Part 3, pp. 82–84.)

421. "PRINCIPLES OF TRANSMISSION IN TELEPHONY" [Letter calling Attention to a Valuable but Little-Known Book].—M. P. Weinbach. (*QST*, Oct. 1943, Vol. 27, No. 10, p. 62.) Bennett's letter points out that the entire volume is concerned with the derivation and application of the general transmission equations.

422. PROPERTIES OF NEGATIVE INDUCTANCE : CIRCUIT EFFECTS REVEALED BY TRANSITRON AS SOURCE OF NEGATIVE RESISTANCE.—C. Brunetti & J. A. Walschmitt. (*Communications*, June 1943, Vol. 23, No. 6, pp. 14–18 and 84, 86.)

From its definition, a negative inductance will, at any given frequency, present the same type of reactance as a condenser. But if the frequency is increased, the reactance of the condenser will decrease in absolute value, while that of the negative inductance will increase. A reliable and practical negative inductance may find use in transmission-line problems, in filters and corrective networks, and may lead to obtaining impedances proportional to the *n*th power of the frequency, where *n* may be any positive or negative integer. An experimental investigation is described, using the one-valve Transitron (1851 [and 2296] of 1939, and elsewhere) to produce the negative resistance necessary for the circuit.

423. ON THE THEORY OF PERFORMANCE OF WIDE- AND ULTRA-WIDE-BAND LATTICE-TYPE CRYSTAL BAND-PASS FILTERS CONTAINING STABILISED NEGATIVE-IMPEDANCE ELEMENTS.—S. P. Chakravarti. (*Indian Journ. of Phys.*, June 1943, Vol. 17, Part 3, pp. 167–188.)

The simplest (ladder-section) crystal band-pass filter, consisting of crystal and capacitance elements, gives only a small band-width : a lattice section employing crystal and capacitance elements gives a wider band which is still, however, much less than what is often required. A lattice section using crystal, capacitance, and inductance elements (Mason : Stanesby & Broad) gives a still wider band, but even this remains much narrower than is desirable for many modern purposes : also, the use of inductance coils of comparatively low "Q" value gives high attenuation in the pass-band. The present writer and Dutt (1032 of 1941) developed wide- and ultra-wide-band low-loss filters, containing crystal, capacitance, inductance, and stabilised negative-impedance elements, capable of many applications in television and broad-band carrier-current systems, as filters or couplings in multi-channel radio-telephone transmission and receiving arrangements, and suitable (by reducing the band-width) for many other purposes in short- and medium-wave working.

The present paper deals with further theoretical investigations of the Class I and Class II types described in the previous work : the first of these has a series resonant circuit (inductance in series with capacitance) in the series arm, the second has a crystal in parallel with a stabilised negative-impedance element in the series arm, while in both types a crystal connected in series with a stabilised negative-impedance element is in the lattice arm. For each of the two types the various important cases, depending on the relative tunings of the several elements, are considered separately (calculated characteristic curves of Figs. 2 and 5, respectively).

424. THE VIBRATING QUARTZ IN COMMUNICATIONS TECHNIQUE : PART II—CIRCUITS WITH VIBRATING QUARTZ CRYSTALS, AND THEIR APPLICATION IN COMMUNICATIONS TECHNIQUE.—W. Arens. (*E.N.T.*, Dec. 1942, Vol. 19, No. 12, pp. 266–284.)

Part I (von Beckerath) was dealt with in 2973 of 1942. The present writer deals first with some of the most useful generator circuits, paying special attention to the advantages of the series-resonance connection over the simpler and more common parallel-resonance arrangements based on the Pierce (Huth-Kühn) circuit : as examples of circuits for specially high precision he discusses the very useful one-valve arrangement due to Heegner (1837 of 1939) and Meacham's bridge-stabilised oscillator (263 of 1939). He then considers crystal-filter circuits, methods of measuring the electrical constants and of production testing, the application of quartz circuits to wireless and carrier-current communication, and finally their use in measuring technique.

425. INVESTIGATIONS ON AN OSCILLATION GENERATOR WITH RESISTANCE AND CAPACITANCE AS THE FREQUENCY-DETERMINING COMPONENTS : *RC* GENERATOR [Bridge-Stabilised Type].—W. Zaiser. (*E.N.T.*, Nov. 1942, Vol. 19, No. 11, pp. 228–234.)

For other recent investigations *see* 2331/3 & 2670 of 1943 (where Meacham's paper is referred to). Author's summary :—" Oscillators which contain in their frequency-determining portion only resistances and condensers are distinguished, if properly proportioned, by a very high constancy of frequency ; moreover, they show a small harmonic content and a good constancy in time of the output voltage, a very slight variation of this output voltage with change of frequency, and a linear dependence of frequency on the variation of the frequency-determining condensers and resistances". These points are investigated theoretically and illustrated experimentally by measurements on an oscillator with the wide frequency-range of 4 c/s to 1.3 Mc/s, based on the theoretical results. The thermal element used appears to have taken the form of four " 6 v/40 mA " barretter tubes in series (Fig. 4).

426. ON THE THEORY OF THE MAINS-RECTIFIER CIRCUITS.—H. Holzwarth. (*E.N.T.*, Nov. 1942, Vol. 19, No. 11, pp. 218–227.)

Author's summary :—" The mains-rectifier circuit of ' charging-condenser ' type [*i.e.* with the rectifier acting directly on a condenser of large capacitance which delivers the d.c. to the load] is investigated mathematically for an infinite capacitance in the smoothing condenser ; a method is given by which, taking the non-linear rectifier characteristic into account, the transformer voltage for a desired d.c. voltage and a desired d.c. current can be determined quickly and accurately. Formulae are also given for the efficiency and for the losses, both of the whole circuit and of the component due to the rectifier-valve itself " [knowing the d.c. voltage and current and the transformer voltage, the total losses can be derived easily from eqn. 43a and Fig. 11 ; then from eqn. 44 and Fig. 10 the input series resistance loss-component is found, and finally the losses in the rectifier-valve are obtained either by subtraction or by the approximate formula of eqn. 50].

"The cathode-choke circuit [choke lying between rectifier and condenser: this arrangement has certain advantages but reduces the voltage obtainable] is compared with the first circuit. Since its efficiency is higher, a larger current can be taken from the rectifier: the basic calculations for this are given. As an example it is shown that a particular rectifier for which a maximum d.c. current of 100 mA is laid down can yield, in the cathode-choke connection, an increased current of 150 mA with the same anode loss."

427. Electromagnetic Screens with Joints and Gaps.—H. Kaden. (*E.N.T.*, July 1943, Vol. 20, No. 7, pp. 159–169.)

The screening actions (ratio of external to internal field strengths) of the three basic forms of screen, the hollow sphere, the hollow cylinder, and the parallel plates, all of the same wall-thickness and of equal diameter or spacing (Fig. 1), bear to one another the ratios only of 1 : 2 : 3. This makes it comprehensible that the usual calculations of practical screens, by assuming them to be of one of these three forms, are fairly reliable. But the calculations also assume that the screen wall is homogeneous. In practice this is not generally true, for screening cans are often made from sheet metal, so that they contain joints, cable and feeder screens are made of spirally wound tape, transformer covers are built up from two halves. It is obvious that the field near such joints and gaps will be very different from what would be given by a completely homogeneous screen. "It is the object of this paper to investigate the effect of such joints and gaps," for the case of screens of non-magnetic material.

The treatment is analytical, and the results are illustrated by diagrams. In practice the ideal is to arrange things so that the unavoidable joints or gaps lie parallel to the eddy-current paths and perpendicular to the primary (external) field, but this cannot always be accomplished; for instance, a spirally-wound screen is bound to have its joints at an angle to the eddy currents. The treatment considers the two extreme cases, where the joints are at right angles to the eddy currents and where they are parallel to them. In the first case the bending of the lines at the joints causes the magnetic field to be completely distorted. This distortion decreases with increasing frequency and diminishes exponentially with the distance from the surface of the screen: at a distance equal to the distance between the joints it has practically vanished. The field in the screened space is therefore periodic, with a period equal to the spacing of the joints. For this reason the length of twist of a two-wire line, running parallel to a screen, should not coincide with the spacing between joints (Fig. 8).

In the second case, where the joints are parallel to the eddy currents, the latter are no longer deflected, and if the gap at the joint is negligible the screen acts perfectly. Section 3 therefore neglects the eddy currents and devotes itself to calculating how the external field "wells" through a gap (at right angles to it) in an infinitely extended plane screen. The method of conformal representation is used, and the starting point is the Laplace differential equation (eqn. 15) with appropriate boundary conditions. It is found that the field distortion decreases with decreasing air gap. The field strength in the internal space is proportional to the square of the gap width and to the square of the reciprocal of the distance from the gap. The gap acts on the internal space like a magnetic linear dipole. For a distance from the gap of (say) 10 times the gap width, the field strength is only 0.625‰ of that of the original interfering field.

With the help of the formulae obtained it is possible to calculate the "coupling resistance" (eqn. 53) of a screen with a gap to an asymmetric conductor inside the screen. The coupling resistance of a coaxial cable with a gap in the outer conductor is also found. "Previously there was no picture available showing the spatial distribution of the field in a screened space in the neighbourhood of such joints or gaps. This is now provided by the diagrams of Figs. 7, 13 and 14. They provide a basis for the correct choice of the arrangement of the screen itself and of the parts to be screened, especially the spacing of the latter from inhomogeneous points in the screen."

428. Polarised-Iron-Core Chokes for Regulating Purposes [including a Comparison between Series & Parallel Connections].—W. Schilling. (*Elektrot. u. Masch:bau*, 29th Aug. 1941, Vol. 59, No. 35/36, pp. 397–406.)

429. A Universal Resonance Chart [for Electrical & Mechanical Systems].—H. G. Yates. (*Engineer*, 1st Oct. 1943, Vol. 176, No. 4577, pp. 268–269.)

"A fact whose significance, so far as the writer can discover, has not been commented upon [in the literature of the subject] is that Z may be written as $r/\cos\phi$ instead of being given in the more cumbersome form of $Z^2 = r^2 + (2\pi fm - s/2\pi f)^2$. With this simplification the response, in terms of the response at resonance, may be written simply as $u/u_0 = \cos\phi \, . \cos(2\pi ft - \phi)$. Thus, as soon as the phase shift ϕ has been calculated from the simple expression $\tan\phi = Q \, . \, (f/f_0 - f_0/f)$, the response may be obtained from a table of cosines. Furthermore, the information given in these last two equations may be conveniently represented by the simple nomogram or alignment diagram shown in the attached figure," which has "several advantages over the universal resonance curves generally used (*e.g.* Terman, *Radio Communication*, 2nd Ed., p. 56)." These advantages are shown, and some points of interest regarding mechanical damping are discussed.

430. On Some Phenomena of Parametric Excitation in Physics and Technics [and the Advantages of "Parametric" Machines].—N. D. Papalexi. (*Journ. of Phys.* [of USSR], No. 4, Vol. 4, 1941, p. 378: summary only, in English.)

The researches of Mandelstam & Papalexi and their colleagues on parametric excitation, stretching over the last fifteen years, have been dealt with in various past abstracts: *see* for example 1780 of 1940. In the present summary it is mentioned that in such machines, where the parametric excitation is carried out by periodic variation of the capacitance or self-induction, accidental currents are always initially present in the form of thermal-fluctuation currents: there is therefore no need for any kind of exciting winding, the construction is greatly simplified, a better use is made of the iron than in ordinary machines, the winding-free rotor can have a very high speed, and the machine can be given a number of characteristics convenient for special practical purposes and otherwise obtainable only by supplementary devices and with difficulty.

TRANSMISSION

431. On the First Model Type HB 14 of a "Resotank" [Cavity-Resonator Valve for a Fixed Wavelength of 14 cm: Constructional Details].—W. Dällenbach. (*Hochf:tech. u. Elek:akus.*, June 1943, Vol. 61, No. 6, pp. 161–163.)

This generator, with the elements of a specially designed small-electrode-spacing retarding-field valve built into a metallic cavity resonator, was exhibited in 1937 and outlined only in the paper dealt with in 2258 of 1938: for theoretical papers *see* 2389 of 1941 and back references. Since such a "cavity valve" represented a complete novelty in the design of valves, and since later years have indicated the fundamental importance of such design in decimetric- and centimetric-wave technique, the writer thinks that full constructional details will be of interest, "particularly as the design unites in itself a number of advantages which are absent in certain schemes at present being worked on."

After giving, with the help of a photograph of a sectional model, a full account of the construction, the writer lists at the end a number of practical points contributing to the attainment of an extremely low-damping resonator such as is necessary if high efficiency is to be obtained. Among these points are (a) u.h.f. surfaces (of copper) are finished with a diamond tool: (b) the coaxial section "11" (Fig. 1), which functions chiefly as an inductance, is made in the intermediate form shown, between the extreme forms of Fig. 2 a and c (narrow tube and flat pot) because this design, with its outer radius and axial length about equal, gives the smallest loss in current and yet has a sufficiently good frequency-stability, though not as good as the extreme forms [*cf.* 29 of January]: (c) the spaces between the cathode and the grid, and between the end-plate "5" and the metal cap "25," are each made as a cavity detuned from the 14 cm working wave: the former in order to prevent the cathode and grid from oscillating together, the latter to avoid u.h.f. energy leaking along the filament-current and d.c.-voltage leads.

Other final points dealt with are the choice of the radius of the retarding-field electrode (there is an optimum relation between this and the grid radius) and the matching of the generator to the load (by the correct selection of the annular gap between "4" and "6"). In conclusion, a special advantage of the whole design is that the generator can be frequency-modulated at its maximum power without consumption of energy, and with a slope of 90 kc/s per volt.

432. Modification of the Peterson "Pot" Oscillator to give Parallel Plate Feed and avoid D.C. Voltage on the "Pot."—F. D. Lewis: Peterson. (*QST*, Oct. 1943, Vol. 27, No. 10, p. 59.) *See* 4366 of 1939. For the correct illustration *see* November issue, p. 53.

433. Ultra-High-Frequency Design Factors [Discussion of Factors such as Contact Resistance, Circuit Stability, Tuning Methods, Circuit "Q," Valve Admittance, Shielding, & Coupling: with Illustrations].—A. H. Meyerson. (*Communications*, June 1943, Vol. 23, No. 6, pp. 20–22 and 24..26, 70.) From the Radio Laboratory of the New York Fire Department.

434. Sideband and "Swinging-Vector" Theory of Frequency Modulation for Sinusoidal and Rectangular Modulation [and for Small & Large Deviations].—O. Zinke. (*E.N.T.*, April 1943, Vol. 20, No. 4, pp. 93–102.)

Starting from the equation $\phi = \omega_0 t + \phi_w$ and the Helmholtz definition ("frequency corresponds to the speed of phase change": $\omega = d\phi/dt$), the writer shows the differences between frequency and phase modulation: in frequency modulation $f - f_0$ is proportional to u_m (the l.f. modulating voltage) and the accompanying phase fluctuation ϕ_w is proportional to $\int_0^t u_m dt$, whereas in phase modulation ϕ_w is proportional to u_m, and the accompanying frequency fluctuation to du_m/dt. Another criterion is that if, for modulation with a constant-amplitude pure sinusoidal voltage u_m of variable modulation frequency f_m, an increase of f_m produces no change in the frequency deviation, it is a case of frequency modulation: the phase deviation decreases with $1/f_m$: whereas with phase modulation the phase deviation is not affected by the variation of f_m, while the frequency deviation increases in proportion to f_m. Of the two, frequency modulation has the greater technical importance: "Hölzler has shown that the greatest reduction of interference occurs only with frequency modulation in the narrower sense."

For frequency modulation the maximum phase fluctuation $\Delta\phi$ is given, with sinusoidal modulation, by eqn. 5 ($\Delta\phi = \Delta f/f_m$), while for rectangular modulation it is given by eqn. 5a ($\Delta\phi = 1.57 . \Delta f/f_m$). The frequency deviation referred to the modulation frequency, the ratio $\Delta f/f_m$, is of vital importance as regards the reduction of interference, and is termed the "modulation index": for sinusoidal modulation it is, by eqn. 5, identical with the phase deviation, whereas with rectangular modulation the phase deviation equals 1.57 times the "modulation index," by eqn. 5a.

After this, the writer introduces the use of swinging-vector diagrams to represent the processes of frequency modulation. "In Fig. 3 the time axis rotates, as usual, clockwise with an angular velocity ω_0. The frequency modulation makes the vector arm, which in the absence of modulation is stationary, swing in the rhythm of the modulation period T_m, to a maximum angle, to right and left of its central position, equal to the phase deviation $\Delta\phi$. The velocity with which the vector arm swings to and fro corresponds to the course in time of $d\phi_w/dt$, that is, to the frequency change $f - f_0$ within the modulation period." After a discussion of Fig. 3 (for a phase deviation $\Delta\phi = 2$: left, rectangular frequency modulation; right, sinusoidal) the writer shows how the frequency spectrum can be obtained (Fig. 4) by splitting the swinging vector of constant length, of the previous diagram, into two stationary vectors of constant frequency ω_0 whose amplitude is modulated. "This procedure is not limited to small phase deviations but can be applied also to large deviations and also to arbitrary, non-sinusoidal modulation. Fig. 4 shows that the vector of constant amplitude A swinging with the phase angle ϕ_w is resolved into $A_t'' = A \cos\phi_w$ in the direction of the central position and $A_t' = A \sin\phi_w$ at right angles to this position. Both vectors A_t' and A_t'' retain their position relative to the time axis and to each other and do not change their angular position even when A swings to and fro with a considerable deviation ($\Delta\phi \gg 1$). This means that A_t' and A_t'', as two amplitude-modulated oscillations of the frequency ω_0, together reproduce the frequency-modulated oscillations. As soon as A_t' and A_t'' have been subjected to Fourier analysis, the frequency spectrum is obtained. This is particularly easy to do with a small deviation" [section VI: Fig. 4 shows the spectrum for sinusoidal modulation, Fig. 5 for rectangular. Both differ from the usual representation of spectra (by vertical lines of varying height distributed along a horizontal axis) which gives only an incomplete picture owing to the neglect of the phase situation: in the present spectra the side-frequencies are represented in their phase relations to the starting-time point, $t = 0$. Without this representation of the phase relations, a comparison of the spectrum of Fig. 4 with that of amplitude modulation is liable to lead to the conclusion that for a small deviation ($\Delta\phi < 1$) the two spectra are identical. There are however two important differences: the f.m. spectrum contains further side-frequencies, $f_0 \pm 2f_m$ being specially prominent: and the side-frequencies $f_0 \pm f_m$ are at 90° to the carrier, in contrast to the phase relation in amplitude modulation].

Frequency spectra for large frequency deviations are then considered by the same procedure. "Particularly illuminating relations between the position of the strongest sidebands in rectangular modulation and the frequency deviation" are obtained (pp. 98–99). The conditions for the disappearance of the carrier frequency are derived. Next, the "theory of the keying spectrum" is developed (Fig. 9), starting from the assumption that a rectangularly frequency-modulated oscillation is made up of two alternately sharply keyed oscillations of the highest and lowest frequencies: the frequency spectrum for rectangular frequency modulation is obtained by the superposition of two known keying spectra, and following on this the band-widths occupied by rectangular and sinusoidal modulations are found (Fig. 12). Finally, the notable phenomenon of the formation, in frequency modulation, of a spectrum asymmetrical with respect to f_0 (a thing which cannot happen in amplitude modulation however irregular the shape of the modulation curve), is explained, and the condition necessary for the production of a symmetrical spectrum, which Vellat (678 of 1942) calls "radial symmetry," is discussed with the help of Fig. 13.

435. Measurement of the Characteristic Values of Frequency-Modulated Oscillations.—W. Stäblein. (*E.N.T.*, April 1943, Vol. 20, No. 4, pp. 102–111.)

Recent researches have tended to confine themselves to the theory of frequency modulation, more particularly to the occurrence of distortions, and comparatively little has been done with regard to methods of measuring the characteristic values of a frequency-modulated wave. Occasionally the frequency spectrum has been plotted (Cudell, 2989 of 1942), but this requires an extremely selective filter and the method becomes hardly workable when the modulation index [*see* Zinke, 434, above] is large. "In the present paper a measuring method is described which is basically applicable to all conditions and especially when the modulation index is large, and which allows all the characteristic values of a frequency-modulated oscillation to be determined simultaneously. It has the advantage that it requires only a small expenditure in apparatus. Some of the values cannot, it is true, be obtained *directly* from the measurements, but must be determined subsequently from an oscillogram obtained photographically.

"In the method to be described a second oscillation of known constant frequency and equal amplitude is superposed on the frequency-modulated oscillation, the mixture is (if necessary) amplified to a suitable value, and the resulting beat image observed on an oscillograph. The following quantities are thus obtained: (1) The value of the modulation index or phase deviation, and hence the frequency deviation; (2) the modulation curve contained in the frequency modulation: that is, the determination whether the modulation follows a sine curve or any other periodic curve; (3) the magnitude, form, and phase relation (referred to the phase of the frequency modulation) of any simultaneously occurring amplitude modulation; and (4) the exact value of the carrier frequency."

For an explanation of the processes involved the writer calls upon "swinging vector" theory and its diagrams: the reference given to the development of this method of representing modulation is to Zinke's book (3311 of 1938), but *see* also 434, above. In such a diagram (Fig. 2) it can readily be seen what relations obtain when the frequency-modulated oscillation has superposed upon it an oscillation of constant amplitude and frequency. If this frequency is equal to the carrier frequency f_0, the rotating vector representing the heterodyning oscillation remains fixed at a definite angle χ with respect to the axis of symmetry. Such a vector is shown in Fig. 2 with its end point represented by Q'. If both oscillations are present simultaneously, at any instant the sum of the two vectors, the rotating vector MQ' and the swinging vector MP, must be obtained. This is accomplished on the diagram more simply if instead of the vector MQ' the reversed vector MQ is used: the resultant oscillation is then given by the vector QP, making an angle ψ_p with the line of the central position, compared with the original variable phase-angle ϕ_p made by the swinging vector MP. The angle ψ_p is obviously also periodically variable with the modulation frequency, but it cannot behave as a simple sine function and displays distortions which increase as the length of MQ increases.

It is pointed out, in passing, that from such a swinging-vector diagram the distortion of a frequency-modulated oscillation heterodyned by a constant oscillation can be arrived at: *see* Wundt & Hoffmann, 1872 of 1943. Fig. 2 also shows that the amplitude of the resultant oscillation QP fluctuates during a swing of the point P within the period T_m of the modulation frequency. It has its minimum value, obviously, when P lies exactly on the prolongation MQ, and its maximum when P lies on the prolongation MQ'. It is on this property that the measurement of the characteristic values of the f.m. oscillation is based. To simplify the procedure the amplitude of the superposed constant-frequency oscillation is made equal to that of the f.m. oscillation under test: in Fig. 2 this means that the point Q, as well as the point P, lies on the circle, and the amplitude of the heterodyned oscillation becomes zero each time that P, rotating along the circle, coincides with Q.

Section C deals with the theoretical side of the method, considering first the formation and significance of the beat image (Fig. 3), then the variation of this image according to the value of the modulation index (Fig. 4), then its variation with the value of χ, the angle between the heterodyning vector MQ' and the horizontal axis of symmetry (Fig. 5, for a constant phase deviation $\Delta\phi = 20$ and for χ ranging from 0° to 315° in 45° steps), its variation with the wave-form of the modulation (Fig. 6), and finally its variation with the frequency of the heterodyning oscillation: Fig. 8, where the numbers 0–11 represent the beat image given when the heterodyning frequency coincides with the f.m. carrier, with its first sideband frequency, and so on. Only in the first case is the image symmetrical, the asymmetry increasing as the order of the sideband increases. It is mentioned that this asymmetry for a varying heterodyning frequency may be utilised to determine in which half-period of the modulation period the frequency is (for example) raised with respect to the carrier frequency: it is only necessary to shift the heterodyning frequency upwards and to observe in which half-period the number of beat antinodes is diminished.

Coming to the practical carrying-out of the method, the writer points out that it is often difficult to obtain stationary images, owing to the impossibility of holding the heterodyning frequency sufficiently accurately on the prescribed value. But the required results may be obtained if the image can be kept steady enough to allow a photograph to be taken. Using photography, either the heterodyning frequency may be kept as constant as possible, yielding a series of pictures such as those of Fig. 5, or it may be altered slowly during the taking of the photographs, as in Fig. 9, in which the point where the heterodyning frequency coincides with the carrier frequency is clearly identifiable by the symmetry condition just discussed: the point is indicated by J_0, while J_1 identifies coincidence with the first sideband. In this record the modulation voltage is also shown; this should always be done, to help in working out results, though it is not actually essential.

The last subsections deal with the application of the method to the case where simultaneous amplitude modulation is present, and with the measurement of the exact value of the carrier frequency and of any fluctuation it may undergo during modulation.

436. The Application of Negative Feedback to Radio Transmitters using Grid Modulation.—S. V. Person & V. A. Khatskelevich. (*Izvestiya Elektroprom. Slab. Toka*, No. 12, 1940, pp. 11–23.)

The possibility is discussed of introducing negative feedback into grid-modulated transmitters already in operation in Russia. Various types of feedback (over the modulator channel, over the modulated high-frequency channel, and over both the high-frequency and low-frequency channels) are considered. Phase and phase-frequency distortions thus introduced are examined, and measures to reduce these are suggested. The possibility of self-excitation of the transmitter is discussed in detail, and in conclusion a brief report is given on an experimental investigation in which feedback over the high-frequency and low-frequency channels was employed.

437. A Valve Circuit for the Keying or Modulation of Oscillators with High Grid Current [as in Pulse Transmitters: the Advantages of a Cathode-Follower Connection].—F. Below. (*Hochf:tech. u. Elek:akus.*, June 1943, Vol. 61, No. 6, pp. 164–167.)

Although a pulse transmitter can be keyed at a much higher power than that corresponding to the rating of its transmitting valves, the occurrence of large grid currents in these valves puts a big strain on the keying stage. For instance, a push-pull transmitter with two 150 w valves, when keyed for pulses of 10^{-5} s, repeated (say) every 10^{-4} s, can be made to give 2 kw without overloading the valves, but grid currents of 1 A or more may easily occur. The writer describes a keying circuit which can provide such currents without the use of a disproportionately large keying valve. Such a stage must also have a small internal resistance, or else the grid current will produce a large voltage drop which will reduce the d.c. voltage at the grids of the transmitting valves. For the same reason the keying impulse must not be led through a condenser, and a high leak resistance is also detrimental.

All these points are best satisfied by the cathode-follower circuit, which has also the advantage that aging effects in its valve have little or no influence on the transmitter adjustment, and even a changing of its valve produces a change in the transmitting-valve bias only if the amplification factor of the new valve is different—and this can easily be counteracted by an adjustment of the voltage of the battery B_1 (Fig. 3).

Calculation of the amplification of the cathode-follower stage leads to eqn. 4, $V = 1/(1 + 1/\mu + 1/SR_k)$, so that for an amplification as large as possible (as near unity as possible) the cathode resistance R_k must be large and a high-mu valve selected. From this point of view, however, the use of a pentode has no advantage here, since the cathode voltage fluctuates, and with it the screen-grid voltage with respect to the cathode, so that the advantage of the screen grid is illusory in this case. Nevertheless the use of a small transmitting pentode is recommended, because this generally combines a small screen-grid penetration-coefficient with a steep slope. Thus a type RS 288 with its screen grid connected to the anode has a $1/\mu$ of 0.05, which with an infinitely large cathode resistance should give an amplification of 0.95. Actually a resistance of 5000 ohms gives the highest attainable value of about that figure (Fig. 2).

438. Oscillator and Amplifier Systems with a Single Tuned Circuit: the Effect of the Tuned-Circuit Characteristic on the Stability and Amplification Factor respectively.—Shitikov. (*See* 402.)

439. The Vibrating Quartz in Communications Technique: Part II.—Arens. (*See* 424.)

RECEPTION

440. Ultra-High-Frequency Design Factors.—Meyerson. (*See* 433.)

441. A Receiving Method with Magnetic-Field Valves worked at Electron-Rotation Resonance [Habann Principle].—H. Jungfer. (*Hochf:tech. u. Elek:akus.*, June 1943, Vol. 61, No. 6, pp. 172–189.)

In contrast to the many papers on the magnetron as a transmitter, hardly anything has been written about it as a receiver: references are made to three papers only, those of Habann (1356 of 1938), Wolff & others (986 of 1935), and Schmersow (1874 of 1941). The ordinary way of using the magnetron for receiving is to utilise a bend in its characteristic: for high frequencies the magnetic field is so adjusted that the frequency of electron-rotation agrees with the signal frequency, in order to obtain a damping reduction in the circuit. In all such methods the optimum adjustments are when $\sqrt{U_a}/B$ remains constant (U_a is the anode d.c. voltage, B the magnetic induction), when the shape of the characteristic keeps unaltered to a first approximation.

Habann, however, refers to an entirely different method (*loc. cit.*), in which no anode d.c. voltage, or at most a very low one, is employed: detection occurs not through a bend in the anode-current characteristic but through a type of resonance between the spiralling-electron rotation and the incoming oscillations. It is this method, termed the "rotation-resonance" method, that the present writer investigates both theoretically and experimentally. It is found that the rotation resonance occurs at the calculated point, and that it is practically independent of the dimensions of the anode system, of the value of the anode alternating voltage, of the inclination of the magnetic field, and of the value of the filament-heating voltage. An increase in anode d.c. voltage necessitates a slight raising of the magnetic induction.

Both theory and experiment show that the method, when applied to a "straight" receiver, is unsuitable for the reception of weak signals owing to its possession of a marked threshold of sensitivity. On the other hand, it can be applied successfully to the mixing stage of a superheterodyne receiver for frequencies above about 100 Mc/s. As regards its "sensitivity" when thus applied, it is about equal on wavelengths around 1 m to the best reception method for such wavelengths, namely superheterodyne reception with pentode mixing, and begins to be superior to this at the upper end of the decimetric-wave band. This superiority holds also as regards the diode mixing usually employed in the lower part of this band, and probably also (data are lacking) as regards mixing by retarding-field valves and by magnetrons used in the ordinary way. An objection to the method is that the optimum receiving properties are only attained for a quite definite value of the heterodyning voltage: but since this may be very low, it can easily be adjusted to the best value by the loose coupling which is all that is necessary. A further advantage of the method is that the special mechanism of rotation-resonance provides a considerable degree of selectivity even without any external tuning system.

The above-mentioned superiority over other methods of mixing is based on the extremely good properties of the system in the matter of valve noise. The highest obtainable "sensitivity" on a 1 m wave is around 35 kT_0. This may be compared with 50 kT_0 for a small pentode: a push-pull pentode (type EFF 50) will also give 35 kT_0

at 1 m, but as the frequency is increased further the sensitivity deteriorates rapidly (p. 188). A magnetron used in the ordinary way will give a good sensitivity, but the adjustments are extremely critical for satisfactory results and cannot be maintained for long: to obtain more stable conditions it is customary to modulate a supply voltage (usually that of the anode) with a medium-frequency voltage: this, however, increases the valve noise and decreases the sensitivity.

The sensitivity of 35 kT_0 at 1 m "may also be expected at shorter wavelengths if the anode diameter is correspondingly reduced": for a discussion of this point *see* pp. 187–188. The writer ends by pointing out that a similar resonance between electron motion and signal frequency can occur in a reciprocating motion in a d.c. electric field: a receiving system on this idea has been mentioned by Hollmann in his book (784 of 1937), but nothing definite about it has been published.

442. CURVES OF CONSTANT CONVERSION AMPLIFICATION IN RECTIFYING-CHARACTERISTIC CURVE FAMILIES.—H. F. Mataré. (*E.N.T.*, June 1943, Vol. 20, No. 6, pp. 144–148.)

Following on the paper dealt with in 2705 of 1943. Author's summary:—"It is shown that by means of isogonal trajectories [by "trajectory" is meant a line which cuts the various curves of a curve family according to a definite law] in the characteristic field of a valve, curves of constant parallel resistance are obtained from the external and internal resistances of the valve ["constant parallel resistance" $R_p = R_{ext} \cdot R_{int} / (R_{ext} + R_{int})$; eqn. 9], from which it is possible to obtain the curves of constant conversion amplification, even when there is no predetermined relation between the external and internal resistances. The method can be applied to any characteristic field [in the above, by "characteristic field" the writer signifies the single-parameter family of rectifying characteristics of the form $F(x,y,c) = 0$: for diodes the single parameter c is generally the alternating-voltage amplitude, while x and y represent the rectified current and the d.c. voltage: for triodes the parameter c is the grid voltage]. A calculation of the equation for the trajectories is possible in the region for which the current-flux angle $\theta = 90°$, and is carried through" (eqn. 30): but the graphical construction is much simpler.

443. CONTRIBUTION TO THE CALCULATION OF THE SIGNAL-VOLTAGE/NOISE-VOLTAGE RATIO AT THE OUTPUT TERMINALS OF RECEIVERS: CORRECTION AND SUPPLEMENT TO A PREVIOUS PAPER [3026 of 1941].—K. Fränz. (*E.N.T.*, Dec. 1942, Vol. 19, No. 12, pp. 285–287.)

"The object of the original work was, among other things, to calculate the transmission of noise voltages over smooth characteristics. Let such a characteristic, giving the relation between input voltage v and output voltage v_a, be represented by $v_a = \Sigma a_p \cdot v^p/p!$ The noise generated by the powers $p \geq 3$ was given too small by mistake; corrections are therefore necessary to the sections III b, d, e and V. After showing the point at which the previous calculations were incomplete [failure to take into account some non-negligible terms in eqn. 49 for $f_p\,(n) f_p^*\,(n)$], we shall derive the correct results by another, clearer method which has been developed in the meantime (3027 of 1941). The results given in the summary of the original paper, so far as they effect receiving technique (A_1, A_2, A_3), are not invalidated by the error": but compared with the earlier calculations, for $p \geq 3$ new terms with $\mu \neq 0$ appear. This means that the behaviour of the rectifier with higher powers than $p = 2$ is much more complicated than was assumed: the energy spectra are no longer identical with the Fourier spectra of the writer's "normal" interference, but are made up of several "normal" spectra: and there is no simple analogy to the incorrect eqn. 60. Finally, the writer points out that the subject has attained an interest above and beyond considerations of the "sensitivity" of receivers, since Bennett's proposal to investigate the non-linearity of multi-channel systems with the help of noise voltages (1333 of 1941).

444. CROSS MODULATION AND INPUT NOISE VOLTAGE.—E. Hudec. (*E.N.T.*, May 1943, Vol. 20, No. 5, pp. 123–135.)

"An ideal receiver has the task of amplifying the oscillations supplied by the aerial in the desired frequency band, even when these are extremely weak, and of suppressing all oscillations outside that band which the aerial may supply, however strong these may be. The fulfilment of the first task is limited by the input noise voltage of the receiver [this is sometimes spoken of as the "sensitivity" of the receiver: *cf.* Fränz, 3126 of 1939 (and 443, above). Such a terminology is considered undesirable, however, since what is generally meant as "sensitivity" is inversely proportional to the input noise voltage: for instance, if that voltage is at a minimum, the receiver is at its most sensitive, *i.e.* its sensitivity is at its maximum]. The fulfilment of the second is limited by the imperfections of the filter and by the interfering oscillations produced by cross modulation in the desired frequency band."

The two requirements are mutually contradictory; for the input noise voltage of a receiver has a minimum for a particular coupling between the aerial and the input circuit, but at this value of coupling the input circuit is considerably damped by the aerial, so that to keep down the cross modulation through the input valve it would be desirable to weaken the coupling by an appreciable amount. The resulting loss of sensitivity, compared with the gain in selectivity for the input circuit, is less serious than it might be, owing to the minimum of input noise voltage, as a function of the coupling, being a fairly flat one.

The cross modulation can, theoretically, be eliminated without increasing the input noise voltage by seeing that the valve characteristic $i_a = f(u_g)$ has a square-law form, so that the higher derivatives $f^{\text{III}}(U_g), f^{\text{V}}(U_g)$ . . . vanish: *see* eqns. 5 & 11. But Fig. 5, giving the cross-modulation-factor curves of three typical modern valves, from measurements made at the optimum adjustment of the cathode-lead resistance (*i.e.* of the grid bias, giving minimum cross modulation), shows that up to the present commercial valves deviate widely from this form of characteristic. A comparison between types of valve shows that their cross modulation is the smaller, the higher their noise voltage, so that a reduction of cross modulation by the selection of a suitable type can only be accomplished at the cost of higher input noise voltage. On p. 130 the writer shows how the cross modulation of a given valve can be reduced, without increasing its input noise voltage, by counter-modulation with the help of the l.f. current resulting from rectification. Eqn. 12 yields an expression (eqn. 30) for a component of the anode current which is a pure l.f. current containing only the modulation frequency and higher harmonics, and which ordinarily is completely suppressed by the filter in the anode circuit. This current can be used to modulate the desired oscillation in opposition to the cross modulation, by an arrangement of cathode-circuit resistance and small bridging-condenser as shown in Fig. 10a or b. The excellent result, for an EF14 valve, is seen in Fig. 11.

Unlike negative feedback (which also decreases cross modulation) counter-modulation exerts no adverse effect on the sensitivity, nor does it decrease the valve amplification. An often useful combination of counter-modulation and negative feedback is given by the circuit of Fig. 12. The results are shown in Fig. 13: for all three types there is a great improvement over Fig. 5, but especially for the EF11. For a mixing valve the same relations hold fundamentally as for an amplifier, if the static characteristic is replaced by the Rothe & Kleen "conversion characteristic." Then the relations already obtained for the cross-modulation factor hold good, but only for small amplitudes of the interfering voltage: for larger values, cross modulation is also produced by the fact that the conversion slope does not vary linearly with the amplitude of the input voltage. Fig. 14 shows how the cross-modulation factor of an ECH11 mixing valve is reduced by counter-modulation. If the condenser C_1 in Fig. 10 is removed, negative feedback comes in through the resistance R_{k1} and reduces the cross modulation still further: but the amplification is diminished. In a superheterodyne receiver highly selective filters are connected in front of the mixing valve, so that cross modulation through the latter only occurs if the interfering voltage has a frequency close to that of the desired signal. But another effect shows itself here, through which the modulation of the interfering frequency is transferred to the desired frequency band: owing to the curvature of the mixing-valve characteristic the interfering voltage produces an anode current containing terms of the form of eqn. 37, and this current combines with the oscillator oscillations to produce a beat frequency which, if the condition of eqn. 38 is fulfilled, passes through the i.f. filter. This effect also can be balanced out with counter-modulation, at any rate under certain conditions (Fig. 15).

After a general discussion in section 10 of the considerations determining the choice of amplification in a h.f. amplifier, section 11 takes as an example the superheterodyne circuit of Fig. 18, consisting of two h.f. amplifier stages (EF14 and EF11), an oscillator (EF14), and a mixing stage (ECH11). The cathode resistance of the input EF14 is so adjusted that cross modulation is compensated by counter-modulation as completely as possible, as in curve "b" of Fig. 11. The EF14 is provided with counter-modulation and also negative feedback, and the mixing valve with counter-modulation only (curve "a" of Fig. 15). The input noise voltage of the receiver is 0.7 μv or 8 μv (according to whether the aerial-coupling switch is in the "sensitive" or "insensitive" position) for a current source of resistance 70 ohms. Fig. 20 shows the effective value of a 100% modulated interfering voltage required to produce a cross modulation of 1, 5, and 10% of the 100% modulated signal. The circuit of Fig. 18 is followed by an i.f. amplifier with quartz bridge filters (*cf.* 2097 of 1943) in its first three stages, so that there is no cross modulation beyond the mixing valve.

445. THE TRACKING OF THE HIGH-FREQUENCY FILTERS OF SUPERHETERODYNE RECEIVERS.—E. Hudec. (*E.N.T.*, Nov. 1942, Vol. 19, No. 11, pp. 235–240.)

Author's summary :—"Formulae are derived for the calculation of the tracking of the rotating condensers of superheterodyne receivers [the particular receiver considered (Fig. 1) has two h.f. amplifier stages with EF14 and EF11 valves respectively, an oscillator with an EF14, and a mixing stage with an ECH11. The grid circuit of the input valve contains an oscillatory circuit, the anode circuits of the two amplifier valves each contains a band filter composed of two capacitively coupled oscillatory circuits; the condensers of all these circuits are driven by a common shaft, which also controls the oscillator-tuning condenser]. The calculation is carried out in particular for a circuit in which the influence of the self-capacitance of the coils, connections, and valves are taken into account. . . It is shown by an example that a 100% synchronisation is attainable with band filters. The calculated results are confirmed by measurements."

446. SOME "HETROFIL" SNAGS AND THEIR SOLUTION.—F. Dearlove. (*QST*, Oct. 1943, Vol. 27, No. 10, pp. 59–60.) For this device for eliminating heterodyne interference *see* Woodward, 4392 of 1939.

447. RADIO NOISE IN SMALL AIRCRAFT [and the Problems presented in Its Suppression].—D. K. Kinsey. (*Communications*, May 1943, Vol. 23, No. 5, pp. 34 and 36, 71, 73.)

448. THE SUPPRESSION OF RADIO INTERFERENCE FROM INTERNAL-COMBUSTION ENGINES [in Road Vehicles (from Private Cars to Tanks) where Complete Screening of Ignition System is Not Necessary: the Advantages of Henley H.R.I. Ignition Cables with Conducting Rubber as the Conductor, giving 25 000 Ohms in an 18-Inch Length: Methods of Terminating: Results: the Need for Post-War Legislation].—H. A. Macdonald. (*Distribution of Electricity*, Oct. 1943, Vol. 16, No. 152, pp. 179–182.)

449. CURRENT AND POTENTIAL DISTRIBUTION IN "SHORTED-EDGE" ROLL-TYPE CONDENSERS [of Special Importance in Interference-Suppression].—L. Leiterer. (*E.N.T.*, July 1943, Vol. 20, No. 7, pp. 170–182.)

This type of construction, in which the many edges of each of the two spiral windings are soldered together or otherwise connected on the outside, and the leads taken to the two common surfaces thus obtained, provides an inductance-free condenser which is of great importance in the high-quality suppression of interference. This is particularly true at high frequencies, for in complete contrast to other condensers this type behaves better and better as the frequency of the interference increases.

There is, however, one exception to this rule: all condensers of the type in question have a certain limited (and generally not very high) frequency zone in which they display multi-resonance effects. These effects are difficult to account for on ordinary lines, for they cannot be attributed to series or parallel resonance of the condenser capacitance combined with an inductance, nor to standing waves of the ordinary kind, since the wavelengths of the resonance frequencies obtained bear no relation to the external dimensions of the roll winding, and may in fact exceed these by many orders of magnitude: thus with a 1 μF condenser resonances occur at wavelengths as long as 30 m.

These new phenomena were first reported on by W. Thormann & R. Zechnall in the 1940 *Jahrbuch der deutschen Luftfahrtforschung*, together with another, intimately related phenomenon: that the "apparent resistance" (voltage/current ratio) of such condensers falls abnormally rapidly as very high frequencies are reached, namely much more steeply than in inverse proportion to the frequency. These writers derived formulae which correctly gave a partial explanation of the occurrence of resonances. But a satisfactory elucidation of the phenomena can only be obtained when a clear idea has been formed of the processes in the roll winding, the current and potential distributions in its interior. The object of the present paper is to contribute to this knowledge.

The writer assumes that the penetration depth of the high frequency in the metal of the foil is large compared with the foil thickness, and he represents the "shorted-edge" condenser by a system of coaxial interleaved condensers of cylindrical type. "For the current distribution in such a cylindrically symmetrical system a group of differential equations (eqns. 11 a–e) is obtained; a characteristic of these is that they are linked together by terms involving the resistance of the foil. The solution of the differential equations leads, since the special boundary conditions (set by the fact of the edges being shorted) must be observed, to a transcendental determinant-condition equation [eqn. 27] having an infinite series of roots. One of these roots has the special property that the current distribution corresponding to it represents the desired current distribution in the condenser [*see* p. 176]. All the remaining particular solutions together have only the significance of a slight edge-correction in the neighbourhood of the outermost turn, and can be neglected.

"From the current distribution thus found, the potential distribution is also obtained, and from the two together the impedance and 'core' resistance of the condenser are derived [by "core" resistance (a term taken from network theory) is meant the ratio of U_a to the condenser input current $I(-1)$, where U_a is the output voltage on open circuit: *see* p. 180, r-h column. For a small number of turns and at low frequencies, the "core" resistance and the impedance coincide, but at high frequencies this is not the case, for then U_a may differ from U_e (the voltage in front of the condenser) by several orders of magnitude. The "core" resistance is of great importance in interference suppression]. The final formulae for the impedance and 'core' resistance correctly give the experimentally observed resonances (eqn. 62 for the frequency of the first peak). The sharpness of the resonances is a question of the foil resistance, the losses in the dielectric, and certain other factors which have not yet been dealt with by the calculations." Comparison (p. 182) of eqns. 60 and 61, for the impedance and "core" resistance respectively, with the corresponding equations for a coaxial line open at the end (eqns. 66 and 67) brings out the relations between the two arrangements.

450. Reception Reports [Comprehensive Code for defining the Quality of a Received Signal].—Internat. Broadcasting Union. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, p. 377.) For the B.B.C. signal-strength code for oversea listeners *see* 2379 of 1943.

451. Basic Fault-Finding: Methods of Training Army Service-men [at Loughborough College].—E. Wilkinson. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 359–361.) Latest developments of 3795 of 1942.

452. Musical Taste [and the True Jazz Fan's Preference].—"Diallist": Fellgett. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, p. 383.) Reply to Fellgett's letter, 93 of January.

453. Further Notes on the Contrast Expansion Unit: Simplifying the Control Circuit: Hints on Operation, *and* Contrast Expansion [Note on Correspondence].—Williamson: Hughes. (*See* 480 & 481.)

AERIALS AND AERIAL SYSTEMS

454. Transmission-Line Matching Simplified: Easier Methods of determining Stub Dimensions.—T. A. Garretson. (*QST*, Oct. 1943, Vol. 27, No. 10, pp. 42–49.) The method of computing is a simplified version of that given by Potter for Rutgers University classes. Both open and shorted stubs are considered. For rapid work Quaranta's quadrant-&-pointer "calculator" is used (*see* Fig. 11).

455. A Generalisation of Foster's Reactance Theorem [with Special Reference to Aerial/Feeder Matching].—Fränz. (*See* 407.)

456. Correspondence on "Medium- and Long-Wave Aerial System for a Radio Laboratory" [3492 of 1943: the Cathode Follower suggested for Matching an Output Valve to a Low-Impedance Line].—Ray: Rymer. (*See* 529.)

457. Calculation of the Current Drop and Radiated Power of Aerials, for Damped Progressive Waves: Part I—Current Drop and Radiated Power of a Parallel-Wire Line.—W. Jacknow. (*E.N.T.*, May 1943, Vol. 20, No. 5, pp. 115–123.)

For previous work *see* 585 of 1940, 2407 of 1941, 1050 of 1942, and 417 of 1943. Author's summary:—"The calculation of the radiated power of lines with damped progressive waves is based on the calculation of the damping factor at the end of the line, which gives the drop in current-amplitude up to the end. For this damping factor an integro-differential equation, to which it must conform, can be derived from the energy equation which must apply to every element of length of the line. In the case of straight, parallel, linear conductors of equal length and equal current-amplitude—in particular, a feeder line—this integro-differential equation can be solved, and an ordinary homogeneous differential equation of the first order obtained from it, with non-constant coefficients for the damping factor. The latter can thus still be represented as an *e* function, whose exponent however is no longer linearly dependent on the radiation length . . . but varies in a complex manner with the length and spacing [eqn. 50, p. 119]. The damping exponent is inversely proportional to the characteristic impedance. It is made up of the sum of two components, of which the first represents the radiation resistance of the system for *undamped* progressive waves and the second represents the total ohmic resistance of the line [eqn. 72, from eqns. 62 & 71]. In the special case where the radiation component vanishes, the known equation for the damping factor for pure line damping is obtained.

"It is to be noted that the calculations yield only the damping factor at the end of the whole line, spoken of here as the current drop. This is, however, sufficient for the calculation of the radiated power [eqn. 76]. The calculation of the damping factor for every point along the line is in preparation. This is accomplished by dropping the idealising assumption that the conductors are infinitely thin."

458. The Calculation of the Characteristic Impedances of Symmetrical Feeders.—V. V. Tatarinov. (*Izvestiya Elektroprom. Slab. Toka*, No. 12, 1940, pp. 1–10.)

A symmetrical feeder consists of equal numbers of symmetrically disposed go and return conductors. Formulae are derived for determining the characteristic impedance in the following cases: (1) Feeder consisting of N similar conductors of which n are go and n return; formula 15, which in the case of a two-conductor feeder is reduced to formula 16; more exact formulae (25 & 26) are also derived for the case when the diameter d of the conductor is large in comparison with the diameter D of the feeder. (2) A multi-conductor feeder (Fig. 7) with a circular screen; formula 34; a more exact formula (35) for the case of a twin feeder is also derived. (3) A feeder consisting of two similar conductors each with a circular screen (Fig. 10); formula 39. (4) A formula (40) is quoted for a twin feeder with a common rectangular screen (Fig. 2). (5) A feeder consisting of two conductors, each with a rectangular screen (Fig. 12); formula 41.

459. The Radiation Field of Long Wires, with Application to Vee Antennas.—C. W. Harrison, Jr. (*Journ. Applied Phys.*, Oct. 1943, Vol. 14, No. 10, pp. 537–544.)

"The analysis previously made for the current distribution along a symmetrical centre-driven antenna of non-vanishing radius, and for the radiation field thereof, is extended to include long-wire centre-driven antennas [the references are to the paper dealt with in 423 of 1943, the papers accepted for *Proc. I.R.E.* and referred to in 2721 of 1943, and "The receiving antenna" and "The receiving antenna in a plane-polarised field of arbitrary orientation," both as yet unpublished. The writer's paper dealt with in 3385 of 1943 is referred to later]. The results of this investigation are then applied to obtain an approximate solution for the field of a long-wire resonant vee antenna."

VALVES AND THERMIONICS

460. A Simple Method of determining the Active-Power Losses in Receiving Diodes at Very High Frequencies.—A. Weissfloch. (*E.N.T.*, April 1943, Vol. 20, No. 4, pp. 89–92.)

"In a previous paper (403, above) a proposition was established showing how, on the basis of impedance measurements, the ratio of the active-power loss N_v in the quadripole to the total active power supplied, N_{ges}, (a ratio called for short the 'relative active-power loss') could be determined. This process is as follows:—To the output terminals of the quadripole under investigation are connected in turn at least three reactive impedances, for example ∞ (open circuit), 0 (short circuit), and an arbitrary capacitance, and the corresponding input impedances of the quadripole are determined experimentally. The three measured impedances are plotted in the complex number plane, and through them is drawn a circle, the boundary circle of the quadripole (Fig. 2). This lies wholly in the right half of the complex number plane and allows an equivalent diagram to be set up for the quadripole, dividing it into a loss-component and a wattless component (Fig. 3). $R_s + jY$ is given by Fig. 2 as that border-point of the circle which is nearest to the imaginary axis. R_p is also obtained from the circle, since the latter's furthest point from the imaginary axis has the value $R_s + R_p + jY$.

"As was shown before (*loc. cit.*) it follows that inside the boundary circle it is possible to construct circles of constant relative active-power loss N_v/N_{ges}. These circles are characterised by the geometrical properties (i) that their centres all lie on the line through jY parallel to the real axis, and (ii) that they cut this parallel line at two points $R_1 + jY$ and $R_2 + jY$ such that R_1R_2 is always equal to $R_s(R_s + R_p)$. For all terminating impedances of the quadripole whose transformed input impedances lie on one and the same circle of the family of Fig. 2, the ratio N_v/N_{ges} is a constant. But for the special input impedance $R + jY$ it is easy to calculate, from the equivalent circuit of Fig. 3, that $N_v/N_{ges} = R_s/R + (R - 2R_s)/R_p + R_s^2/RR_p)$ . . . (eqn. 1). It is obvious that the losses are the greater, the nearer the transformed input impedance of the terminating impedance in question is to the edge of the boundary circle, and that on the circle itself they amount to 100%. Eqn. 1 also shows that in general the losses are larger, the smaller the ratio R_p/R_s.

"Now a diode may be regarded, from a high-frequency standpoint, as a quadripole in which the electron space is the quadripole output: the quadripole itself is made up of the valve base, socket, h.f. screening of the d.c. leads, etc., while the quadripole input is represented by the uniform line along which the high frequency is led to the diode (Fig. 4). On this uniform line the input impedance of the quadripole can be determined directly from the potential distribution.

"The relative active-power losses thus measured contain, according to circumstances, any losses due to the valve mounting or to inadequate h.f. screening, so that the quality of all these is included in the result."

The procedure is as follows:—the input impedance is first measured with the diode functioning in the anticipated conditions, with the suitable external resistance and the h.f. power led to it over the test line: the value of this power can be estimated from the value of the rectified current. The input impedance thus obtained corresponds to the loading of the quadripole which is of particular interest as regards the relative active-power loss: it is set down as a point on the complex number plane. It is then necessary to determine the course of the boundary circle: to do this, pure reactive impedances must be applied to the electron space and the corresponding input impedances measured on the test line. The simplest terminating impedance is an open circuit: this may be obtained merely by breaking the d.c. anode circuit, when the anode will automatically become negative and suppress any energy-carrying flow of electrons: or the anode may be provided externally with a strong negative bias to make certain of this suppression—no difference was found between the two methods. One point on the boundary has thus been determined, but to find further points the diode must be opened, when the anode/cathode gap can be altered, a dielectric introduced into the gap, or the anode and cathode short-circuited. The corresponding input-impedance values of the quadripole then provide the whole boundary circle, and from it the relative active-power loss can be determined. For very exact purposes, errors arising from changes in the cathode surface due to the opening-up of the diode must be allowed for: no details are given.

In many cases, however, and especially with centimetric waves where the losses are larger, it is possible to determine the relative loss approximately by a simplification of the procedure not involving any opening-up of the diode. If it is desired to test a series of diodes of the same type, one single specimen may be sacrificed to produce the complete boundary circle, and the rest compared by application of the same circle to input-impedance measurements with each diode under its anticipated working conditions and on open circuit, as described above. On the other hand, it is possible to estimate approximately the course of the circle without opening-up any diode, by measuring the input impedance successively with the anode d.c. circuit broken and with various values of resistance in that circuit; thus in the example given (Fig. 6) the approximate formula of eqn. 4 leads to an estimated loss of 60% of the total power when the resistance is 20 ohms, 62% when it is 10 kilohms, and 75% when it is 100 kilohms: this increase of loss with increasing negative bias of the diode is of importance in selecting the best working conditions.

Comparative measurements on the above simplified lines, including as they do the various losses in bases, sockets, etc., often lead to the discovery of avoidable sources of inefficiency.

461. Curves of Constant Conversion Amplification in Rectifying-Characteristic Curve Families.—Mataré. (*See* 442.)

462. Distribution in Time of Spontaneous Fluctuation Voltage.—Surdin. (*See* 415.)

463. Thermionic Emission from an Oxide-Coated Cathode.—H. Y. Fan. (*Journ. Applied Phys.*, Oct. 1943, Vol. 14, No. 10, pp. 552–560.)

From Kunming, China. The subject has been studied extensively, but "many results are, however, contradictory and the properties of such emitters are not yet thoroughly clarified. We present in this paper the results of experiments made on a cathode coated with barium oxide. Emissions in retarding and accelerating fields [the latter with voltages up to 1300 v] were studied, and the variations of emission constants with the state of the cathode were investigated. The experiments were made at low temperatures, so that the state of the cathode was not affected by the measurement itself" [such work at low temperatures was made possible by the thermocouple technique employed to measure the cathode temperature].

464. THE THEORY OF SECONDARY ELECTRON EMISSION FROM DIELECTRICS AND SEMICONDUCTORS.—A. E. Kadyschewitsch. (*Journ. of Phys.* [of USSR], No. 4, Vol. 4, 1941, pp. 341–348: in German.) The Russian original was dealt with in 2423 of 1941.

465. ELECTRONICS: ITS START FROM THE "EDISON EFFECT" SIXTY YEARS AGO [with Some Newly Uncovered Historical Material].—W. C. White. (*Gen. Elec. Review*, Oct. 1943, Vol. 46, No. 10, pp. 537–541.)

DIRECTIONAL WIRELESS

466. NAVIGATION AIDS IN AIRCRAFT COMMUNICATIONS [Adcock System: Marker-Transmitter: Cone of Silence: Fan Markers: Approach Marker: Simultaneous Range Stations: Aerials on Aircraft: Future Developments].—R. G. Peters. (*Communications*, June 1943, Vol. 23, No. 6, pp. 50..54 and 87.)

467. AN ANALYSIS OF GONIOMETERS FOR RADIO BEACONS.—A. N. Plemyannikov. (*Izvestiya Elektroprom. Slab. Toka*, No. 12, 1940, pp. 44–51.)

On the basis of an investigation by Foelsch (2758 of 1936) formulae are derived for calculating the magnetic field of a cylindrical coil, and various methods for increasing the uniformity of the field along the axis of the coil are surveyed. Two types of radio beacon are then considered: one in which the polar diagram is rotated (Fig. 3) and the other, the American type TR.400, which operates on the principle of a fixed equi-signal zone (Fig. 5). The operation of the two types is discussed and various factors affecting their accuracy are analysed. A number of experimental curves are plotted. Some practical suggestions are also made to increase the accuracy of the beacons.

468. ENEMY AIRBORNE RADIO EQUIPMENT.—C. P. Edwards. (*Electrician*, 26th Nov. 1943, Vol. 131, No. 3417, p. 533: *Elec. Review*, 3rd Dec. 1943, Vol. 133, No. 3445, pp. 753–754: summaries of I.E.E. paper.)

ACOUSTICS AND AUDIO-FREQUENCIES

469. NEW HEADSET PERMITS SIGNAL MEN TO WEAR HELMETS [New Type, coming well down over Sides of Head: Miniature Receivers with Neoprene Inserts].—(*Bell Lab. Record*, Aug. 1943, Vol. 21, No. 12, p. 459.)

470. THE MEASUREMENT OF EXTREMELY SMALL SOUND PRESSURES WITH THE CONDENSER MICROPHONE.—G. Weymann. (*E.N.T.*, June 1943, Vol. 20, No. 6, pp. 149–158.)

Author's summary:—"The condenser microphone of modern design is treated mathematically and an expression derived for the sensitivity: this can be put into a convenient form [eqns. 37, 38] for the type in which the diaphragm mass predominates [$m_M \gg m_s$] or for the type [eqns. 39, 40] in which the mass of the medium moving with the diaphragm predominates [$m_M \ll m_s$. "Of these four expressions only eqn. 38 is generally applicable to the usual designs, since C_n (the minimum attainable circuit capacitance) has to be small compared with the microphone capacitance C, and the foil thickness (for duralumin, for instance) can hardly be below 5μ for technological reasons. By eqn. 38 the sensitivity increases in inverse proportion to the thickness, so that the inclination would be to try much thinner diaphrams, by the use of collodion films for example. That, however, would bring the type at once into the second group . . . where eqn. 39 holds good, according to which the sensitivity increases with the thickness. This confirms the conclusion drawn already from Fig. 7, that the most sensitive microphone is to be found in the transition zone between the two groups, where the thickness lies between 1μ and 0.1μ"].

"It is also found that the molecular noise of the air can be measured experimentally with the highly sensitive types, and that it delivers at the input about the same voltage as is given by the noise sources of the circuit itself, so that a further increase in sensitivity is no longer of decisive importance. The measured noise sound-pressure agrees satisfactorily with the calculated values." The measurements were made with the help of the high-frequency microphone-circuit shown in Fig. 8, in which the h.f. voltage was taken off a quartz oscillator: the voltage thus obtained had a much lower noise content than that of an ordinary self-excited oscillator.

471. OUR EDITORIAL FACE IS RED [Correction to "Historic Firsts: the Condenser Microphone" (3039 of 1943), where the Wrong Picture was shown].—(*Bell Lab. Record*, Aug. 1943, Vol. 21, No. 12, p. 463.)

472. DATA ON NEW ACOUSTIC STETHOSCOPE [3434 of 1943: Method of obtaining Frequency/Response Characteristics: Curves for the New Instrument and Diaphragm-Type & Bell-Type Stethoscopes].—H. F. Olson. (*Electronics*, Aug. 1943, Vol. 16, No. 8, pp. 184 and 186.)

473. IS DISC RECORDING OBSOLETE? [*Wireless World* Brains Trust].—Stuart Black. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 378–379.)

"Fundamentally and mechanically, the present methods are worse than those of the original Edison phonograph . . . The miracle is that [tolerable reproduction] has been attained, to a large extent by sheer misplaced ingenuity." But "it is easy to see where the advantage of the disc lies, and to some extent must probably always lie": however, "there is hardly one of the advantages which the old cylinder possessed that could not be incorporated in that disc." As for the endless other possibilities by other methods, "all I hope is that we shall soon cease to get our music by scraping a steel point carrying some tons of weight per square inch over what is virtually a refined macadamised roadway."

474. TOWARDS BETTER SOUND IN THE THEATRE AND THE HOME [Survey: Fantasound, Vitasound, & Stereo-Control Sound: Development of Stereophonic Sound in Australia, and General Analysis of the Problem of More Faithful Sound Reproduction & Increased Illusion].—E. L. Walther. (*Proc. Inst. Rad. Eng. Australia*, May 1942, Vol. 4, No. 8, pp. 126–133.)

475. CINEMA SOUND QUALITY: INVESTIGATING THE CAUSES OF GOOD AND BAD REPRODUCTION [in Different Cinemas, but with Identical Reproducers: Path or Paths taken in reaching Listener of Much

Greater Importance than Reverberation Times or Frequency Characteristics: Analysis of Architectural Designs by Impulse Tests].—J. Moir. (*Wireless World*, Nov. & Dec. 1943, Vol. 49, Nos. 11 & 12, pp. 320–323 & 362–366.) From the British Thomson-Houston Research Department. *See* also 3071 of 1941.

476. Loudspeaker Installation for Large Surfaces or Spaces [Central Unit for Low Frequencies, with Large Output: Distributed Directive Loudspeakers for Medium & High Frequencies, Each supplied with Suitable Delay Device, so that Echo-Free Reproduction is obtained].—R. Thomson. (*Hochf:tech. u. Elek:akus.*, April 1943, Vol. 61, No. 4, p. 127.) A Telefunken patent, D.R.P. 726 439, applied for 6/9/40.

477. Discussion on "Public Address Systems" [1133 & 1705 of 1943].—S. Hill. (*Journ. I.E.E.*, Part III, Sept. 1943, Vol. 90, No. 11, pp. 151–152.)

478. All-Electronic Sound Reproducer [Suggested Arrangement without Moving Parts: Zig-Zag Track on Stationary Film is scanned by Cathode-Ray Spot].—F. M. Parry. (*Electronics*, Aug. 1943, Vol. 16, No. 8, pp. 280..282.)

479. Audio-Perspective System for Home Radio Receivers.—Volpe. (*See* 94 of January.)

480. Further Notes on the Contrast Expansion Unit: Simplifying the Control Circuit: Hints on Operation.—D. T. N. Williamson. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 375–377.) Based on correspondence prompted by 3048 of 1943.

481. Contrast Expansion [and the Two Essentially Different Standpoints in the Vigorous Correspondence dealt with in 3429 of 1943].—J. R. Hughes. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, p. 382.)

482. New Bass-Boosting Circuit [Simplified Circuit resulting from Correspondence on the Two-Valve Screen-Grid Injection Circuit, 3819 of 1940].—L. M. Barcus. (*Electronics*, June 1943, Vol. 16, No. 6, pp. 216..222.)

483. The Two Most Important "Tone-Diaphragm" Circuits [Simple Low-Pass Filters for A.F. Amplifiers].—de Gruyter. (*See* 90 of January.)

484. A Limiting Amplifier [for Programme Control] with Peak Control Action.—J. K. Hilliard. (*Communications*, May 1943, Vol. 23, No. 5, pp. 13–16.) In which "the principal defects of previous amplifiers [excessive intermodulation or amplitude distortion, and bad transient characteristics] have been greatly reduced." For broadcasting or recording.

485. Simplifying the [*Wireless World* Push-Pull] "Quality Amplifier": Wartime Modifications to a Well-known Design.—(*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 355–358.)

486. Universal Equaliser provides A.F. Amplifier Design Data ["Tonalizer" Unit, inserted in Audio-Amplifier System and adjusted to satisfy Listeners, provides Response Curve which yields Equaliser-Circuit Constants to improve the Response Characteristics].—P. H. Thomsen. (*Electronics*, Aug. 1943, Vol. 16, No. 8, pp. 120–121 and 269..271.)

487. Transformers for Telecommunications.—Glazier. (*See* 417.)

488. Investigations on an Oscillation Generator with Resistance and Capacitance as the Frequency-Determining Components: *RC* Generator [Bridge-Stabilised Type].—Zaiser. (*See* 425.)

489. An Improvised Oscillator for Pip-Tone Supply [Generation of 900 c/s Supply by "Oscillating Amplifier" Circuit, comprising Line Repeater, Equaliser, & Pair of Rectifiers].—N. W. Lewis. (*P.O. Elec. Eng. Journ.*, Oct. 1943, Vol. 36, Part 3, p. 90.)

490. High-Speed Sound-Effect Signal Device [for Introducing a News Period: gives Effect of High-Speed Radiotelegraph Station].—H. E. Adams. (*Communications*, June 1943, Vol. 23, No. 6, pp. 28 and 68.) Simple and convenient: the use of records has several objections, including short life for the records.

491. Industrial Music and Morale: *also* Music as a Safety Factor: *and* Attitudes towards Types of Industrial Music [with Graphs, for Types from Hill-Billy to Classical, for Various Sections of Workers]: Programming Music for Industry: *and* The Statistical Method in Determining the Effects of Music in Industry.—D. D. Halpin & others. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 116–135.)

492. On Subjective Tones [Repetition of Criticism of Loose Usage of Term, and of the Procedure called "Measuring the Intensity of S.T. by the Method of Most Pronounced Beats"].—J. D. Trimmer. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, p. 136.) Prompted by Moe's paper (796 of 1943), which is written "in blunt disregard of two facts previously pointed out" (3726 of 1937).

493. Relation between Dissonance and Context.—P. A. D. Gardner & R. W. Pickford. (*Nature*, 25th Sept. 1943, Vol. 152, No. 3856, p. 358.) For a letter from C. G. Gray, and Pickford's reply, *see* issue for 13th Nov. 1943, No. 3863, p. 570.

494. The Problem of the Keyboard Instrument: II.—Ll. S. Lloyd. (*Phil. Mag.*, Sept. 1943, Vol. 34, No. 236, pp. 624–631.)

The writer ends: "The real problem of the keyboard instrument remains to be fully solved. It is: Why does the average ear of the trained musician, which knows that there is something wrong with the tuning of the harmonium, accept as satisfactory the tuning of the piano?" Some suggestions are made. For III *see* October issue, No. 237, pp. 674–684.

495. String under Intermittent Impulses, and Experimental Verification of Kar's Theory of Intermittent Action [as applied by Biswas to the Bowed String].—H. G. Mane & B. N. Biswas. (*Indian Journ. of Phys.*, April 1943, Vol. 17, Part 2, pp. 97–101.)

496. How Well Do I Hear? [Use of the Audiometer: Results of Public Health Survey & of World's Fair Tests].—M. B. Gardner. (*Bell Lab. Record*, Sept. 1943, Vol. 22, No. 1, pp. 6–11.) Supplementing Munson's article, 3055 of 1943.

497. ESTIMATION OF PERCENTAGE LOSS OF HEARING [Résumé of the Various Methods, and the Method formulated by the Council on Physical Therapy].—H. A. Carter. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 87–90.)

498. "AN INTRODUCTION TO BIOPHYSICS" [Book Review]—O. Stuhlman, Jr. (*Journ. Acous. Soc. Am.*, Oct 1943, Vol. 15, No. 2, p. 138.) "His lucid expositions will serve to bring many scientists up-to-date in fields not immediately their own."

499. ELECTRICAL RESPONSES FROM THE PRIMARY ACOUSTIC CENTRE OF THE FROG.—J. Ek & C. L. von Euler. (*Nature*, 31st July 1943, Vol. 152, No. 3848, p. 132.)

500. THE DETERMINATION OF SPEECH-INTELLIGIBILITY IN A CHANNEL WITH A LIMITED FREQUENCY BAND.—Zholdakov. (*See* 609.)

501. INVESTIGATIONS ON THE OPTIMUM RANGE OF VALUES OF THE TRANSMISSION EQUIVALENT IN TELEPHONIC COMMUNICATION.—F. Strecker & G. von Susani. (*E.N.T.*, Nov. 1942, Vol. 19, No. 11, pp. 241–252.)

502. RECORDING AUDIO ANALYSER [Western Electric Model RA-281 (for Frequencies 10–9500 c/s) for Noise & Vibration Study].—Western Electric Company. (*Electronics*, July 1943, Vol. 16, No. 7, pp. 100–103 and 210.)

503. ACOUSTIC ROOM AND TEST APPARATUS [for Automatic Recording of Microphone & Loudspeaker Characteristics, at Murray Hill Laboratory].—(*Bell Lab. Record*, Sept. 1943, Vol. 22, No. 1, p. 15.) For this laboratory *see* also 663, below.

504. ACOUSTIC LABORATORY IN THE NEW R.C.A. LABORATORIES [Princeton, New Jersey].—H. F. Olson. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 96–102.)

505. A DISCUSSION OF ACOUSTIC CELLS FOR AIRPLANE-ENGINE TEST BUILDINGS, *and* SOUND-PROOF AIRPLANE-MOTOR TEST CHAMBERS.—D. Fitzroy : I. E. Katel. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 106–113 : pp. 114–115.)

506. SOUND ATTENUATION IN AIR DUCTS [Metropolitan-Vickers Investigations, including the Testing of Lining Materials having a Good Absorption Coefficient at Low Frequencies].—B. G. Churcher & A. J. King. (*Engineering*, 17th Sept. 1943, Vol. 156, No. 4053, pp. 221–222.)

507. A SIMPLE METHOD OF MEASURING THE WAVELENGTH OF SOUND IN FREE AIR [Modification of Pulse Method (3264 of 1938) by Use of 500 c/s Oscillator supplying Non-Polarised Electrostatic Loudspeaker (giving 1000 c/s Note) : Error in Phase Determination is Smaller for the 2 : 1 Lissajous' Figure than for the Usual 1 : 1.]—E. G. Knowles. (*Journ. of Scient. Instr.*, Oct. 1943, Vol. 20, No. 10, p. 165.)

508. REVERBERATION IN SMALL GLASS TUBES [during & after Blowing : an Apparently Anomalous Resonance & Its Explanation].—S. M. Cox. (*Nature*, 25th Sept. 1943, Vol. 152, No. 3856, pp. 357–358.)

509. THE THERMOREGENERATION OF SOUND.—K. F. Theodortschik. (*Journ. of Phys.* [of USSR], No. 6, Vol. 2, 1940, pp. 437–440 : in German.) *See* 1991 of 1942, and for later work *see* 460 & 2390 of 1942.

510. THE EXTENSION OF THE DOPPLER PRINCIPLE TO DISCONTINUOUS PERIODIC EFFECTS [such as Time between Hits of Bullets fired by Machine Gun on Aeroplane in Motion].—L. Fleischmann. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 103–105.)

511. USE OF THE PIERCE INTERFEROMETER FOR MEASURING THE ABSORPTION OF SOUND IN GASES [including the Rigorous Theoretical Derivation of the Empirical Pielemeier Equation].—H. C. Hardy. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 91–95.)

512. ACOUSTICAL TABLES FOR AIR AND SEA-WATER [showing Relationships among Pressure Level, Sound Pressure, Intensity Level, Particle Velocity, & Product of Particle Displacement & Frequency].—W. L. Woolf. (*Journ. Acous. Soc. Am.*, Oct. 1943, Vol. 15, No. 2, pp. 83–86.)

513. SUPERSONICS FOR COMMUNICATION : EXPERIMENTAL RESULTS AND CIRCUIT SUGGESTIONS.—S. J. Weitzer. (*QST*, Oct. 1943, Vol. 27, No. 10, pp. 9–13.)

514. SUPERSONICS IN BIOLOGY [Researches with Concentrated Beams, using Concave Quartz Crystals (Grützmacher, 225 [and 2705] of 1936)].—J. G. Lynn & others. (*Electronics*, May 1943, Vol. 16, No. 5, pp. 154 and 156 : summary only.)

515. DETECTION OF FLAWS IN PLATES BY SUPERSONIC WAVES [using Flowing Water as Link Between Quartz Oscillator, Material, & Receiver].—A. Trost. (*Zeitschr. V.D.I.*, 12th June 1943, Vol. 87, No. 23/24, pp. 352–354.)

516. PROPAGATION OF ELASTIC WAVES THROUGH ELECTROLYTIC SOLUTIONS [only a Rise of Supersonic Velocity with Concentration has been reported previously : Heavy Alkali Halides show Decrease (enhanced at Higher Temperatures) followed by Increase : Implications].—A. K. Dutta & B. B. Ghosh. (*Indian Journ. of Phys.*, Feb. 1943, Vol. 17, Part 1, pp. 19–25.)

517. SUPERSONIC MEASUREMENTS IN CO_2 AND H_2O AT 98° C [Velocity & Absorption Behaviour], *and* SUPERSONIC MEASUREMENTS IN CO_2 AT 0° TO 100° C.—W. H. Pielemeier & W. H. Byers : W. H. Pielemeier. (*Journ. Acous. Soc. Am.*, July 1943, Vol. 15, No. 1, pp. 17–21 : pp. 22–26.) *See* also *Phys. Review*, 1st/15th July 1943, Vol. 64, No. 1/2, p. 44.

PHOTOTELEGRAPHY AND TELEVISION

518. A NEW OPTICAL-MECHANICAL SYSTEM OF TELEVISION ["permits Pictures to be transmitted practically with Any Desired Sharpness by means of Very Small Rotating Parts & Simple Light Sources (Incandescent Lamps)"].—O. B. Lurye. (*Journ. of Phys.* [of USSR], No. 3, Vol. 4, 1941, pp. 227–234 : in English.) The Russian original was dealt with in 3458 of 1943.

519. THE BASIC METHODS FOR THE PRODUCTION OF SPATIAL REPRESENTATIONS, ACCORDING TO WALUS'S CLASSIFICATION [in connection with Stereoscopic Cinematography].—W. Selle. (*Zeitschr. f. Instrum:kunde*, June 1943, Vol. 63, No. 6, pp. 207–214.)

520. LETTER AND PAPERS ON THE CATHODE FOLLOWER.—Ray : Kleen : Müller-Lübeck. (*See* 409/11.)

521. Emission-Type [Vacuum, Gas-Filled, & Secondary-Emission Types] Photoelectric Cells [Their Characteristics, Uses, & Circuits: Possible Future Development & Applications].—A. C. Lynch & J. R. Tillman. (*P. O. Elec. Eng. Journ.*, July 1943, Vol. 36, Part 2, pp. 43–46.) Among other things, greater total sensitivity may be attained—the theoretical limit has not yet been approached.

522. Ejection of Photoelectrons by X-Rays near the Angle of Total Reflection [including the Suggestion that the Investigation of Photocurrent at Very Small Glancing Angles of X-Rays may prove a Sensitive Method of analysing Photoelectric Emission].—N. B. Gorney. (*Journ. of Phys.* [of USSR], No. 3, Vol. 4, 1941, pp. 247–258: in English.)

MEASUREMENTS AND STANDARDS

523. On the Measurement of Dielectric Constants with the help of the Hollow Wave Guide.—G. Fejér & P. Scherrer. (*Helvet. Phys. Acta*, No. 7, Vol. 15, 1942/3, p. 645 onwards.)

A long paper on measurements in the 1–3 cm wave-range, using a guide of rectangular cross-section: for earlier work *see* 1176 of 1943 and back reference. The special properties of the $H_{0,m}$ wave in such a guide enable crystals to be investigated. A particularly simple procedure is given if the plate thickness is so chosen that the total phase jump can be made equal to that occurring in reflection at a metal piston of infinite conductivity; that is, equal to π. When this is the case, the required dielectric constant can be obtained from comparatively simple formulae, involving only the cross-sectional dimensions of the guide, the thickness of the plate, and the wavelength.

524. The Bolometer as a Power Meter at Ultra-Short Wavelengths: Remarks on a Previous Paper [3048 of 1942].—H. Meinke. (*E.N.T.*, Dec. 1942, Vol. 19, No. 12, p. 287.)

In that paper the writer described his trial of a "triple tie-line" (from Fränz's paper dealt with in 3240 of 1942), and came to the conclusion that with such a triple device adjustments were liable to occur where resonance effects produced appreciable losses of energy which vitiated the measurements and were difficult to detect. He now points out that this difficulty would not present itself when the device was used in the way in which Fränz used it, where the criterion for correct adjustment was the indication of maximum power in the bolometer beyond the quadripole.

525. A Garage for the Diode-Head of the Multi-Purpose Test Meter [3500 of 1943].—E. H. Simmonds. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, p. 380.)

526. Combination Modulation and Field-Strength Indicator and External S Meter.—I. F. Miller. (*QST*, Oct. 1943, Vol. 27, No. 10, p. 60.).

527. A New Test Method for Amplifiers and Components of Communication Engineering.—H. Knapp & A. Gerrmann. (*E.N.T.*, June 1943, Vol. 20, No. 6, pp. 137–140.)

The quantity most generally used to specify the transmission properties of a four-terminal network is the operative attenuation b_B. The usual way of measuring it is by comparison with a variable calibrated line, independent of frequency, connected to the same source (Fig. 1). An amplifier is measured in the same way, a fixed attenuation, larger than the amplification, being connected in front of it. Such a method has two disadvantages for production testing: (i) if the operative attenuation (or the amplification) is dependent on frequency, the test frequency must be adjusted carefully before each measurement to the correct value; and (ii) it is generally necessary to work the change-over switch *S*, and to vary the calibrated line, many times in order to make the output voltage from the quadripole equal to that from the calibrated line. Both these processes mean a waste of time and a more difficult operation.

The first objection, point (i), is removed by replacing the calibrated line by a quadripole of the same type as the quadripole under test and known to be correct: then there is no need to adjust the frequency to a definite value, since in comparing two similar quadripoles the exact frequency is unimportant. Point (ii) is avoided by changing the procedure to one in which the beat-note generator is swept over the desired frequency band, and simultaneously the difference between the output voltage provided by the quadripole under test and the standard quadripole is read on a difference-meter designed to measure the difference between two approximately equal a.c. voltages. Since the operative attenuation involves only the voltage magnitudes and not their phase, this instrument must not be phase-sensitive.

Now the easiest way to eliminate phase-influences is through rectification. Separate rectification of the two output voltages, and a comparison between the two rectified voltages, would fail in practice because the two transformations would not be likely to agree within 0.01 N over the suitable range (*see* the discussion on the requirements of the difference-meter, top of p. 138) of 4 N. So a common amplifying–rectifying system is supplied alternately (several times a second) with the two voltages by means of a relay ("Rls. 1" in Fig. 3). After amplification in the amplifier I the voltages are rectified, and the rectangular-wave-voltage of the envelope curve (Fig. 4b) is further amplified (in amplifier II) as an alternating voltage of the commutating frequency *F* (actually about 4.5 c/s). This amplified envelope voltage (Fig. 4c) serves as a measure of the difference between the two original voltages U_N and U_x: it is rectified (Fig. 4d) by the relay "Rls. 2" (which works synchronously with the first relay) and proceeds to the indicating instrument. The current through the latter will reverse its direction when U_N is not (as in Fig. 4) larger than U_x but smaller, so that in order that the instrument may measure what is required, the voltage beyond the second amplifier must vary not with the difference between U_x and U_N but with their ratio: this is accomplished by making the d.c. voltage across the resistance *R* (Fig. 3) control the amplification of the first amplifier I. The alternating voltage at amplifier II is thus roughly constant whatever the input voltage may be, but the residual dependence on this voltage is still liable to be too great (over the desired range of 4 N), and the second amplifier is therefore also controlled as to its amplification in the same manner: the magnitude of the voltage beyond it is thus made to depend only on the ratio of U_x to U_N.

The construction of the complete equipment is described. A recording instrument may be used, and the paper supply coupled to the shaft of the variable condenser of the beat-note generator.

528. Measurements on High-Frequency Cables and Feeders: Part I—Tuned-Circuit Substitution Methods [Series-Resistance & Parallel-Resistance Methods: Reactance-Variation Method, and Its Extension to Frequencies up to 200 or 300 Mc/s]:

PART II—IMPEDANCE, TRANSMISSION AND CROSS-TALK ATTENUATION MEASUREMENTS [including Localisation of Faults & Investigation of Impedance Irregularities].—R. F. J. Jarvis & J. C. Simmonds. (*P.O. Elec. Eng. Journ.*, July & Oct. 1943, Vol. 36, Parts 2 & 3, pp. 37–42 & 76–82.)

529. CORRESPONDENCE ON "MEDIUM- AND LONG-WAVE AERIAL SYSTEM FOR A RADIO LABORATORY" [3492 of 1943: the Cathode Follower suggested for Matching an Output Valve to a Low-Impedance Line].—S. N. Ray: T. B. Rymer. (*Journ. of Scient. Instr.*, Nov. 1943, Vol. 20, No. 11, pp. 181–182.) Rymer's reply, unfavourable to the use of a cathode follower in the case in point, includes two improvements introduced into his amplifier since his paper went to press, and makes one correction.

530. REFRESHER FOR DIELECTRIC CALCULATIONS [Correlation of Fundamental Characteristics of Dielectric with Measured Quantities such as Current, Loss, Power Factor, & Capacitance].—E. W. Greenfield. (*Elec. Engineering*, Oct. 1943, Vol. 62, No. 10, pp. 445–448.)

531. PAPERS ON THE USE OF COMPRESSED-GAS AND MICA CONDENSERS IN MEASURING TECHNIQUE.—Keller: Ebinger & Linder. (*See* 563 & 564.)

532. THE TECHNIQUE OF MATERIALS TESTING: IV—THERMAL TESTS [continued from October Issue].—W. D. Owen. (*BEAMA Journal*, Nov. 1943, Vol. 50, No. 77, pp. 337–342.) For previous parts *see* 157 of January.

533. INSULATION TEST SET [designed by J. S. Forrest (2465 of 1943)].—British Thomson-Houston. (*Journ. of Scient. Instr.*, Nov. 1943, Vol. 20, No. 11, p. 182.)

534. LABORATORY FOR HIGH-TENSION RESEARCH.—Oerlikon Company. (*Engineer*, 1st Oct. 1943, Vol. 176, No. 4577, pp. 269–272.)

535. THE PRODUCTION OF QUARTZ CRYSTALS [Wartime Alternate Methods: Photographs, Circuit Diagrams, & Captions].—DX Crystal Company. (*Communications*, May 1943, Vol. 23, No. 5, pp. 22–23 and 64, 65.)

536. THE VIBRATING QUARTZ IN COMMUNICATIONS TECHNIQUE: PART II.—Arens. (*See* 424.)

537. TIME SIGNALS [Published Time should be arranged, say Once a Week, when Master Clock at Power Station is as nearly as possible in Synchronism with Greenwich Time: for setting Synchronous Clocks & Time-Switches].—E. W. Davies. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 381–382.)

538. INVESTIGATIONS ON AN OSCILLATION GENERATOR WITH RESISTANCE AND CAPACITANCE AS THE FREQUENCY-DETERMINING COMPONENTS: *RC* GENERATOR [Bridge-Stabilised Type].—Zaiser. (*See* 425.)

539. GAS-TUBE HARMONIC GENERATOR.—L. G. Kersta. (*Bell Lab. Record*, Oct. 1943, Vol. 22, No. 2, pp. 53–58.)

"The studies [originally in connection with the search for methods of obtaining currents of high and closely-controlled frequency for carrier systems] showed that gas tubes [mercury-vapour or argon low-pressure triodes] can produce much higher frequencies than had been thought possible, if operated in a new circuit which accelerates the ionisation and deionisation of the tube (Shepherd, 1655 of 1943). The harmonics thus generated provide standard frequencies, as high as 25 Mc/s, for calibrating oscillators. The sharp current pulses are useful in testing television systems, for making phase-distortion measurements in coaxial cables, and in high-speed triggering devices."

540. A RECORDING VOLTMETER WITH EXTENSIVELY SUPPRESSED LOWER-END READINGS [for the Monitoring of an Exact Voltage: Bridge Circuit with Voltage-Dependent Resistances].—W. Oesinghaus. (*Elektrot. u. Masch:bau*, 23rd May 1941, Vol. 59, No. 21/22, pp. 250–251.)

541. ELECTRIC INSTRUMENTS: AMMETERS, VOLTMETERS, WATTMETERS: PERFORMANCE AT HIGHER THAN NORMAL OPERATING FREQUENCIES: SIMPLE, PRACTICAL METHODS OF CHECKING ACCURACY.—S. C. Richardson. (*Gen. Elec. Review*, Oct. 1943, Vol. 46, No. 10, pp. 565–569.)

542. STANDARDISATION AS APPLIED TO INDUSTRIAL ELECTRICAL INSTRUMENTS.—K. Edgcumbe. (*Journ. I.E.E.*, Part I, July 1943, Vol. 90, No. 31, pp. 265–281: Discussions pp. 282–297.) Summaries were referred to in 517 of 1943.

543. INSTRUMENT TRANSFORMERS [and a New Compensator for reducing Errors].—A. Hobson. (*Electrician*, 26th Nov. 1943, Vol. 131, No. 3417, pp. 539–541: summary & Discussion of I.E.E. paper.)

SUBSIDIARY APPARATUS AND MATERIALS

544. A COMPACT, TWO-UNIT CATHODE-RAY OSCILLOSCOPE [Sensitivity from 2 cm/V to 0.04 cm/V, Frequency Range from Zero to 200 kc/s: Time Base Linear from 1 c/s to 10 kc/s: Independent Controls on Both Axes].—Hadfield. (*P.O. Elec. Eng. Journ.*, April 1943, Vol. 36, Part 1, pp. 1–5.)

545. A BIBLIOGRAPHY OF ELECTRON MICROSCOPY.—Marton & Sass. (*Journ. Applied Phys.*, Oct. 1943, Vol. 14, No. 10, pp. 522–531.)

546. COHERENCE OF RAYS IN THE FORMATION OF IMAGES IN THE MICROSCOPE [and the Part it plays in the Visibility of Objects].—Rogestwensky. (*Journ. of Phys.* [of USSR], No. 4, Vol. 4, 1941, pp. 293–317: in English.) For previous work *see* 4404 of 1940.

547. THE THEORY OF SECONDARY ELECTRON EMISSION FROM DIELECTRICS AND SEMICONDUCTORS.—Kadyschewitsch. (*Journ. of Phys.* [of USSR], No. 4, Vol. 4, 1941, pp. 341–348: in German.) The Russian original was dealt with in 2423 of 1941.

548. GAS-TUBE HARMONIC GENERATORS [giving Current Pulses useful in High-Speed Triggering Devices, etc.].—Kersta. (*See* 539.)

549. A NEW METHOD OF GENERATING VERY HIGH DIRECT VOLTAGES [240 kV: Defects of Previous Methods: Economical & Straightforward Method developed from the "Micafil" Multiplier Circuit, with Condenser Stages each having Feeding Condenser & Rotating Switch-Arm].—Böning. (*Elektrot. u. Masch:bau*, 14th Feb. 1941, Vol.

59, No. 7/8, pp. 69–73.) For a "Micafil" generator for 3 Mv *see* issue for 28th Feb. 1941, No. 9/10, p. 107, and Imhof, 787 of 1940.

550. FURTHER EXAMINATION OF THE PRODUCTION OF IONS BY ELECTRONS OSCILLATING IN AN ELECTRIC FIELD [531 of 1941].—Korsunski & Shavlo. (*Journ. of Phys.* [of USSR], No. 3, Vol. 4, 1941, pp. 285–286: summary only, in English.)

551. THE SUPPRESSION OF THE EVAPORATION OF MERCURY [limiting Pumping Speed & Ultimate Vacua] FROM DIFFUSION PUMPS.—Bull & Klemperer. (*Journ. of Scient. Instr.*, Nov. 1943, Vol. 20, No. 11, pp. 179–181.) In the new equipment described, all critical surfaces in the pump are water-cooled, the trap offers unusually small pumping resistance, and a metal shutter replaces the conventional tap.

552. INITIATION OF GLOW DISCHARGES [Investigation (at Pressures below 5 cm Hg) with Two-Beam Oscillograph, using 50 c/s A.C., but checking with D.C: Comparison between Behaviours of Various Gases: Occurrence of Streamers: etc.]. —Craggs & Meek. (*Nature*, 2nd Oct. 1943, Vol. 152, No. 3857, pp. 386–387.)

553. POSITIVE COLUMN PLASMA IN A STRONG LONGITUDINAL MAGNETIC FIELD [and the Question of Contraction].—Reichrudel & Spiwak. (*Journ. of Phys.* [of USSR], No. 3, Vol. 4, 1941, pp. 211–226: in English.)

English version of the paper referred to in 2526 of 1941: *see* also 1962 of 1943. Authors' summary:—"The effect of a longitudinal magnetic field on the positive-column plasma is investigated at low pressure (10^{-3} to 1 mm Hg) in mercury vapour or argon. . . . Data were obtained concerning plasma contraction in the magnetic field [and reasons for the departure from Tonks's theory are suggested]. The peculiarities of probe characteristics for strong magnetic fields are discussed. Parallel measurements carried out with an anode made up of concentric rings are in good agreement with results of the probe measurements."

554. ERRATUM: THE MERCURY-ARC CATHODE [2806 of 1942].—Smith. (*Phys. Review*, 1st/15th July 1943, Vol. 64, No. 1/2, p. 40.)

555. ON THE "CLEAN UP" OF RARE GASES IN HOLLOW CATHODES, AND THE PROCESSES INVOLVED THEREIN.—Bartholomeyczyk. (*Ann. der Phys.*, 13th May 1943, Vol. 42, 1942/3, No. 7/8, pp. 534–560.)

556. THE MECHANISM OF STRIKING IN HIGH-VOLTAGE GAS-FILLED "SECTIONAL" RECTIFIERS.—Rakov & Fetisov. (*Izvestiya Elektroprom. Slab. Toka*, No. 9, 1940, pp. 42–47.)

Continuing a previous work (322 of 1940), parameters determining the operation of the rectifiers are established and the mechanism of striking is investigated, taking into account the effect of the circuit in which the rectifier operates.

557. ON THE THEORY OF THE MAINS-RECTIFIER CIRCUITS [and a Comparison between the Two Principal Types].—Holzwarth. (*See* 426.)

558. POLYPHASE BRIDGE CONNECTION FOR CURRENT CONVERTERS.—Wasserrab. (*Elektrot. u. Masch: bau*, 3rd Jan. 1941, Vol. 59, No. 1/2, pp. 3–9.)

559. CHARACTERISTICS AND APPLICATIONS OF SELENIUM-RECTIFIER CELLS [made by General Electric Company].—Harty. (*Elec. Engineering*, Oct. 1943, Vol. 62, No. 10, Transactions pp. 624–629.)

560. TRANSITIONAL RESISTANCES OF SEMICONDUCTORS.—Davydov. (*Journ. of Phys.* [of USSR], No. 4, Vol. 4, 1941, pp. 335–339: in English.) The Russian original was dealt with in 2545 of 1941: for other work *see* 828 of 1942.

561. DISCUSSION ON "A DIFFERENTIAL ELECTRONIC STABILISER FOR ALTERNATING VOLTAGES, AND SOME APPLICATIONS" [1957 of 1943: including a Query as to Variation of the "Grid" Frequency].—Glynne. (*Journ. I.E.E.*, Part II, Oct. 1943, Vol. 90, No. 17, pp. 367–368.)

562. CURRENT AND POTENTIAL DISTRIBUTION IN "SHORTED-EDGE" ROLL-TYPE CONDENSERS.—Leiterer. (*See* 449.)

563. THE USE OF COMPRESSED-GAS CONDENSERS FOR MEASURING PURPOSES.—Keller. (*Elektrot. u. Masch: bau*, 20th June 1941, Vol. 59, No. 25/26, pp. 292–296.)

564. MICA CONDENSERS, THEIR PROPERTIES AND IMPORTANCE IN MEASURING TECHNIQUE.—Ebinger & Linder. (*Elektrot. u. Masch: bau*, 20th June 1941, Vol. 59, No. 25/26, pp. 286–292.) From the Siemens & Halske laboratories.

565. MICA FOR WAR PURPOSES [and the Use of the Stained Variety: Test Methods & Quality Control: the Mica Mission Agreements.]—Given. (*Bell Lab. Record*, Oct. 1943, Vol. 22, No. 2, pp. 60–63.) For other recent work on mica *see* 527, 2495, 2818, & 3369 of 1942 and 2492, 2829, & 3541/4 of 1943.

566. ON THE DEFINITIONS OF THE TERMS SURFACE LEAKAGE AND SURFACE-LEAKAGE PATH [and the Need for Clarification].—von Cron. (*E.T.Z.*, 17th June 1943, Vol. 64, No. 23/24, p. 324.) Arising from recent researches, including the writer's own (3133 of 1943).

567. REFRESHER FOR DIELECTRIC CALCULATIONS.—Greenfield. (*See* 530.)

568. DOSIMETRY AND LOCAL DISTRIBUTION OF ENERGY IN THE ELECTRIC HIGH-FREQUENCY FIELD [of Interest also in Insulation Research: Use of Electrodeless-Discharge Measuring Instrument: Examples of Short-Wave Field Distribution].—Lion. (*See* 706.)

569. ON THE THERMAL CONDUCTIVITY OF DIELECTRICS AT TEMPERATURES HIGHER THAN THE DEBYE TEMPERATURE.—Pomeranchuk. (*Journ. of Phys.* [of USSR], No. 3, Vol. 4, 1941, pp. 259–268: in English.) *See* also 3112 of 1942 and back reference.

570. THE DIPOLE THEORY AND THE CHARACTERISTICS OF ORGANIC INSULATORS.—Wall. (*Engineering*, 30th July 1943, Vol. 156, No. 4046, pp. 81–83.)

571. ELECTRICAL PROPERTIES OF SOLIDS: XIII [Polymethyl Acrylate, etc.].—Mead & Fuoss. (*Journ. Am. Chem. Soc.*, Oct. 1942, Vol. 64, p. 2389 onwards.) *See* Abstract 1248, *Sci. Abstracts*, Sec. A, 1943.

572. THE USE OF POLYVINYL CHLORIDE IN ELECTROTECHNICS.—Paselli. (*Elektrot. u. Masch:bau*, 24th Oct. 1941, Vol. 59, No. 43/44, pp. 517–518: long summary only.)

573. PRODUCTION AIDS [Various Developments suggested by Workers, including the Arbor for Ceramic Coil Forms (202 of January)].—General Electric. (*Communications*, June 1943, Vol. 23, No. 6, pp. 40 and 70, 86.)

574. "KUNSTHARZPRESSSTOFFE IM MASCHINENBAU" [Synthetic-Resin Plastics in Machine Construction: including the Testing of Finished Parts: Book Review].—Weigel. (*Schweizer Arch. f. angew. Wiss. u. Tech.*, May 1943, Vol. 9, No. 5, p. 164.)

575. DIPOLE MOMENTS OF THE CHIEF CONSTITUENTS OF LAC AND ROSIN.—Bhattacharya. (*Indian Journ. of Phys.*, June 1943, Vol. 17, Part 3, pp. 153–162.) The last pages are missing in some copies.

576. ELECTRICAL AND MECHANICAL PROPERTIES OF THE SYSTEM BUNA S–GILSONITE [Loss of Electrical Quality avoided when Gilsonite is used as Filler instead of Carbon Black: primarily for Submarine Cables].—Selker, Scott, & McPherson. (*Journ. of Res. of Nat. Bur. of Stds.*, Sept. 1943, Vol. 31, No. 3, pp. 141–161.)

577. INTRINSIC ELECTRIC STRENGTH AND CONDUCTIVITY OF VARNISH FILMS, AND THEIR VARIATION WITH TEMPERATURE.—Thomas & Griffith. (*Journ. I.E.E.*, Part I, Nov. 1942, Vol. 89, No. 23, pp. 487–498.)

578. THE CRACKING OF THE ENAMEL LAYER OF ENAMELLED ALUMINIUM WIRE AT ORDINARY AND HIGHER TEMPERATURES.—Greulich. (*E.T.Z.*, 6th May 1943, Vol. 64, No. 17/18, pp. 241–242.)

579. RADIANT HEATING FOR INDUSTRIAL PURPOSES [Abridgement of Institute of Fuel Paper], *and* INFRA-RED INDUSTRIAL LAMPS: SOME OF THE RESULTS OF PRACTICAL APPLICATION.—Andrew & Chamberlain: Anon. (*Engineering*, 24th Sept. & 8th Oct. 1943, Vol. 156, Nos. 4054 & 4056, pp. 245–246 & 298–300: *Electrician*, 16th July 1943, Vol. 131, No. 3398, pp. 59–60.) *See* also 3100 of 1942.

580. SUPER-STRENGTH BONDING ["Reanite" Bonding Process for uniting Rubber, Plastics, Wood, etc., to Metal or to Each Other].—U.S. Stoneware. (*Scient. American*, Oct. 1943, Vol. 169, No. 4, p. 185.) *See* also *Electronics*, Sept. 1943, Vol. 16, No. 9, p. 270.

581. TEMPERATURE-DEPENDENT RESISTANCES ("HEISSLEITER") AND THEIR APPLICATION IN TECHNIQUE [Survey, including the Urdox Regulator as Variable H.F. Resistance].—Sachse. (*Elektrot. u. Masch:bau*, 28th Feb. 1941, Vol. 59, No. 9/10, pp. 107–108: summary only.)

582. ON THE MECHANISM UNDERLYING THE BEHAVIOUR OF NON-LINEAR RESISTORS [Experimental & Theoretical Investigation of Carborundum Resistors].—Braun & Busch. (*Helvet. Phys. Acta*, No. 6, Vol. 15, 1942, p. 571 onwards: abstract in *Sci. Abstracts*, Sec. B, May 1943, Vol. 46, No. 545, p. 81.)

583. THE PRODUCTION OF FIXED CARBON RESISTORS [and the Use of Automatic Machines].—Carter. (*P.O. Elec. Eng. Journ.*, April 1943, Vol. 36, Part 1, pp. 6–9.)

584. ELECTROMAGNETIC SCREENS WITH JOINTS AND GAPS.—Kaden. (*See* 427.)

585. MAGNETIC PROPERTIES OF NICKEL SUPERSATURED WITH CARBON.—Gerlach & von Rennenkampff. (*Naturwiss.*, Vol. 31, 1943, p. 96.)

586. MAGNETIC PROPERTIES OF SOLID SOLUTIONS: III—THE PARAMAGNETIC ALLOYS OF COPPER AND NICKEL.—Kaufmann & Starr. (*Phys. Review*, 1st/15th June 1943, Vol. 63, No. 11/12, pp. 445–450.) For previous parts *see* 878 of 1942.

587. A MAGNETIC STUDY OF THE TWO-PHASE IRON–NICKEL ALLOYS: II.—Hoselitz & Sucksmith. (*Proc. Roy. Soc.*, Series A, 6th May 1943, Vol. 181, No. 986, pp. 303–313.) For I *see* 3204 of 1940.

588. NEW FORMULAE FOR THE DETERMINATION OF MAGNETIC QUANTITIES H, l, M, and m [by the Deflection Magnetometer].—Singal. (*Current Science* [Bangalore], April 1943, Vol. 12, No. 4, p. 113.)

589. A STUDY OF IRON CORES: TYPES, METHODS OF PRODUCTION, AND APPLICATIONS AT L.F. AND H.F. ANALYSED [including the Raising of "Q" by Severe Magnetic Shock].—White. (*Communications*, June 1943, Vol. 23, No. 6, pp. 42–43 and 46, 48, 80, 83.) Performance at ultra-high frequencies is dealt with from p. 80 onwards.

590. EFFECT ON B/H CURVES OF ELECTRIC CURRENTS THROUGH A TRANSFORMER CORE [Investigation with Cathode-Ray Oscillograph: Broadening of Hysteresis Loop (over the Two Quarter-Cycles when Magnetising Current is Increasing) on passing A.C. (Not D.C.) through Core].—Sarna, Kalia, & Dhingra. (*Current Science* [Bangalore], Sept. 1943, Vol. 12, No. 9, pp. 251–252.) "The change depicted is quite peculiar, and although induced effects will be there, the atomic magnetic properties may also have to be considered."

591. ON THE THEORY OF MAGNETISATION AND HYSTERESIS CURVES OF POLYCRYSTALLINE FERROMAGNETICS.—Kondorsky. (*Journ. of Phys.* [of USSR], No. 3/4, Vol. 6, 1942, pp. 93–110: in English.)

"In a previous paper (3197 of 1940) an attempt was made to obtain, on the basis of the modern conceptions of magnetisation processes, an approximate theory of hysteresis for cases when the magnetic interaction between the separate parts of a body can be neglected (there was examined a ferromagnetic with infinitely long cylindrical homogeneous grains). In the present paper is considered the possibility of certain approximate theory of the magnetic properties of a polycrystalline ferromagnetic, taking into account the magnetic interaction between its separate parts." The theoretical curves constructed by means of the formulae obtained are compared with curves of actual materials: the comparison is made with a normal magnetisation curve for cobalt and also with the normal curves for iron, steel, and permalloy in the region of small values of magnetisation. "The comparison shows that the theory describes the processes of magnetisation of polycrystalline bodies sufficiently well. For two specimens (iron and permalloy) the theoretical curves coincide exactly with the experimental ones in the regions of small magnetisation values." For another paper by the same writer, "On the theory of the normal and ideal magnetisation curves of a polycrystalline ferromagnetic," see *Comptes Rendus (Doklady) de l'Acad. des Sci. de l'URSS*, No. 5, Vol. 32, 1941, pp. 323–326: in English.

592. TELEPHONE RELAYS [and Their New Applications: Notice of Booklet to assist Designers in Their Choice of Type].—Inter-Services Components Manufacturers' Council. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, p. 383.)

593. SUNVIC ENERGY REGULATOR [Bimetallic Device as Alternative to Variable Rheostat: saves Energy and is Independent of Supply-Voltage Fluctuations].—Sunvic Controls. (*Electronic Eng:g*, Oct. 1943, Vol. 16, No. 188, p. 208.)

594. BIMETALLIC ELEMENTS [Chace Thermostatic Bimetal No. 6850: Electrical Resistivity considerably Higher than Previously Available: High Thermal-Deflection Rate].—Chace Company. (*Journ. of Scient. Instr.*, Nov. 1943, Vol. 20, No. 11, p. 183.)

595. FUZIT ELECTRODE WIRE JOINTER [using "Stanelco" Method of Fusing instead of Soldering].—Stanelco Products. (*Electrician*, 26th Nov. 1943, Vol. 131, No. 3417, p. 542.) For the "Rakos" process *see* 3564 of 1943.

596. LOW-TENSION SOLDERING IRONS: HINTS ON CONSTRUCTION.—(*Wireless World*, Nov. 1943, Vol. 49, No. 11, pp. 340–341.)

597. CONSERVATION OF TIN IN SOFT SOLDERS.—Colwell & Lang. (*ASTM Bulletin*, Aug. 1943, No. 123, pp. 37–43.)

598. INVESTIGATION OF LOW-TIN AND TIN-FREE SOLDERS: INDUSTRIAL CHEMISTRY CIRCULAR No. 2.—Worner & others. (*Journ. of Council for Scient. & Indust. Res., Australia*, Aug. 1943, Vol. 16, No. 3, p. 203: summary only.)

599. LEAD-ALLOY COATING FOR COPPER WIRE: SATISFACTORILY REPLACES SCARCE TIN.—Anaconda Wire & Cable. (*Scient. American*, Sept. 1943, Vol. 169, No. 3, p. 114.)

STATIONS, DESIGN AND OPERATION

600. 100 MEGACYCLES AND BEYOND, FOR RELAY TRANSMISSION [WGAR's Relay Broadcast Equipment, with Mobile Unit WEMV: a Three-Element Aerial with Griffin Inductive Coupling (permitting Continuous Rotation): Experiences on 225.6 Mc/s: etc.].—Widlar. (*Communications*, June 1943, Vol. 23, No. 6, pp. 11–13.)

601. "HANDY ANDY": a 112 Mc/s HAND SET FOR WERS.—Palmer. (*QST*, Oct. 1943, Vol. 27, No. 10, pp. 35–38.)

The diagram of this "handie-talkie" resembles that of a transceiver in that the same valve is used for oscillator and detector, but separate tank circuits are provided so that the receiver and transmitter may be tuned independently and each circuit designed for best performance, without compromise.

602. WERS DURING THE MAY MISSISSIPPI FLOODS, *and* CONSTRUCTIONAL ASPECTS OF WERS MOBILE INSTALLATIONS: SOME SUGGESTIONS FOR CAR GLOVE-COMPARTMENT JOBS.—Keating: Forster. (*QST*, Aug. 1943, Vol. 27, No. 8, pp. 30–33: pp. 34–37.)

603. MESSAGE HANDLING IN WERS: SPEEDING-UP OPERATION BY ASSIGNING MESSAGE PRIORITIES AND REVISING CONTROL-STATION PROCEDURE.—Russell & King. (*QST*, Aug. 1943, Vol. 27, No. 8, pp. 42–43 and 96.)

604. WERS FOR SEVEN MILLION PEOPLE: HOW NEW YORK CITY'S CD-WERS SYSTEM IS ORGANISED.—Long & Kenney. (*QST*, Oct. 1943, Vol. 27, No. 10, pp. 14–18.)

605. "WERS" CALLING: RADIO AMATEURS AND OTHER VOLUNTEER WORKERS ARE SETTING UP A COMMUNICATION SYSTEM FOR WAR OR PEACE.—Grant. (*Scient. American*, July 1943, Vol. 169, No. 1, pp. 23–25.) Major-General Grant writes from the Protection Branch, U.S. Office of Civilian Defence. For WERS *see* numerous past abstracts in this Section.

606. POLICE-UTILITY SYSTEMS: RADIO AND TELEPHONE SERVICE IN NEW ENGLAND: *also* THE MASSACHUSETTS STATE POLICE SYSTEM: *and* UTILITY COMMUNICATIONS [in Gas Supply Systems].—Gifford: MacLean: Murphy. (*Communications*, May 1943, Vol. 23, No. 5, pp. 30–32: pp. 32, 34: pp. 73–75.)

607. FREQUENCY MODULATION FOR POWER-LINE CARRIER CURRENT [Investigations on Frequencies below 200 kc/s, using a 1:1 Deviation Ratio: Usual Advantages of F.M. obtained even with this Low Deviation: Special Suitability for Power-Line Working].—Kenefake. (*Elec. Engineering*, Oct. 1943, Vol. 62, No. 10, Transactions pp. 616–619.) Unlike wide-swing f.m., the small-deviation f.m. delivers "usable intelligence" from a signal whose signal/noise ratio is less than 2.

608. VARIOPLEX TELEGRAPH PATENT [3781 of 1942] IS GRANTED WITH 107 CLAIMS.—Holcomb. (*Electronics*, May 1943, Vol. 16, No. 5, pp. 190 and 192.)

609. THE DETERMINATION OF SPEECH-INTELLIGIBILITY IN A CHANNEL WITH A LIMITED FREQUENCY BAND [in connection with Secrecy Systems].—Zholdakov. (*Izvestiya Elektroprom. Slab. Toka*, No. 12, 1940, pp. 32–41.)

Modern conceptions regarding the effect on speech-articulation of the width of the frequency band transmitted are reviewed, and methods are indicated for determining articulation for narrow frequency bands. The results obtained are applied to the case of scrambler secrecy equipment, and the minimum number of the frequency bands is determined, into which the main band should be split up to ensure that speech will remain unintelligible from any one of these bands if this is picked up by an unauthorised listener.

Since no information is yet available on the relationship between syllables of various forms in the Russian language, the discussion is based on Bell Laboratories' experiments with the English language, but it is not expected that this will seriously invalidate the conclusions reached.

610. BRIEF REMARKS ON THE POSSIBILITY OF NARROWING THE FREQUENCY BAND BY SLOWING DOWN THE TRANSMISSION.—Gadiev. (*Izvestiya Elektroprom. Slab. Toka*, No. 12, 1940, pp. 41–44.)

It is suggested that the material should be first recorded and then transmitted at a lower speed. If the times of recording and transmission are t_1 and t_2 respectively, then the required frequency band will be $k\,(=t_2/t_1)$ times narrower than that for ordinary transmission. In practice the band-width is limited only by considerations of the frequency stability of the transmitter, which under the present conditions enables it to be narrowed down to a few hundred cycles. With this method of transmission the level of interference is also reduced k times.

Since the programme so transmitted is unintelligible unless it is recorded and reproduced again at the normal speed, a certain amount of secrecy can be achieved by using this method.

611. Enemy Airborne Radio Equipment.—Edwards. (*Electrician*, 26th Nov. 1943, Vol. 131, No. 3417, p. 533: *Elec. Review*, 3rd Dec. 1943, Vol. 133, No. 3445, pp. 753–754: summaries of I.E.E. papers.)

612. Britain's Radio System for the Armed Forces. —Norris. (*Communications*, May 1943, Vol. 23, No. 5, pp. 11–12 and 58.)

613. Commando Signals: Wireless in a "Corps d'Élite."—(*Wireless World*, Nov. 1943, Vol. 49, No. 11, pp. 332–334.)

614. History and Organisation of Wireless Communication in Italy.—Pirani. (*Zeitschr. f. Fernmeldetech.*, 20th May 1943, Vol. 24, No. 4/5, pp. 64–66.) Based on a 1941 paper by G. Provenza.

615. Development of American Broadcasting Stations at the Beginning of 1942 [with Data regarding Power, Numbers, & Numbers of Receivers.]—von Reding. (*E.T.Z.*, 17th June 1943, Vol. 64, No. 23/24, p. 327: summary only.)

616. Reorganising Broadcasting: Some Drastic Proposals examined [Cossor Research Laboratories' Plan (K. I. Jones & D. A. Bell, I.E.E. Paper) and *Economist* Article].—(*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 351–354.) Editorial followed by summaries of the two proposals.

GENERAL PHYSICAL ARTICLES

617. The Potential Distribution in an Infinitely Long Conducting Hollow Cone, due to Regions of Electric Charge disposed Symmetrically to the Cone Axis.—Schmidl. (*Ann. der Phys.*, 21st July 1943, Vol. 43, No. 3, pp. 193–202.)

618. The Calculation of the Forces on Cylindrical Conductors in Plane Electrostatic and Electromagnetic Fields [by Methods based on Theory of Functions & specially used in Hydrodynamics].—Müller. (*See* 3292 of 1943).

619. Ejection of Photoelectrons by X-Rays near the Angle of Total Reflection.—Gorney. (*See* 522.)

620. The Quantum Theory of Radiation of an Electron uniformly moving in a Medium [with Velocity Greater than Phase Velocity of Light in that Medium].—Ginsburg. (*Journ. of Phys.* [of USSR], No. 6, Vol. 2, 1940, pp. 441–452: in English). *Cf.* 555 of 1942 and back reference: also 3592 of 1943 (Čerenkov radiation).

621. An Integral related to the Radiation Integrals [with Table of Values].—Powell. (*Phil. Mag.*, Sept. 1943, Vol. 34, No. 236, pp. 600–607.)

622. Magnetic Potential Energy [Treatment using Hamiltonian Function defined in Terms of Mechanical Momentum: providing Simple Interpretation of the Gyromagnetic Experiments & Electron Spin].—Warburton. (*Phys. Review*, 1st/15th June 1943, Vol. 63, No. 11/12, p. 460: summary only.)

623. Magnetolysis and the Electric Field around the Magnetic Current [Experimental Demonstration].—Ehrenhaft. (*Phys. Review*, 1st/15th June 1943, Vol. 63, No. 11/12, pp. 461–462: summary only.) *See* also 2867 of 1943.

624. "Reports on Progress in Physics: Vol. 9 (1942/43)" [Book Review].—Mann (Edited by). (*See* 399.)

MISCELLANEOUS

625. A Problem in the Summation of Series [in connection with Electromagnetic Theory (Cavity Resonators)].—Goddard. (*See* 401.)

626. On Certain Non-Linear Differential Equations of the Second Order.—Levinson. (*Proc. Nat. Acad. Sci.*, July 1943, Vol. 29, No. 7, pp. 222–223.) For a correction *see* September issue, No. 9, p. 281.

627. Method of Integration of Relaxation Equations by Successive Arcs.—Parodi. (*Comptes Rendus* [Paris], 5th/27th Oct. 1942, Vol. 215, No. 14/17, pp. 268–270.)

628. The Probability-Integral of the t-Function [and the Unsuitability, for Engineer & Chemist, of the Statistical Tables available (constructed for Biological Purposes): Thornhill's t-Curves].—Evans. (*Engineering*, 8th Oct. 1943, Vol. 156, No. 4056, p. 295.) With special reference to corrosion tests.

629. Simplified Method of utilising the Single Mean "t" Test [in Application of Statistics to Telecommunications Engineering]. — Mumford. (*P.O. Elec. Eng. Journ.*, April 1943, Vol. 36, Part 1, pp. 18–19.) A supplement to Doust & Josephs' series dealt with in 3550 of 1941 and 2551 of 1942. "Removing much of the drudgery."

630. A General Test for Finding whether Two Random Samples are Consubstantial: *also* A Note on the Problem of k Samples: *and* Statistical Formulae [New Variance Formula for k Varieties].—Savur: Nair: Ayyangar. (*Current Science* [Bangalore], Nov. 1940, Vol. 9, No. 11, pp. 491–492: April 1943, Vol. 12, No. 4, pp. 112–113: May 1943, No. 5, p. 145.)

631. "The Advanced Theory of Statistics" [Book Review].—Kendall. (*Nature*, 16th Oct. 1943, Vol. 152, No. 3859, pp. 431–432.)

632. Computation of Polynomial Functions by Summation of Finite Differences [Quick Approximate Method for use with Calculating Machine].—Bennett. (*Journ. Opt. Soc. Am.*, Sept. 1943, Vol. 33, No. 9, pp. 519–526.)

633. Harmonic Analysis by Photographic Method [Fourier-Series Coefficients determined from Area Measurements on Photographs of Function under Analysis, with Graph Sheet formed into Series of Alternately Convex & Concave Semicylinders: Example, and Comparison with Values determined Analytically].—Straiton & Terhune. (*Journ. Applied Phys.*, Oct. 1943, Vol. 14, No. 10, pp. 535–536.)

634. Non-Commutative Algebra [and the Measurement of Angles in "Millirevs" or "Jots": Convenience for Vectorial Work & in Other Ways].—Turnbull. (*Electrician*, 8th Oct. 1943, Vol. 131, No. 3410, p. 352.)

635. "Laplace-Transformation: eine Einführung für Physiker, Elektro-, Maschinen-, und Bauingenieure" [Book Review].—Hameister. (*T.F.T.*, April 1943, Vol. 32, No. 4, p. 96.) A distinctly scathing review by H. Schulz, who quotes various existing sources for the information.

636. "Mathematische Grundbegriffe für Fernmeldetechniker" [Book Review].—Rinkow. (*T.F.T.*, April 1943, Vol. 32, No. 4, p. 96.)

637. Elementary A.C. Mathematics: Part VII—Power, Power Factor, Losses in Reactances.—Grammer. (*QST*, Aug. 1943, Vol. 27, No. 8, pp. 56–59.) For some previous parts *see* 2349 & 2683 of 1943.

638. A Universal Resonance Chart [for Electrical & Mechanical Systems].—Yates. (*See* 429.)

639. The Construction of Nomograms.—Woods. (*Distribution of Electricity*, Oct. 1943, Vol. 16, No. 152, pp. 184–187.) Including references to a book "The Nomogram" by Allcock & Jones.

640. "Empirical Equations and Nomography" [Book Review].—Davie. (*Electronics*, June 1943, Vol. 16, No. 6, pp. 343 & 344.)

641. Determination, by the "Condition of Least Inaccuracy," of the Coefficients of a Formula representing an Experimental Curve in which they figure Linearly.—Vernotte. (*Comptes Rendus* [Paris], 7th/21st Dec. 1942, Vol. 215, No. 23/25, pp. 568–570.) Continuation of the work dealt with in 2547 of 1942 (and *see* ref. "2").

642. "Mathematics Dictionary" [Book Review].—James. (*Elec. Engineering*, July 1943, Vol. 62, No. 7, p. 329.)

643. Graphical Symbols [Tabulation].—(*Electronics*, April 1943, Vol. 16, No. 4, pp. 84–87.)

644. *Wireless World* Brains Trust: "Oddness" of Standard Voltages [Origin of Factor of 11 entering into Transmission & Supply Voltages: the Daniell-Cell Hypothesis].—Cosens: Forrest. (*Wireless World*, Sept. & Oct. 1943, Vol. 49, Nos. 9 & 10, pp. 274 & 302.) *See* also "Diallist," pp. 280–281 & 316.

645. Wireless "Jargon": Avoiding Misleading and Unintelligible Terms [Editorial].—(*Wireless World*, Oct. 1943, Vol. 49, No. 10, p. 287.)

646. Definition of Electronics [deals with Electrons "which are Free in the Sense of being Substantially at Much Greater Distances from the Nuclei of Atoms than the Radii of the Outermost Stable Orbits of the Normal Atom": also a Simpler Definition].—Slepian. (*Electronics*, May 1943, Vol. 16, No. 5, pp. 158 & 160: summary only.)

647. Transliteration Once More [Criticisms of Dunlap's Views].—Furfey. (*Science*, 13th Aug. 1943, Vol. 98, No. 2537, p. 153.) Continuation of the correspondence referred to in 317 of January. "For every language written with a non-Roman alphabet there exists a system of transliteration which is accepted as more or less standard by specialists in the field. . . ." For a letter on the still worse confusion in the transliteration of English names into Russian, *see* issue for 3rd Sept. 1943, No. 2540, p. 219.

648. The Coordination of Abstracting.—(*Engineering*, 8th Oct. 1943, Vol. 156, No. 4056, p. 292.) Nathan's proposals for international cooperation: need for national scheme first: the assumption that "an abstract of an article is a specific thing, like its title," is false when specialisation is involved: the recent A.S.L.I.B. discussion: Rosenhain's "evaluated abstracts" scheme ("it would appear that a system of this kind would be best operated by abstracting bureaux each covering a special field. . . . This is not to say that some type of central office, constituted by the individual services, might not be of value. It might prevent some duplication, although it could not do so completely, and it is not desirable that it should").

649. B.K. Microfilm Copying Camera [New Design with Copy placed Face Downwards on Glass Plate about 3 ft. above Camera. Various Advantages, including Absence of Focusing].—Wray Optical Works. (*Journ. of Scient. Instr.*, Nov. 1943, Vol. 20, No. 11, p. 183.) Designed by B. K. Johnson of Imperial College.

650. Radio Founders' Day, 1942 ["Commemorating the Founders of Radio Science and Radio Industry" on 12th December (the Day of Marconi's "Three Dots" across the Atlantic): Presidential Address, including the Rôle of the Engineer in the Post-War World].—Fisk. (*Proc. Inst. Rad. Eng. Australia*, April 1943, Vol. 5, No. 1, pp. 4–8.)

651. Nikola Tesla's Achievements in the Electrical Art: I—Tesla's Contribution to Electric Power: II—Tesla's Contribution to High Frequency: III—Papers and Lectures: References.—Scott: Wheeler. (*Elec. Engineering*, Aug. 1943, Vol. 62, No. 8, pp. 351–355 and 355–357.) *See* also 2623/4 of 1943, and for Fleming's I.E.E. Commemorative Lecture see *Electrician*, 3rd Dec. 1943, Vol. 131, No. 3418, pp. 563–566.

652. Robert W. Paul: Pioneer Instrument Maker and Cinematographer.—Eccles. (*Electronic Eng:g*, Aug. 1943, Vol. 16, No. 186, pp. 99–102.) For R. W. Paul's gift to Research *see* p. 128.

653. Papers on the Fortieth Anniversary of the Russian Radio Industry.—Rybkin, Dobropistsev, & Vologdin. (*Izvestiya Elektroprom. Slab. Toka*, No. 11, 1940, pp. 1–5, 6–8 and 8–12.) The year 1900 stands out in the history of the Russian

radio industry as the year when the first radio workshop in Russia was organised by A. S. Popov (*see* 3218 of 1943) in Kronstadt to supply the needs of the Russian Navy, in which radio had been used as far back as 1897. This workshop was later enlarged and transferred to St. Petersburg in spite of difficulties encountered with the Laboratory of High Explosives, who were housed in a near-by building and who were alarmed by the prospect of electrical oscillations being induced in their mains network. Fortunately, a special commission set up to investigate the matter was able to allay their anxiety. Although the Russians were thus pioneers in radio-communication, the progress of the radio industry during the subsequent years was very slow in comparison with successes achieved abroad. Except for the above workshop, where independent work was carried out by Russian engineers, nearly the whole of radio production was concentrated in the hands of two firms representing respectively the Telefunken and Marconi companies. The real foundation of modern Russian industry was laid in 1918, when by a decree of Lenin a radio laboratory was established in Nizhni-Novgorod (Gorki); it is interesting to note that up to 1926 Germany was buying high-power valves manufactured in that laboratory.

654. "SCIENCE IN SOVIET RUSSIA" [Book Review].—Needham (Edited by). (*Journ. Roy. Soc. Arts*, 17th Sept. 1943, Vol. 91, No. 4648, p. 564.) Papers by seven British scientists.

655. SCIENCE IN WESTERN SZECHUAN: I—PHYSICO-CHEMICAL SCIENCES AND TECHNOLOGY.—Needham. (*Nature*, 25th Sept. 1943, Vol. 152, No. 3856, pp. 343–345.)

656. THE ADVANCEMENT OF SCIENCE IN CHINA DURING THE PAST THIRTY YEARS.—Liu. (*Science*, 16th July 1943, Vol. 98, No. 2533, pp. 47–51.)

657. INDUSTRIAL RESEARCH AND INDIAN INDUSTRIES [Editorial].—(*Sci. & Culture* [Calcutta], June 1943, Vol. 8, No. 12, pp. 465–466.)

658. "FORWARD WITH SCIENCE" [Book Review].—Rusk. (*Nature*, 25th Sept. 1943, Vol. 152, No. 3856, pp. 341–342.)

659. ELECTRONICS: ITS START FROM THE "EDISON EFFECT" SIXTY YEARS AGO [with Some Newly Uncovered Historical Material].—White. (*Gen. Elec. Review*, Oct. 1943, Vol. 46, No. 10, pp. 537–541.)

660. SIXTY YEARS OF ELECTRONICS, 1883–1943, *and* ELECTRON TUBES FOR INDUSTRY AND COMMUNICATION [Graphic Portrayals].—(*Electronics*, July 1943, Vol. 16, No. 7, between pp. 112 and 113.)

661. JUBILEE OF THE VERBAND DEUTSCHER ELEKTROTECHNIKER ["V.D.E.": Special Number devoted to Papers about Its History and Its Work, Past and Present].—(*E.T.Z.*, 28th Jan. 1943, Vol. 64, No. 3/4, pp. 29–68.) *See* also No. 7/8 (25th Feb. 1943), pp. 93–98.

662. WESTERN HEMISPHERE I.T. & T. SYSTEM COMMUNICATION CONTRIBUTIONS OF 1942.—(*Elec. Communication*, No. 2, Vol. 21, 1943, pp. 75–83.)

663. NEW MURRAY HILL LABORATORY OF BELL TELEPHONE LABORATORIES.—Hunt. (*Journ. Applied Phys.*, June 1943, Vol. 14, No. 6, pp. 249–257.) *See* also 503, above.

664. RADIO AND ELECTRONICS IN THE NAVY [Then and Now].—Sashoff. (*Electronics*, April 1943, Vol. 16, No. 4, pp. 72–75.)

665. CONCERNING MILITARY RADIO DEVELOPMENTS AND THE AMATEUR [Editorial].—(*QST*, Sept. 1943, Vol. 27, No. 9, pp. 7–8.) "When the green light for post-war operation is flashed we need not expect that the closed doors of military secrecy will open, revealing brand-new developments on which we can capitalize . . ."

666. BASIC FAULT-FINDING: METHODS OF TRAINING ARMY SERVICE-MEN [at Loughborough College].—Wilkinson. (*Wireless World*, Dec. 1943, Vol. 49, No. 12, pp. 359–361.) Latest developments of 3795 of 1942.

667. THE RESTORATION OF TELEPHONE APPARATUS AFFECTED BY ENEMY ACTION [mechanically Undamaged, but Useless from Action of Dirt or Water: the Use of "Dekalin" Spray, etc.]—Richards & Glover. (*P.O. Elec. Eng. Journ.*, April 1943, Vol. 36, Part 1, pp. 25–27.) *See* also Oct. issue, Part 3, p. 94.

668. REORGANISING BROADCASTING: SOME DRASTIC PROPOSALS EXAMINED.—(*See* 616.)

669. COURT DECISIONS AFFECTING BROADCASTING.—Parker. (*Electronics*, June 1943, Vol. 16, No. 6, pp. 117–118 and 268..272.)

670. A REPORT ON THE N.A.B. WAR CONFERENCE [April 1943].—Winner. (*Communications*, May 1943, Vol. 23, No. 5, pp. 18–20.)

671. REGULATIONS CONCERNING THE SAFETY AND PROTECTION OF ELECTRICAL APPARATUS FOR THE TRANSMISSION AND REPRODUCTION OF SOUND AND IMAGES AND OF APPARATUS FOR TELECOMMUNICATION AND TELECONTROL.—(*Bull. Assoc. Suisse des Élec.*, 11th Aug. 1943, Vol. 34, No. 16, pp. 491–494: in French.)

672. SYMPOSIUM ON MUSIC IN INDUSTRY.—Halpin & others. (*See* 491.)

673. "MYE TECHNICAL MANUAL" [Book Review].—Mallory & Company. (*Electronics*, June 1943, Vol. 16, No. 6, p. 343.)

674. FREQUENCY MODULATION FOR POWER-LINE CARRIER CURRENT.—Kenefake. (*See* 607.)

675. PUMPS CONTROLLED AUTOMATICALLY [to maintain Level within One Foot of Any Predetermined Depth] OVER 'PHONE LINE TO RESERVOIR.—Norman. (*Electronics*, April 1943, Vol. 16, No. 4, pp. 148..152.) At Kalamazoo, Michigan.

676. THE TELEMETERING OF TEMPERATURE AND VACUUM VALUES FOR THE REMOTE CONTROL OF UNATTENDED SUBSTATIONS [by the Compensation-Amplifier & Impulse-Frequency Methods].—Klinker. (*Elektrot. u. Masch:bau*, 30th April 1943, Vol. 61, No. 17/18, pp. 202–204.)

677. GAS-TUBE HARMONIC GENERATORS [giving Current Pulses useful in High-Speed Triggering Devices, etc.].—Kersta. (*See* 539.)

678. ON SOME PHENOMENA OF PARAMETRIC EXCITATION IN PHYSICS AND TECHNICS [and the Advantages of "Parametric" Machines].—Papalexi. (*See* 430.)

679. What is the Amplidyne? How Does It Work?—Felix. (*Gen. Elec. Review*, Aug. 1943, Vol. 46, No. 8, pp. 442–445.) For this device *see* also 1206 & 2928 of 1942 and back references.

680. Fundamental Principles of Amplidyne Applications.—Crever. (*Elec. Engineering*, Sept. 1943, Vol. 62, No. 9, Transactions pp. 603–606.)

681. The Dependence on Load of Automatic Control Systems, and Its Elimination.—Weis. (*E.T.Z.*, 20th May 1943, Vol. 64, No. 19/20, pp. 261–267.)

682. Electrical Measuring Methods for Automatic-Regulation Technique [Survey, including Various Types of Converters (Thermoelectric, Inductive, Chemical, Magnetoelastic, Resistive, Bolometric, & Valve Devices)]. — Dahnken. (*E.T.Z.*, 17th June 1943, Vol. 64, No. 23/24, pp. 319–323: to be concluded.)

683. Motors do a Better Job [Advantages & Industrial Applications of Adjustable-Speed Electronic Motor Drive: the Thy-mo-Trol & Mot-O-Trol Systems].—General Electric: Westinghouse. (*Scient. American*, Oct. 1943, Vol. 169, No. 4, pp. 166–168.) *See*, for example, 340/2 of January.

684. A Two-System Meter for Process Monitoring and Control [with Two Pointers crossing at Right Angles, giving Ordinate & Abscissa Values].—Laub. (*Elektrot. u. Masch:bau*, 17th Jan. 1941, Vol. 59, No. 3/4, pp. 44–45.)

685. Resonant Electrical Control System [Defects of Existing Positioning & Control Equipments in Industrial & Aircraft Installations: System depending on Phase-Shifting Characteristics of Series-Resonant Circuit: applicable to Single- & Multi-Channel Radio Control: Control is Continuous, Not Step-by-Step: etc.].—Moore. (*Elec. Engineering*, Oct. 1943, Vol. 62, No. 10, pp. 436–439.)

686. Resistance-Welding Controls: New Electronic Controls are Contributing to the War Effort.—Palmer. (*Gen. Elec. Review*, Sept. 1943, Vol. 46, No. 9, pp. 507–510.)

687. A New Method for the Measurement of the Friction Coefficients μ_0 and μ for Sliding Friction [based on the Magneto-Mechanical System dealt with in 3827 of 1942 (Relay & Control Technique)].—Piesker. (*Zeitschr. f. tech. Phys.*, No. 2, Vol. 24, 1943, pp. 34–38.)

688. Electronics applied to Heat-Transfer Tests.—Brown. (*Electronics*, July 1943, Vol. 16, No. 7, pp. 113–115.)

689. Heat-Transfer Problems solved with Roomful of *R-C* Networks ["Heat & Mass Flow Analyser" at Columbia University, New York City].—Paschkis. (*Electronics*, April 1943, Vol. 16, No. 4, pp. 180..183.) *See* also 1977 of 1942.

690. Analysing Heat Flow in Cyclic Furnace Operation [by Use of Electrical Network with Resistors representing Thermal Conductivity and Capacitors representing Thermal Lag].—Bradley & Ernst. (*Sci. Abstracts*, Sec. B, June 1943, Vol. 46, No. 546, p. 109.)

691. Supersonics in Biology [Researches with Concentrated Beams].—Lynn & others. (*See* 514.)

692. Detection of Flaws in Plates by Supersonic Waves.—Trost. (*See* 515.)

693. Data on New Acoustic Stethoscope.—Olson. (*See* 472.)

694. British Electrocardiograph uses Smoked-Glass Discs [Cox-Both Portable Three-Valve Instrument giving Trace One-Fortieth Conventional Size, viewed in Standard Size, during Recording, through Microscope].—(*Electronics*, June 1943, Vol. 16, No. 6, p. 264.) From *Wireless World*.

695. Electroencephalograph Design [Fluctuation Noise and Microphonics: Effect of Input-Impedance Values on Signal/Noise Ratio: Calibration Problems: Frequency Response: Interference (including the Tonnies Circuit and Its Limitations): a New Portable Instrument for Use without Shielded Room].—Traugott. (*Electronics*, Aug. 1943, Vol. 16, No. 8, pp. 132..144.)

696. Encephalaphone converts Brain Potentials into Sounds—Beevers & Fürth. (*Electronics*, May 1943, Vol. 16, No. 5, pp. 144..148.) From the paper dealt with in 2041 of 1943 and back reference.

697. Cathode-Ray Equipment tests Paralysed Muscles to determine whether They can be Restored.—(*Electronics*, Aug. 1943, Vol. 16, No. 8, p. 170: photograph & caption only.) Apparently by giving the muscle-potential traces.

698. Recording Blood Flow in Stomach with Thermocouples and Phototube [Technique and Some Results].—Richards & others. (*Electronics*, July 1943, Vol. 16, No. 7, pp. 190..194.) Cornell University Medical College researches.

699. Model Researches on the Theory of the Excitation of Biological Objects [and the Close Analogies between the Spontaneous & Impulse-Excited Discharges of Electrical Relaxation-Oscillation Systems and the Automatic & Stimulated Excitation of Nerve, Muscle, etc.].—Bethe. (*Naturwiss.*, 4th June 1943, Vol. 31, No. 23/24, pp. 276–277.)

700. On the Theory of the Electrical Stimulus [and Nernst's Threshold Law for the Excitation of Nerves, etc.].—Bonhoeffer. (*Naturwiss.*, 4th June 1943, Vol. 31, No. 23/24, pp. 270–275.) *See* also Bethe, 699, above.

701. An Electronic Stimulator for Use in General Physiology [developed from Those of Connerty & Johnson and Delaunois, and overcoming Certain Defects].—Thorp & Robinson. (*Electronic Eng:g*, Oct. 1943, Vol. 16, No. 188, pp. 188–190.)

702. New Researches on Electric Fish.—Cox. (*Science*, 21st May 1943, Vol. 97, No. 2525, Supp. p. 10.) *See* also *Sci. Abstracts*, Sec. A, July 1943, Vol. 46, No. 547, p. 152.

703. Acetylcholine and the Physiology of the Nervous System [including the Processes in the Electric Organs of *Torpedo, Ray*, etc.].—Fulton & Nachmansohn. (*Science*, 25th June 1943, Vol. 97, No. 2530, pp. 569–571.) "The new concept removes the chief difficulty for conciliating the 'electrical' and 'chemical' theories of transmission of nerve impulses."

704. ULTRA-HIGH-FREQUENCY THERAPY [Survey from Electrotechnical Viewpoint].—Owen. (*Proc. Inst. Rad. Eng. Australia*, May 1942, Vol. 4, No. 8, pp. 134–143.)

705. HIGH-FREQUENCY THERAPY: PART I—GENERAL INTRODUCTION: PART II—DIELECTRIC PHENOMENA AT HIGH FREQUENCIES.—Oliphant. (*Electronic Eng:g*, Sept. & Oct. 1943, Vol. 16, Nos. 187 & 188, pp. 142–144 & 206–208: to be contd.)

706. DOSIMETRY AND LOCAL DISTRIBUTION OF ENERGY IN THE ELECTRIC HIGH-FREQUENCY FIELD [of Interest in Biological & Medical Work and also in Insulation Research: Objections to Previous Dosimeter Methods: the Use of the Electrodeless-Discharge Measuring Instrument (2770 of 1941 and 3665 of 1942): Examples of Short-Wave Field Distribution].—Lion. (*Journ. Applied Phys.*, Oct. 1943, Vol. 14, No. 10, pp. 545–551.)

"The material presented up to date does not allow any conclusion to be drawn about the effect of time, neither for the stimulation nor for the lethal effect. It seems, however, indicated that the assumption of a simple law of the form: Dose = Intensity × Time, is not justified." Regarding insulation research, "it may be interesting to note that a major effort of this research is devoted to the development of new insulation materials with low losses, while the reduction of losses by suitable shaping of the electric fields has been of relatively little interest."

707. ELECTRONIC TUBES FOR ULTRA-VIOLET RADIATION, *and* "ULTRAVIOLETTE STRAHLEN . . ." [Generation, Measurement, and Application in Medicine, Biology, and Technics: Book Review].—Laub: Meyer & Seitz. (*Electronics*, May 1943, Vol. 16, No. 5, pp. 80–85 and 138..142: *Gerlands Beiträge z. Geophysik*, No. 3/4, Vol. 59, 1943, pp. 380–381.)

708. COLOUR MICROSCOPY IN ULTRA-VIOLET RAYS [Photographic and Visual Methods].—Brumberg. (*Nature*, 25th Sept. 1943, Vol. 152, No. 3856, p. 357.)

709. "AN INTRODUCTION TO BIOPHYSICS" [Book Review].—Stuhlman. (*See* 498.)

710. CASE-HARDENING WITH RADIO-FREQUENCY ENERGY. —Sherman. (*Electronics*, Aug. 1943, Vol. 16, No. 8, pp. 170 and 172.) From the paper dealt with in 3223 of 1943.

711. ELECTRONIC DEHYDRATION: REDUCTION IN PROCESSING TIME AND INCREASED VITAMIN CONTENT.—Federal Tel. & Rad. Corporation. (*Electrician*, 3rd Dec. 1943, Vol. 131, No. 3418, p. 554.) *See* also 3224 of 1943 and 363 of January.

712. OIL MIST PRECIPITATED: ELECTRICAL CHARGING REMOVES FIRE HAZARD IN SHOPS [Use of "Precipitron" Electronic Tube].—Westinghouse. (*Scient. American*, Oct. 1943, Vol. 169, No. 4, p. 168.)

713. A COUNTING METHOD FOR THE DETERMINATION OF SMALL AMOUNTS OF RADIUM AND OF RADON [with Scaling Circuit (using Twin-Triode Valves), Strobotron Circuit for actuating Traffic-Recorder Counting Dials, etc.].—Curtiss & Davis. (*Journ. of Res. of Nat. Bur. of Stds.*, Sept. 1943, Vol. 31, No. 3, pp. 181–195.)

714. A MICA-WINDOW GEIGER COUNTER TUBE FOR MEASURING SOFT RADIATIONS.—Copp & Greenberg. (*Review Scient. Instr.*, July 1943, Vol. 14, No. 7, pp. 205–206.) *See* also H. Weltin, Sept. issue, No. 9, p. 278.

715. A COMBINED VOLTAGE-REGULATING AND QUENCHING CIRCUIT FOR THE GEIGER-MÜLLER COUNTER.—Bale & Bonner. (*Review Scient. Instr.*, July 1943, Vol. 14, No. 7, pp. 222–223.)

716. FACTORS INFLUENCING THE PLATEAU CHARACTERISTICS OF SELF-QUENCHING GEIGER-MÜLLER COUNTERS.—Spatz. (*Phys. Review*, 1st/15th June 1943, Vol. 63, No. 11/12, p. 462: summary only.)

717. AN EXPERIMENTAL MASS SPECTROMETER [specially designed for investigating Different Source Arrangements: a New Ion Source].—Coggeshall & Jordan. (*Review Scient. Instr.*, May 1943, Vol. 14, No. 5, pp. 125–129.)

718. MASS SPECTROMETER ADAPTED TO ITS WAR JOB OF ANALYSING THE GASES FORMED IN MAKING BUTADIENE.—Hipple. (*Science*, 20th Aug. 1943, Vol. 98, No. 2538, Supp. pp. 8–9.) *See* 2280 of 1943.

719. RADIOACTIVE MERCURY IN MERCURY-VAPOUR MEASUREMENTS [Detection of Mercury in Quantities of 10^{-8} g or Less].—Irvine & Goodman. (*Journ. Applied Phys.*, Sept. 1943, Vol. 14, No. 9, pp. 496–498.)

720. RECENT APPLICATIONS OF THE QUARTZ AND X-RAY SPECTROGRAPHS TO POST OFFICE PROBLEMS.—Taylor. (*P.O. Elec. Eng. Printed Paper No. 177*, pp. 1–10.)

721. X-RAY DIFFRACTION MEASUREMENTS WITH GEIGER-MÜLLER TUBE [as Rapid Indicator of Diffraction Maxima and Their History during Reactions].—Pepinsky & Weisz. (*Phys. Review*, 1st/15th June 1943, Vol. 63, No. 11/12, p. 457: summary only.) *See* also 1352 of 1943.

722. "X-RAYS IN RESEARCH AND INDUSTRY" [Book Review].—Hirst. (*Nature*, 7th Aug. 1943, Vol. 152, No. 3849, pp. 145–146.)

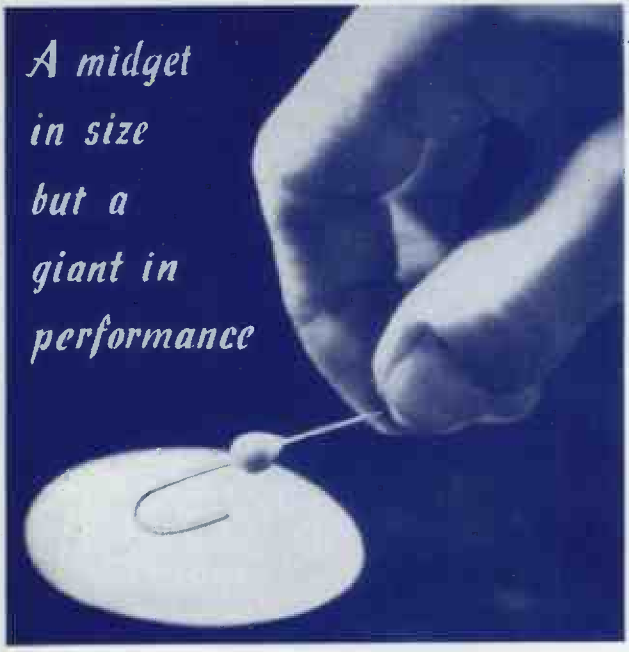
723. "INDUSTRIAL RADIOLOGY" [Second Edition: Book Review].—St. John & Isenburger. (*Science*, 23rd July 1943, Vol. 98, No. 2534, p. 86.)

724. A NEW SCHOOL OF INDUSTRIAL RADIOGRAPHY.—Kodak, Ltd. (*Electronic Eng:g*, July 1943, Vol. 16, No. 185, pp. 58–59.)

725. SURVEY OF PROBLEMS IN X-RAY CINEMATOGRAPHY. —(*Electronics*, May 1943, Vol. 16, No. 5, pp. 166..170.)

726. ON THE OUTPUT OF X-RAYS FOR VERY HIGH VOLTAGE EQUIPMENTS [Calculated Outputs for Voltages up to 8×10^6 V, for Lead & Tungsten Anticathodes: with Discussion].—Dänzer. (*Ann. der Phys.*, 21st July 1943, Vol. 43, No. 3, pp. 182–186.)

727. EMISSION-TYPE PHOTOELECTRIC CELLS [Characteristics, Uses, & Circuits: Possible Future Development & Applications].—Lynch & Tillman. (*See* 521.)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