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THE JOURNAL OF RADIO RESEARCH & PROGRESS

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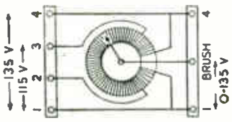

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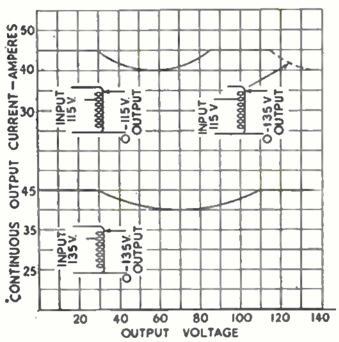

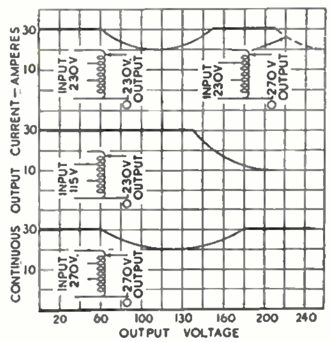

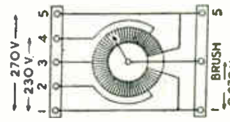

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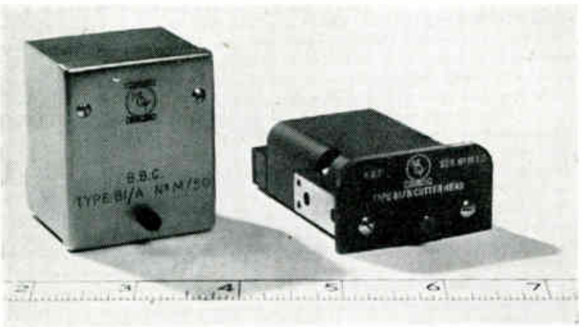

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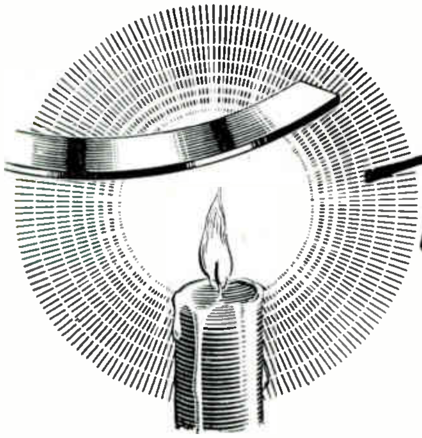

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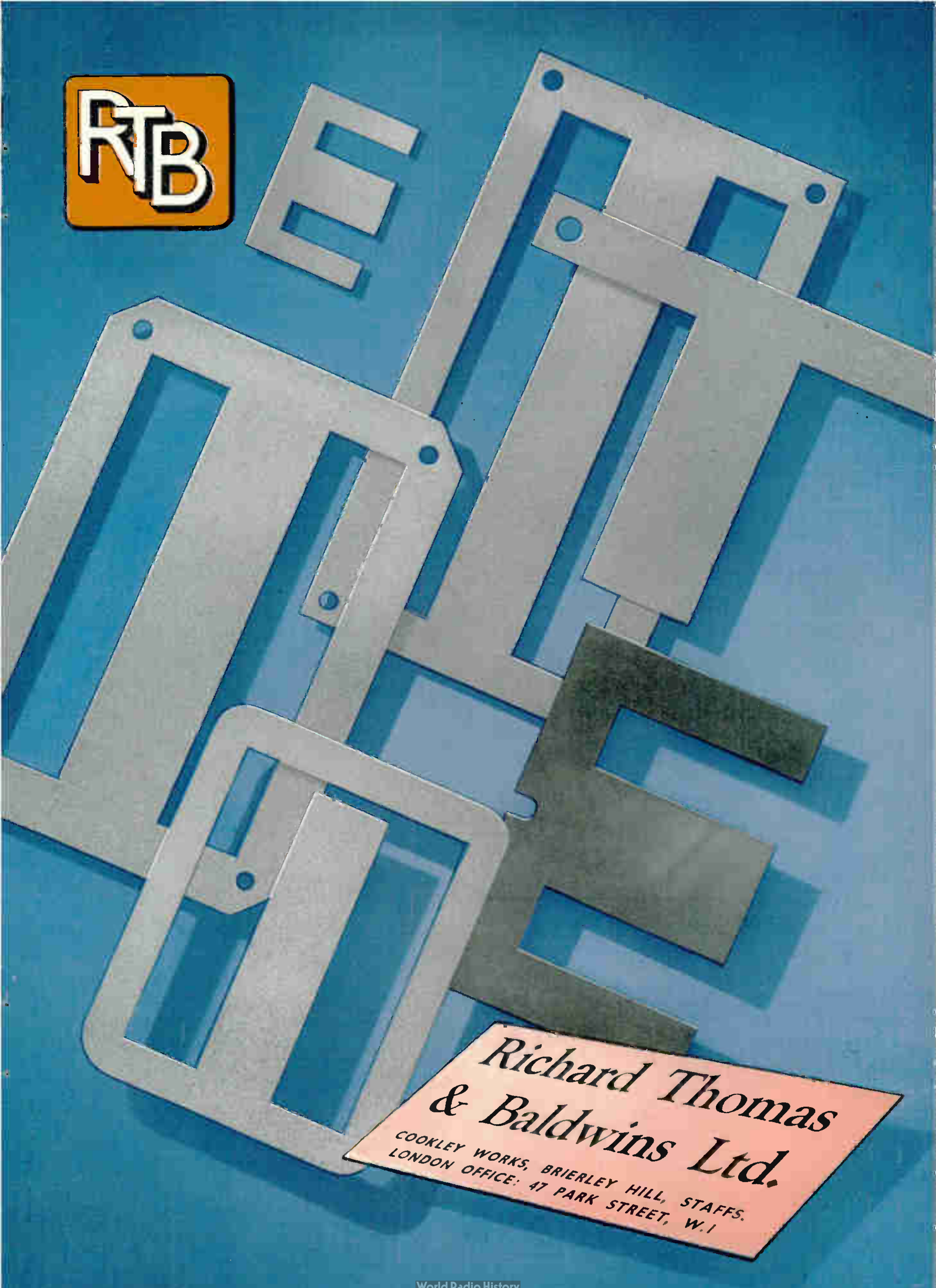
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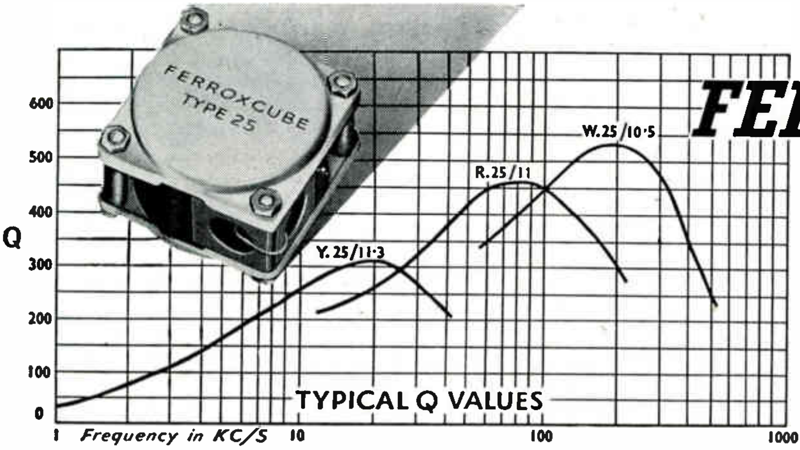

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WIRELESS ENGINEER

The Journal of Radio Research and Progress

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FEBRUARY 1952

Vol. XXIX No. 341

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Published on the sixth of each month

Annual Subscription: Home and Overseas, one year £2 4s. 6d., six months, £1 2s. 3d. Canada and U.S.A. $7.00

Editorial, Advertising and Publishing Offices

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WIRELESS ENGINEER

Vol. XXIX FEBRUARY 1952 No. 341

Cavity Resonator regarded as a Transmission Line

IN last month's Editorial we remarked that one of the diagrams indicated that the cavity resonator of the simple pill-box type could be regarded as a number of half-wave aerials connected in parallel. A more usual and useful line of approach is to regard it as a transmission line radiating from the centre in every direction, so that the two conductors of the ordinary transmission line become two parallel planes. One can picture the top and bottom discs divided into a number of segments, so that one has a number of lines connected in parallel, each of which is as shown in Fig. 1. On arriving at the edge of the cavity, that is, when the distance x is equal to the radius a, the line is terminated; neglecting the resistance of the wall, the line is short-circuited. We therefore have a line open-circuited at one end and short-circuited at the other, but instead of the inductance and capacitance per unit length being constant, L per unit length $= l/x$ and C per unit length $= cx$ where l and c are constants. If i and v are the current and voltage at the distance x at any moment, then

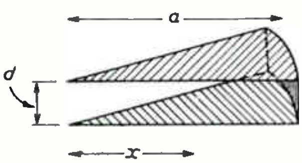

Fig. 1.

$$\frac{dv}{dx} = L\frac{di}{dt} = \frac{l}{x}\cdot\frac{di}{dt} \qquad (1)$$

and

$$\frac{di}{dx} = C\frac{dv}{dt} = cx\cdot\frac{dv}{dt} \qquad (2)$$

From (1)

$$\frac{d^2v}{dx^2} = -\frac{l}{x^2}\cdot\frac{di}{dt} + \frac{l}{x}\cdot\frac{d}{dx}\left(\frac{di}{dt}\right) \qquad (3)$$

and from (2)

$$\frac{d}{dx}\left(\frac{di}{dt}\right) = \frac{d}{dt}\left(\frac{di}{dx}\right) = cx\frac{d^2v}{dt^2} \qquad (4)$$

Hence

$$\frac{d^2v}{dx^2} = -\frac{l}{x^2}\cdot\frac{x}{l}\cdot\frac{dv}{dx} + \frac{l}{x}\cdot cx\cdot\frac{d^2v}{dt^2} \qquad (5)$$

and since, if v is varying sinusoidally with regard to time, $d^2v/dt^2 = -\omega^2 v$ (5) may be written

$$\frac{d^2v}{dx^2} + \frac{1}{x}\cdot\frac{dv}{dx} + k^2 v = 0 \qquad (6)$$

where $k^2 = lc\omega^2$.

This differs from the ordinary transmission-line formula by the inclusion of the second term, which is due to the variation with x of the inductance and capacitance. We are assuming that resistance and leakance are negligibly small.

The solution of this equation is given by the simplest form of cylindrical or Bessel function, viz.

$$v = A\cdot J_0(kx)$$

$$= A\left[1 - \frac{k^2x^2}{2^2} + \frac{k^4x^4}{2^24^2} - \frac{k^6x^6}{2^24^26^2} + \ldots\right]$$

where A is a given constant; viz., the value of v when $x = 0$.

At the risk of boring our more mathematical readers we may prove the correctness of the solution as follows:—Putting $A = 1$,

$$\frac{d^2v}{dx^2} = -\frac{k^2}{2} + \frac{3}{2^2 4} k^4x^2 - \frac{5}{2^2 4^2 6} k^6x^4 + \ldots$$

$$\frac{1}{x}\cdot\frac{dv}{dx} = -\frac{k^2}{2} + \frac{1}{2^2 4} k^4x^2 - \frac{1}{2^2 4^2 6} k^6x^4 + \ldots$$

$$k^2v = +k^2 - \frac{1}{2^2} k^4x^2 + \frac{1}{2^2 4^2} k^6x^4 - \ldots$$

$$\text{total} = 0 \quad + 0 \quad + 0 \quad + \ldots$$

The values of $J_0(kx)$ for various values of kx are given in tables of Bessel functions (see, for example, Jahnke and Emde's "Tables of Higher Functions," p. 156).

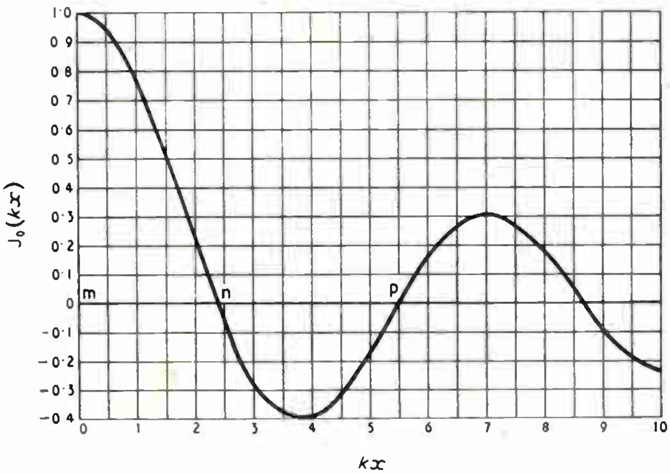

Fig. 2.

The value of lc for the line consisting of the two parallel discs can be found as follows:—

$$H = \frac{4\pi}{10} i/2\pi x = \frac{2i}{10x}$$ where i is the current in amperes and x the distance in cm.

$$l = \frac{\Phi}{10^8 i} = \frac{2d}{10^9 x}$$ henry, where d is the distance between the discs.

Similarly

$$c = \frac{2\pi x}{4\pi d}\cdot\frac{1}{9\times 10^{11}} = \frac{x}{2d}\cdot\frac{1}{9\times 10^{11}}$$ farad

Hence $lc = \dfrac{1}{9\times 10^{20}}$ and $k = \omega\sqrt{(lc)} = \dfrac{\omega}{3\times 10^{10}}$

Taking for example a frequency of 4,000 Mc/s, $k = \dfrac{2\pi \times 4\times 10^9}{3\times 10^{10}} = 0{\cdot}8375$; applying this to the various values of x in the table, we obtain the values of kx, for which the values of $J_0(kx)$ are obtained from the Tables of Bessel Functions.

$f = 4{,}000$ Mc/s

x (cm)	kx	$J_0(kx)$
0	0	1·0
1·815	1·52	0·5
2·86	2·4	0
4·575	3·83	− 0·4
6·6	5·52	0
8·375	7·015	0·3
10·3	8·65	0
12·25	10·275	− 0·25

The values of x have been chosen to give maximum and zero values of $J_0(kx)$. In Fig. 2 the values of $J_0(kx)$ are plotted against the values of kx, so that the curve can be used for any frequency by inserting the appropriate value of k.

In an ordinary line with constant inductance and capacitance per unit length this would be a sine curve, and the distance mp would be exactly three times mn. In the present case $mn/mp = 2{\cdot}4/5{\cdot}52 = 0{\cdot}435$, so that, if the frequency is adjusted until the cavity is resonating at its 'third harmonic,' the distance from the centre to the voltage node is not a third of the radius as one might expect, but is equal to 0·435 a. At this point the current and the magnetic field have their maximum values. Beljers* made use of this in experiments on the gyromagnetic resonance of five spheres of 'Ferroxcube 4'; the spheres were only 0·2 mm diameter and were supported by a thin cylinder of low-loss dielectric material with a radius of 0·435 of that of the cavity. The spheres were therefore situated in the resonant magnetic field; for the gyromagnetic tests they were also in an externally applied steady magnetic field.

Although we have referred to the 'third harmonic,' it should be specially noticed that its frequency is not three times that of the fundamental. From Fig. 2 it is seen that for the fundamental resonance mn/k_1 must be equal to the cavity radius, whereas for the 'third harmonic' mp/k_3 must be equal to the cavity radius; hence

$$\frac{k_3}{k_1} = \frac{\omega_3}{\omega_1} = \frac{mp}{mn} = \frac{5{\cdot}52}{2{\cdot}4} = 2{\cdot}3,$$ and if the cavity has a fundamental resonant frequency of 4,000 Mc/s, the frequency of its 'third harmonic' is not 12,000 but 9,200 Mc/s. In the experiments referred to above the frequency was 9,250 Mc/s.

G. W. O. H.

* H. G. Beljers, *Physica*, Vol. 14, pp. 629–641, 1949. See also H. B. G. Casimir, *Philips Research Report*, Vol. 6, pp. 162–182, June 1951.

RESISTORS AT RADIO FREQUENCY

Characteristics of Composition Type

By T. J. F. Pavlasek, M.Eng., M.I.R.E. and F. S. Howes, M.Sc., Ph.D., F.I.R.E.

(Department of Electrical Engineering, McGill University, Montreal, Canada)

SUMMARY.—The characteristics of JAN-R-11 type, fixed composition resistors ranging from 1000 ohms to 2 megohms are investigated over a range of frequencies from 500 kc/s to 40 Mc/s. The radio-frequency resistance and equivalent shunt capacitance are measured for a group of approximately 45 samples, 15 from each of the types RC 20, RC 30 and RC 41. Curves of equivalent impedance and phase angle, useful for design purposes are shown. 'Universal curves' of $\frac{R_{rf}}{R_{dc}}$, $\frac{|Z|}{R_{dc}}$ and ϕ vs fR_{dc} are plotted and a comparison made with G. W. O. Howe's theoretical curve.

1. Previous Investigations

When a high-value fixed composition resistor, either the solid-rod type or one consisting of a conductive film deposited on a glass or porcelain tube, is measured at radio frequencies, it is found that the resistance is considerably less than that obtained by direct-current measurements. This phenomenon was first described by Boella[1] and was originally referred to as the 'Boella Effect'. The effect was further studied and corroborated by Sowerby and Marshall and described by Puckle.[2]

A theoretical explanation of this phenomenon was attempted by Howe[3] who treated the resistor as a transmission line with distributed capacitive and resistive circuit elements. Howe assumed that the resistor has an average distributed capacitance per unit length. The type of transmission line assumed is that indicated in Fig. 1 and its behaviour may be calculated from transmission-line theory.

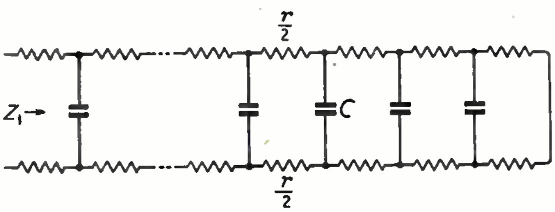

Fig. 1. *Transmission-line equivalent of a resistor.*

The input impedance of a short-circuited transmission line is given by

$$Z_1 = Z_0 \tanh \gamma l$$

where Z_0 is the characteristic impedance,
γ is the propagation constant and
l is the length of the line (in this case, one-half of the length of the resistor).

As applied to the resistor, this formula reduces to

$$Z_1 = R_{dc}\frac{\tanh \theta}{\theta} = |Z_1| \underline{/\phi}$$

where $\theta = l\sqrt{\omega C r}\ \underline{/45^\circ}$

R_{dc} is the total d.c. resistance
r is the resistance per unit length and
C is the average distributed capacitance per unit length.

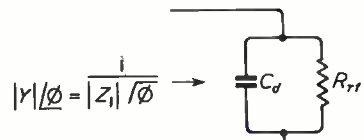

Fig. 2. *Equivalent circuit of a resistor, valid at a particular frequency.*

At any particular frequency, Z_1 may be expressed as an equivalent resistance R_{rf} in parallel with an equivalent shunt capacitance C_d. This is shown in Fig. 2. Then $R_{rf} = \frac{|Z_1|}{\cos\phi}$ and $C_d = \frac{\sin\phi}{\omega|Z_1|}$. Thus if the value of C, the average distributed capacitance per unit length can be determined, the effective resistance R_{rf} can be calculated for that frequency.

If the ratio $\frac{R_{rf}}{R_{dc}}$ is plotted against the parameter $flCR_{dc}$ where f is the frequency, then a normalized curve is obtained from which the effective resistance R_{rf} may be obtained for any frequency.

Howe attempted to calculate C from the physical dimensions of the resistor by assuming the distribution of charge along the surface of the resistor to be proportional to the distance from the midpoint. The potential distribution along the surface of the resistor, and thus the distributed capacitance, were then obtained.

The treatment of the resistor as a uniform transmission line may be regarded as a first

MS accepted by the Editor, August 1950

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approximation. However, the assumption of an average distributed capacitance C and the methods of calculation used do not give predictable results. Nevertheless other investigators have shown that the concept is applicable when the universal curve of $\frac{R_{rf}}{R_{dc}}$ vs $flCR_{dc}$ can be obtained by actual measurement. Thus, if the R_{rf} of a number of different resistors is measured over a range of frequencies, the experimental values may be plotted in such a form and the results compared with the theoretical curve.

Measurements of this nature were made at about the same time by Hartshorn and Ward,[4] Bressi[5] and Pontecorvo.[6] These measurements show a certain degree of correspondence with the theoretical curve, but a difficulty was encountered in calculating the parameter $flCR_{dc}$ because of the uncertainty of evaluating C.

Pontecorvo pointed out, however, that if the term lC is omitted and $\frac{R_{rf}}{R_{dc}}$ is plotted against the parameter fR_{dc}, then a family of curves is obtained. When plotted logarithmically these curves all have the same shape but are displaced relatively to one another along the abcissa. A curve may thus be found for a given type and physical size of resistor. The effective resistance, at any frequency, of a resistor of the same type may then be found from such a curve.

Measurements of this nature made by Pontecorvo, Hartshorn, Essen[8] and Simmonds[9] indicate that such curves may be obtained experimentally. Measurements by Miller and Salzberg[7] and by Ney[10] on resistors below 100,000 ohms, do not give any conclusive results.

The original measurements by Boella, Sowerby and Marshall were obtained by modified resistance variation methods. The results of Hartshorn and Ward were obtained by a susceptance variation method (Sinclair[11]), and those of Bressi and Pontecorvo by a dynamic negative resistance method. The later measurements of Miller and Salzberg, Essen and Simmonds utilize transmission-line techniques (Chipman[12]). The measurements by Ney were made with a Schering bridge at frequencies up to several hundred kilocycles per second.

2. Present Investigations

2.1. *Method of Measurement*

The measurements performed during this investigation were made by means of a parallel-T null circuit. The instrument used was a General Radio Co. Twin-T Impedance Measuring Network, Type 821 A.

The operating range of the Twin-T is from 0·46 to 30 Mc/s and may be extended with precautions to 40 Mc/s. Although this instrument does not cover as wide a frequency range as might be desired for a general study of the problem, the range was nevertheless considered sufficient for the purpose, especially in view of the other advantages offered.

The range of resistances to be measured (1000 ohms to 2 megohms) is covered satisfactorily by the Twin-T. The accuracy is reduced, however, for the highest resistances at the upper frequency limit because the readings are in the vicinity of the zero of the dial which is calibrated in terms of conductance.

At frequencies up to 2 Mc/s, the resistances below 10,000 ohms lie outside the range of the instrument. However, the change in resistance for these resistor values at 2 Mc/s is quite small and the deficiency is therefore not serious.

The theory of the parallel-T circuit and of the Twin-T impedance measuring network is fully discussed in the literature (Tuttle[13] and Sinclair[14]). The basic circuit is shown in Fig. 3. The susceptance component of the unknown admittance is measured by a direct parallel compensation in terms of the variable capacitor C_B, while the conductance component is measured in terms of the change of capacitor C_G. The accuracy of the instrument is thus limited by the calibration and reading accuracy of C_B and C_G and by the effect at high frequencies of residual parameters associated with the network components. The most critical of these are those associated with C_B and R and with the leads used to connect the unknown. Corrections for these (Sinclair[14]) can be made to the measured values.

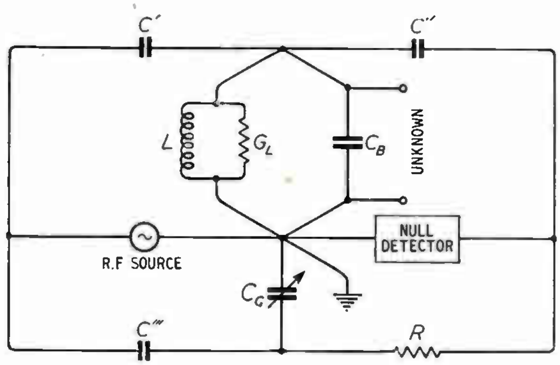


Fig. 3. *Basic circuit of twin-T.*

For an investigation of this type, the parallel-T circuit has the following advantages: The instrument is a null device and, therefore, it is not necessary to have accurate r.f. voltage-indicating instruments such as are required in resonance curve methods. For the same reason, the voltage stability of the signal source is not critical. The measurement of both the conductance and susceptance components is in terms of variable capacitors whose calibration is capable of high accuracy and is relatively independent of frequency. Again, the nature of the circuit is such that the source, the bridge circuit, the unknown and the detector all have a common earth point. This facilitates shielding and results in most of the residual capacitances of the components having no effect on the balance.

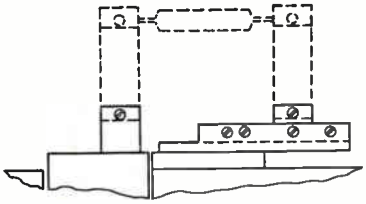


Fig. 4. *Resistor mounting jig.*

2.2. *Arrangement of Apparatus and Method of Operation*

The apparatus was installed in a shielded room to insure freedom from interfering signals and noise. All of the equipment was mounted on a large copper sheet to which each instrument was securely earthed. A voltage-regulating transformer was used to provide power-supply stability.

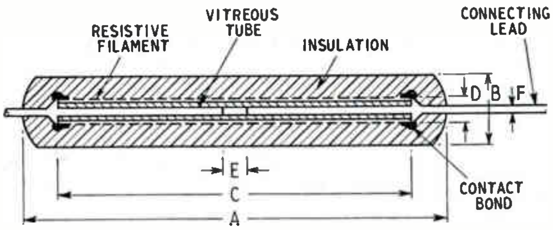
The frequency setting of the signal generator was determined for each measurement by a crystal calibrator whose accuracy had previously been checked in comparison with the U.S. Bureau of Standards' Standard Frequency Transmitter WWV. All connecting leads were shielded to eliminate stray radiation.

The detector was a communications-type radio receiver operated on c.w. The b.f.o. tone was observed simultaneously aurally by headphones and visually on an oscilloscope. It was found that this combination gave a more reliable indication of the null, especially at the higher frequencies when receiver noise became noticeable. The shielding of the receiver case was improved and the connections to the Twin-T were made by a coaxial connector with a coaxial plug at both the Twin-T and at the receiver end.

A jig (Fig. 4) was devised for mounting the resistors in order that all measurements should be made under the same conditions. The jig was made adjustable to allow the mounting of the different physical sizes. The spacing of the mounting posts was chosen to allow a clearance of about $\frac{1}{4}$-in. between each end of the resistor and the posts. Two sets of mounting posts of different lengths were made to investigate their effect on the measured values.

2.3. *Results*

Although a considerable variety of resistors were measured in this investigation, the results for only three types, RC-20, RC-30 and RC-41 are given here. These are all of the same manufacture and are of the metallized-filament type. A total of 10 samples of each resistor was available and two were chosen at random from each group for measurement. Table 1 shows the resistors



	TYPE		
DIM'N	RC-20	RC-30	RC-41
A	$\frac{13}{32}''$	$\frac{23}{32}''$	$1\frac{3}{4}''$
B	$\frac{1}{8}''$	$\frac{1}{4}''$	$\frac{21}{64}''$
C	$\frac{1}{4}''$	$\frac{9}{16}''$	$1\frac{9}{16}''$
D	$\frac{3}{64}''$	$\frac{3}{64}''$	$\frac{1}{16}''$
E	$\frac{1}{32}''$	$\frac{7}{64}''$	$\frac{7}{64}''$
F	·032″	·040″	·036″

Fig. 5. *Cross-section and dimensions of filament-type resistors.*

TABLE 1

JAN-R-11 Type	Resistance (Ω)	JAN-R-11 Type	Resistance (Ω)	JAN-R-11 Type	Resistance (Ω)
RC20BF102K	1000	RC30BF102K	1000	RC41BF102K	1000
472K	4,7000	472K	4,700	472K	4,700
103K	10,000	103K	10,000	103K	10,000
473K	47,000	473K	47,000	473K	47,000
104K	0·10 meg	104K	0·10 meg	104K	0·10 meg
474K	0·47 meg	473K	0·47 meg	474K	0·47 meg
105K	1·0 meg	105K	1·0 meg	105K	1·0 meg
205K	2·0 meg			225K	2·2 meg

studied. The cross-section of these resistor types is shown in Fig. 5.

The results of the measurements are presented in graphical form. The plotted curves represent the averaged values for several measurements on each of the samples listed.

Four groups of curves are given as follows:

Group 1: Fig. 6 (a), (b) and (c), $\frac{R_{rf}}{R_{dc}} \times 100\%$ vs frequency for resistor types RC-20, RC-30 and RC-41 respectively. These show the variation of the ratio $\frac{R_{rf}}{R_{dc}}$ with frequency. Here, R_{rf} is the effective r.f. resistance as shown in Fig. 2. Since the Twin-T measures the conductance G, the resistance R_{rf} is $1/G$. The value of R_{dc} used for each resistor was its actual resistance as measured on a Wheatstone Bridge.

Group 2: Fig. 7 (a), (b) and (c), C_d vs frequency for types RC-20, RC-30 and RC-41 respectively.

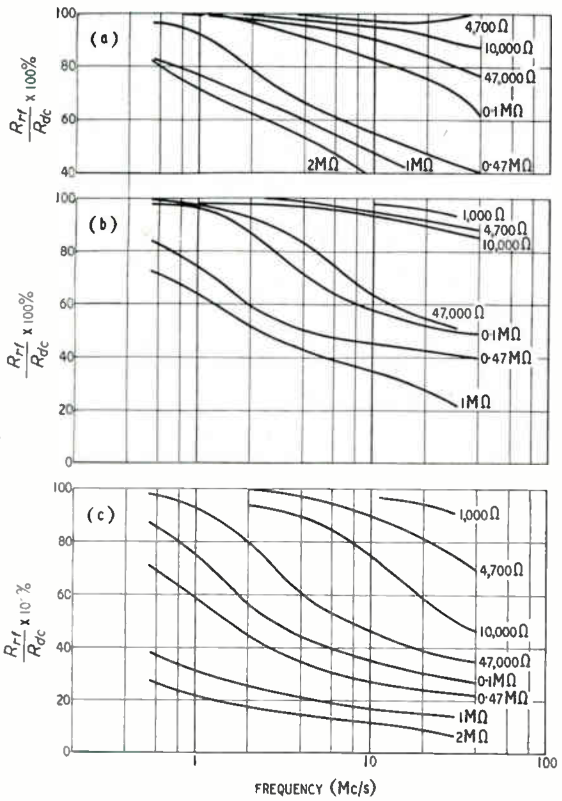


Fig. 6. *R.F. resistance of types RC-20 (a), RC-30 (b) and RC-41 (c) resistors.*

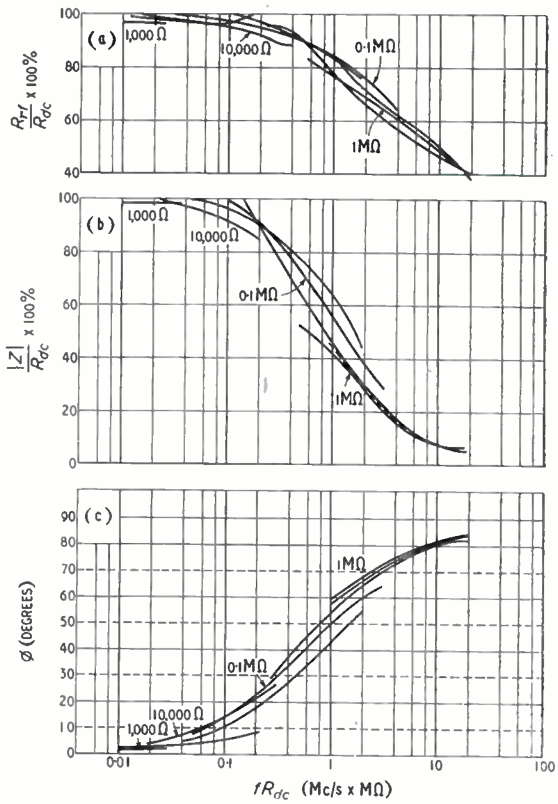


Fig. 8. *R.F. resistance (a), impedance (b) and phase angle (c) of RC-20 resistor.*

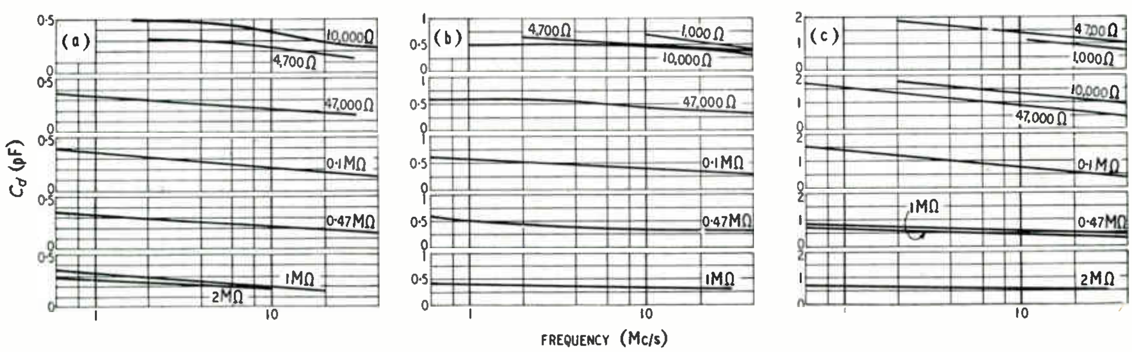


Fig. 7. *Capacitance of types RC-20 (a), RC-30 (b) and RC-41 (c) resistors.*

Group 3: Fig. 8 (a), (b) and (c), $\frac{R_{rf}}{R_{dc}}$, $\frac{|Z|}{R_{dc}}$ and ϕ vs fR_{dc} for type RC-20.

This set of curves has been shown for type RC-20 only to indicate the manner in which the curves of Fig. 9 are obtained. The curves of Fig. 8(a) show the variation of the ratio $\frac{R_{rf}}{R_{dc}}$ with the parameter fR_{dc} and constitute the evidence obtained in this investigation in relation to the theory proposed by G. W. O. Howe.

It is interesting to note that if the values of the magnitude of the complex impedance $|Z|$ and of the phase angle ϕ are plotted in a manner similar to that followed in obtaining the normalized curve $\frac{R_{rf}}{R_{dc}}$ vs fR_{dc} then similar normalized curves are obtained giving $\frac{|Z|}{R_{dc}}$ and ϕ vs fR_{dc}. The manner of plotting such curves for the RC-20 type resistors is shown in Fig. 8 (b) and (c).

Group 4: Fig. 9 (a), (b) and (c), the composite curves of $\frac{R_{rf}}{R_{dc}}$, $\frac{|Z|}{R_{dc}}$ and ϕ vs fR_{dc} for types RC-20, RC-30 and RC-41. In each case, these curves are the mean of a series of curves obtained in the manner shown in Fig. 8 for different values of resistance of a given type.

The values for Figs. 8 and 9 were calculated from the data provided by the curves of Figs. 6 and 7.

The residual parameters of the Twin-T and of the mounting jig were determined and the necessary corrections were applied to the measured values of R_{rf} and C_d. However, it was found that these corrections were insignificant in relation to the probable accuracy of the measurements except in the case of the residual series inductance of capacitor C_B which was found to change the measured values of R_{rf} from 1% to 5% at the frequencies from 20 Mc/s to 40 Mc/s.

2.4. *Observations and Conclusions*

The theory of G. W. O. Howe shows that the ratio $\frac{R_{rf}}{R_{dc}}$ of a resistor is a function of the frequency, the d.c. resistance and the physical dimensions (in particular, the length/diameter ratio). By making the assumption that the resistor may be regarded as a transmission line, with uniformly-distributed resistive and capacitive constants, and that the distributed capacitances are dependent on the physical dimensions alone, a generalized curve is obtained for $\frac{R_{rf}}{R_{dc}}$ vs the parameter KfR_{dc}. Since the term $K = lC$ is a constant for a given type of resistor, a family of such curves, one for each type of resistor, is obtained on plotting $\frac{R_{rf}}{R_{dc}}$ vs fR_{dc}.

The values of $\frac{R_{rf}}{R_{dc}}$ obtained in this investigation were plotted in this manner and can be compared with the theoretical curve. In examining these plots (see curves of Fig. 7 and 8), it is found that the points are distributed about a mean curve which resembles the form of the theoretical curve. The decrease of $\frac{R_{rf}}{R_{dc}}$ with increasing fR_{dc} is not, however, as rapid as in the theoretical curve and in fact the actual shapes of the mean curves differ from one another for the different types. Furthermore, the grouping of the plotted points about the mean is not very close and some rather wide deviations can be observed.

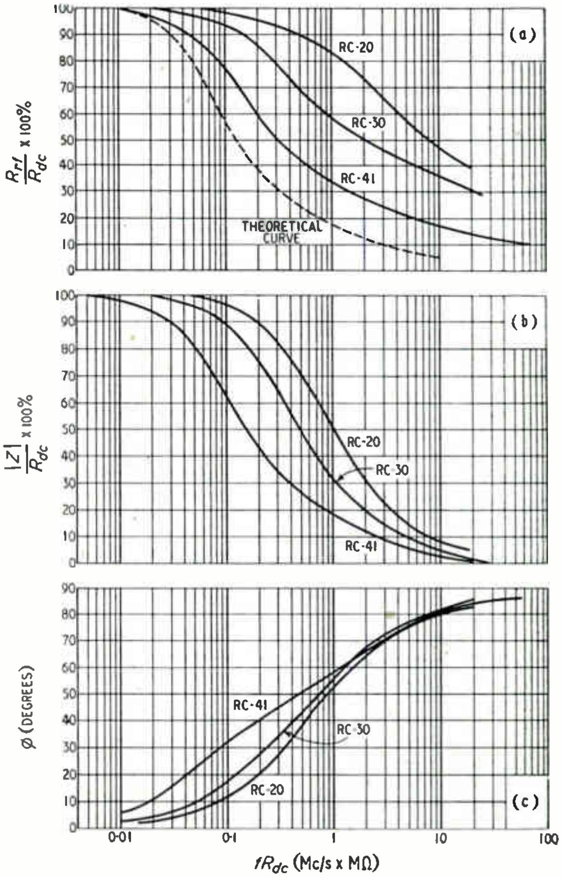


Fig. 9. *Mean curves for r.f. resistance (a), impedance (b) and phase angle (c).*

It is, therefore, indicated that other factors affect the r.f. resistance and that the transmission-line concept, while useful, does not fully account for the actual variation. Nevertheless, the predominant factors affecting the r.f. resistance appear to be those outlined above. That the percentage decrease in the r.f. resistance is a function of the d.c. resistance is shown since the mean curves are continuous.

The effect of physical size may be seen to good advantage by comparing curves for types RC-20, RC-30 and RC-41. These three types of resistor have the same diameter (of the actual resistor, not the insulating moulding) and differ from one another in length. The percentage decrease in resistance is the greatest for the RC-41 type which has the largest length/diameter ratio and also the largest shunt capacitance. The percentage decrease in the resistance and the shunt capacitance are the least for the RC-20 type which has the smallest length/diameter ratio of these three types.

It will be observed that in all cases, the decrease in resistance for the 1000-ohm resistors was initially very small and that at higher frequencies, the resistance increases again. The curves showing the variation of the shunt capacitance with frequency demonstrate that C_d decreases with frequency and is generally higher for low resistances than for the high-resistance values.

In addition to indicating the factors affecting the variation of resistance with frequency and serving as a means of comparing the different resistor types, the curves of $\frac{R_{rf}}{R_{dc}}$ vs fR_{dc} may also be used for predicting the r.f. resistance of any resistor of a given type once the curve is determined from measurements on a few samples. These curves may likewise be used for establishing specifications of the performance of resistors at radio frequencies.

Acknowledgment

The investigation was made possible by a grant from the Canadian Defence Research Board; the resistors were made available by the Canadian Signals Research and Development Establishment through the Interservice Sub-Committee on Electronic Standards. All of this assistance is hereby gratefully acknowledged.

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RADAR AWARDS

The Royal Commission on Awards to Inventors has, for some time, been investigating claims connected with the development of radar and has now made known its recommendations to the Treasury. The largest individual amount, £50,000, has been awarded to Sir Robert Watson-Watt, F.R.S., for his "initiation of radar and his contribution to the development of radar installations." The other awards are:—

£12,000 each to Dr. E. G. Bowen and Mr. A. F. Wilkins.

£2,400 each to Mr. L. H. Bainbridge-Bell, Mr. H. Larnder, Mr. G. A. Roberts, Dr. D. Taylor and Dr. A. G. Touch.

£1,200 each to Mr. W. A. S. Butement, Mr. R. Hanbury-Brown and Mr. P. E. Pollard.

£750 each to Mr. R. H. A. Carter, Mr. H. Dewhurst, Dr. J. H. Mitchell, Mr. S. Jefferson and Mr. B Newsam.

£500 jointly to Wing Commander E. J. Dickie and Dr. B. J. O'Kane.

£250 each to Mr. P. A. Marchant and Mr. R. V. Whelpton.

£250 jointly to Mr. P. A. Marchant and Mr. D. A. Weir.

PHYSICAL SOCIETY'S EXHIBITION

The 36th Annual Exhibition of Scientific Instruments and Apparatus will be held this year from 3rd April to 8th April, excluding Sunday. It will be held partly in the main building of the Royal College of Science, Imperial Institute Road, and partly in the Huxley Building, Exhibition Road, London, S.W.7. On the first day the exhibition will be open to members of the Society only.

RANDOM NOISE

Rate of Occurrence of Peaks

By V. J. Francis, B.Sc., M.I.E.E., A.R.C.S., F.Inst.P.

(Communication from the Staff of the Research Laboratories of The General Electric Co., Ltd., Wembley, England.)

SUMMARY.—In this paper are given the results of approximate calculations on the number of peaks occurring above various amplitude levels on a noise trace, the work bearing particular reference to the London-Birmingham Television Radio Relay Link.[1] The calculated results are compared with those of experiments made in the Laboratory to count the number of noise peaks above various amplitude levels on a system simulating as nearly as possible that of the Relay Link. An attempt is also made to relate these tests to the perceptible level of noise peaks on actual television pictures.

1. Introduction

IN the definition of the quality of transmission in a communication circuit of the 'waveform' type, the peak level of impulsive interference is a very important component. This is particularly true in the case of a television channel in which impulsive interference may cause a perceptible effect on the picture. A phenomenon of this kind can, of course, arise from interfering sources outside the channel under consideration, but in view of the formal specifications now being considered for transmission-channel quality and since the peak levels of random noise may contribute to the apparent interference, it becomes a problem of some importance to assess the likely number of peaks in the trace occurring at or above certain amplitude levels due to the inherent random noise of the system.

In view of the practical significance of this matter, the work reported here was undertaken in connection with the design of the London-Birmingham Television Radio Relay Link. To make any calculations at all, it was, of course, necessary to make several assumptions concerning the system and the noise source.

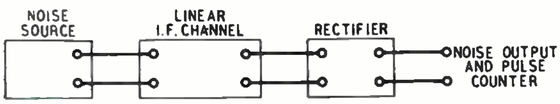
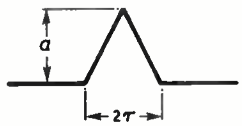


Fig. 1. *Arrangement of apparatus for the measurements.*

These were (1) that the system was linear up to the detector stage, and that following this the bandwidth was not so narrow as to cause appreciable distortion (Fig. 1), (2) that the noise originated in the linear system in a series of rapidly and randomly occurring 'events'[2], (3) that the response of the system to the random event was roughly that shown in Fig. 2. For the system in question, (3) is approximately true whether the fundamental event is a function zero until a certain time after which it is finite and constant, or an impulse function which is zero everywhere except over a very small period when the amplitude is very large. Here the interval τ is a function of the transfer characteristics of the system and is independent of the amplitude of the pulse. The calculations can easily be made for any law of detection which is a function only of the instantaneous amplitude. The calculations here are, however, confined to the simple linear and square law. The system for which the calculations were carried out, and for which experimental results are quoted was known to have something between a linear and a square law.

MS accepted by the Editor, February 1951

Fig. 2. *Response of the system to a random event.*



2. The Calculations for Identically-shaped Pulses

The method of calculation of the rate of occurrence of cuts and peaks at any level is known.[3] For the present purpose it is convenient to work with the rate of occurrence of 'upward cuts,' that is, the frequency with which the trace crosses the given level in an upward direction, rather than with the rate of occurrence of peaks above that level. There are difficulties in applying the formula for the peaks for some of the special cases of interest, and moreover it is in fact more likely that, from a subjective point of view, the number of upward cuts at a certain level which is approximately equal to the number of primary (or distinguishable) peaks is a better criterion of nuisance than the number of 'mathematical' peaks above that level, which would include all minor peaks many of which would not be observable.

If $s(t)$ is the time function for the individual event or pulse, the number of upward cuts at level y_0 is

$$C(y_0) = \frac{\sigma_1}{2\pi\sigma} e^{-y_0^2/2\sigma^2} \quad \ldots \quad \ldots \quad \ldots \quad (2.1)$$

where $\sigma^2 = \lambda \int_{-\infty}^{+\infty} [(t)]^2 \, dt$ .. .. (2.2)

$\sigma_1{}^2 = \lambda \int_{-\infty}^{+\infty} [\dot{s}(t)]^2 dt$ .. .. .. (2.3)

and λ is the number of pulses per second; σ^2 is the ordinary mean-square noise. The following results are expressed in terms of this parameter but, if the signal-noise performance of the system is known, the signal itself may be taken as the unit of amplitude.

For the waveform of Fig. 2

$$\sigma^2 = \frac{2a^2\tau\lambda}{3} \quad .. \quad .. \quad .. \quad .. \quad (2.4)$$

$$\sigma_1{}^2 = \frac{2a^2\lambda}{\tau} \quad .. \quad .. \quad .. \quad .. \quad (2.5)$$

$$C(y_o) = \frac{\sqrt{3}}{2\pi\tau} e^{-y_o^2/2\sigma^2} \quad .. \quad .. \quad .. \quad (2.6)$$

If τ is in seconds this gives the number of cuts per second at y_o.

To ascertain how sensitive $C(y_o)$ is to the precise form of the pulse, it was calculated for the pulses of Fig. 3 (a) and Fig. 3 (b) and for that of Fig. 3 (c) where the linear rise is followed by an exponential decay. For Fig. 3 (a)

$$s(t) = a \sin \frac{\pi t}{2\tau} \; (t = 0 \text{ to } 2\tau)$$

$$s(t) = 0 \text{ (elsewhere)}$$

and for Fig. 3 (b) the left-hand half of the cusp is given by $s(t) = at^n$ and the right-hand half is the obvious reflection of this. For Fig. 3 (c) the linear rise to amplitude a taking place in time τ is followed by an exponential decay of the form

$$s(t) = ae^{-\alpha t}$$

For the pulse form of Fig. 3 (a)

$$C(y_o) = \frac{1}{4\tau} e^{-y_o^2/2\sigma^2} \quad .. \quad .. \quad .. \quad (2.7)$$

For that of Fig. 3 (b)

$$C(y_o) = \frac{1}{2\pi\tau} \sqrt{\frac{n^2(2n+1)}{2n-1}} \, e^{-y_o^2/2\sigma^2} \quad .. \quad (2.8)$$

For the wave form of Fig. 3 (c)

$$C(y_o) = \frac{1}{2\pi} \sqrt{\frac{3\alpha(2+\alpha\tau)}{\tau(3+2\alpha\tau)}} \, e^{-y_o^2/2\sigma^2} \quad .. \quad (2.9)$$

The ratios of the coefficient (C') of the exponential in equations (2.7), (2.8) and (2.9), corresponding to the waveforms of Figs. 3 (a), 3 (b) and 3 (c) respectively, to that (C) of equation (2·6) are given in Table 1.

TABLE 1

Pulse Form	Ratio (C'/C)
Fig. 3 (a)	0·905
Fig. 2	1·0
Fig. 3 (b) $n = 2$ ..	1·5
Fig. 3 (b) $n = 3$ ..	2·8
Fig. 3 (c) $\alpha\tau = \frac{1}{10}$ ..	0·25

For the pulse of the general form of Fig. 2 or 3 (a) or (b), the effect of the increasing peakiness of the pulse shape is evident, but it is clear that, unless high values of n in equation (2.8) are taken, for practical purposes the effective coefficient is not likely to differ seriously from that for the form of Fig. 2. On the other hand a slow rate of decay for the form of Fig. 3 (c) has a more marked effect. It should be remembered that the figures in Table 1 give the ratio of the number of upward cuts at any level for the various pulse shapes with σ taken as the unit of amplitude.

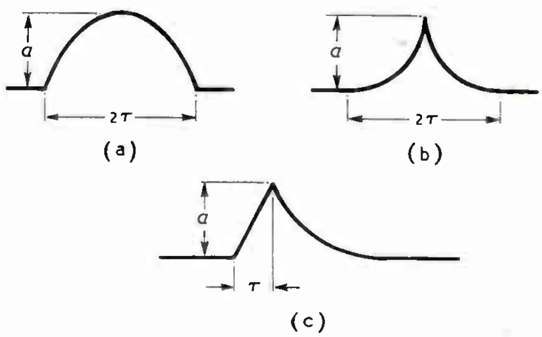


Fig. 3. *Three forms of pulse.*

3. Linear Rectification

For the pulse of Fig. 2 the relation between the amplitude y_o/σ and the number of upward cuts per second at this level is shown in Fig. 4 for the values $\tau = 0{\cdot}2\,\mu$sec and for $\tau = 0{\cdot}6\,\mu$sec corresponding to i.f. bandwidths of 6 Mc/s and 2 Mc/s respectively. The number of upward cuts differs in these two cases by a factor of 3.

4. Square Law Rectification

If detection takes place according to a square law, the instantaneous and r.m.s. voltages for the resulting trace are given except perhaps for a small numerical factor (see also reference 4) by the square of these for the linear trace. Accordingly consideration shows that the number of upward cuts at an amplitude $v_e (= k\sigma)$ on the linear trace is the same as the number of upward cuts at an amplitude $v_s = v_e{}^2$ on the square law trace.

But $v_s = v_e{}^2 = k^2\sigma^2 = k^2\sigma_s$ .. .. (4.1) where σ_s is the r.m.s. voltage on the square-law trace. Then, since $v_e/\sigma = k$ and $v_s/\sigma_s = k^2$, it follows that the curve for the square-law in Fig. 4 is obtained by squaring the values of y_0/σ for the linear curve; that is, doubling the abscissæ.

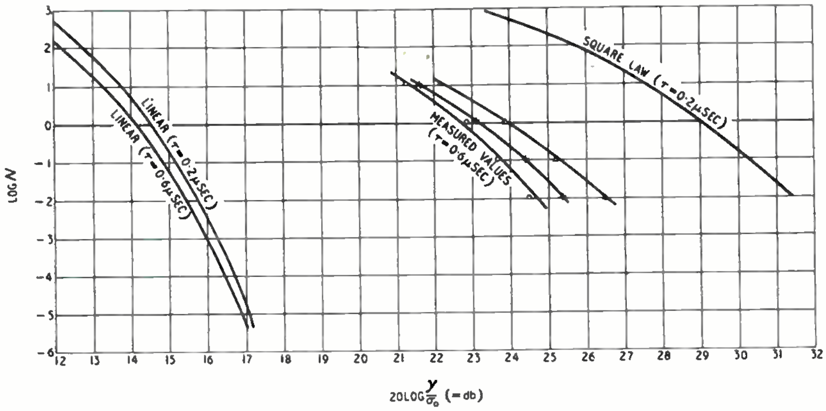


Fig. 4. *Calculated performance for linear and square-law rectifiers and measured results for a detector approaching the latter. N = number of upward cuts per second, y_o = voltage amplitude, and σ = r.m.s. voltage.*

5. The Mean 'Base' of a Peak

As a matter of interest and to add confirmation to the validity of the results it is worth while to estimate the variation in the mean 'base' of a peak with y_0.

This can be found from the consideration that the time the trace spends above the level y_0 is given by

$$t(y_0) = \frac{1}{\sigma\sqrt{2\pi}} \int_{y_0}^{\infty} e^{-y_0^2/2\sigma^2} dy_0 = \tfrac{1}{2}\left[1 - \text{erf}\ \frac{y_0}{\sigma\sqrt{2}}\right] \quad \ldots \quad \ldots \quad \ldots \quad (5.1)$$

The results are given in Table 2 for $\tau = 0{\cdot}2\ \mu$sec.

TABLE 2

y_0/σ	$\frac{1}{2}[1 - \text{erf}\ y_0/\sigma\sqrt{2}]$	$t(y_0)/C(y_0)$ sec
1	0·16	$0{\cdot}19 \times 10^{-6}$
2	$2{\cdot}3 \times 10^{-2}$	$0{\cdot}123 \times 10^{-6}$
4	$3{\cdot}3 \times 10^{-5}$	$0{\cdot}072 \times 10^{-6}$
6	$1{\cdot}0 \times 10^{-9}$	$0{\cdot}046 \times 10^{-6}$
8	$6{\cdot}7 \times 10^{-16}$	$0{\cdot}030 \times 10^{-6}$

It is seen that, at small values of y_0/σ the mean base is of the order of $0{\cdot}2 \times 10^{-6}\ \mu$sec, the value of τ. This is what would be expected from the frequency band-pass characteristics of the system.

6. Random Amplitudes

If the noise arises partly from resistors or even, except in special circumstances, if it arises from the current in a valve, the pulses will not be all alike. If the time taken to complete the actual event is small compared with the frequency characteristics of the circuit, the pulse length may, if the circuit transform impedance is suitable, be the same, but the amplitude will vary. We may therefore have a distribution of positive- and negative-going pulses with the shape, say, of Fig. 2, but with the amplitude distributed randomly.

Let $\lambda(a)da$ be the number of events occurring per second with amplitude between a and $a + da$; then since the mean squares of the noise for uncorrelated events are simply additive, we have

$$\sigma^2 = \frac{2\tau}{3} \int_{-\infty}^{+\infty} a^2 \lambda(a) da \quad \ldots \quad \ldots \quad \ldots \quad (6.1)$$

If λ is the total number of events occurring per second

$$\lambda = \int_{-\infty}^{+\infty} \lambda(a) da \quad \ldots \quad \ldots \quad \ldots \quad (6.2)$$

and

$$\overline{a^2} = \int_{-\infty}^{+\infty} a^2 \frac{\lambda(a)}{\lambda} da \quad \ldots \quad \ldots \quad \ldots \quad (6.3)$$

so that

$$\sigma^2 = \frac{2\tau}{3} \overline{a^2}\, \lambda \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (6.4)$$

It is clear, therefore, that it is the mean-square amplitude of the pulses which determines the number of peaks and that for any physically-occurring distribution, the conclusions of Section 2 will still hold.

7. Experimental Measurements

The apparatus used in the noise measurements comprised a substantially resistive noise source followed by an intermediate-frequency amplifier of 2-Mc/s bandwidth which delivered an output signal to a rectifier stage working under such low voltage conditions that it departed substantially from an ideal linear rectifier. The rectifier characteristic certainly contained terms of first and second orders so that its characteristic would appear between linear and quadratic but with an approximation to the latter. The rectifier output was followed by a video amplifier with a bandwidth of about twice that of the i.f. amplifier, so that substantially no further deformation of the signal followed the detector stage. A pulse counter, with a controllable threshold, was connected to the output of the video amplifier The gain of the amplifier was chosen to give a measurable level of noise, derived mainly from the grid circuit of the first stage; the threshold of the pulse counter was set at several levels (given as abscissae in Fig. 4) with respect to the r.m.s. noise. The pulse counts (N) were taken in each case.

In Fig. 4 are shown plotted three independent sets of such results for the i.f. bandwidth of 2 Mc/s corresponding to a value $\tau = 0{\cdot}6\,\mu$sec. It is clear from the calculations that the curves for a 6-Mc/s bandwidth would be only slightly displaced from these, so that any comparison made between the laboratory counts and the link system with a 6-Mc/s bandwidth is approximately valid.

An attempt was then made to relate these experimental counts to the perceptible levels of noise peaks in actual television pictures. Using a television receiver of normal type, it was found, for example, that peaks independently injected into the system and reaching a level of − 23 db with respect to the white signal (15–20 e.f.c.) are just perceptible on an unmodulated scan at some 6 e.f.c. if they occur at the rate of a few pulses per second or more. In the presence of a complex picture at the same mean brightness it seemed that the perceptible level rose to some − 18 db.

Conclusions

The calculated results (Fig. 4) suggest that the rectifier law was between linear and square law and nearer to the latter. This was believed to be effectively true on the system used, so that the agreement appears at least qualitatively satisfactory. From the observations made on the actual pictures, it appears that noise peaks occurring at the rate of a few per second will be noticeable if they reach a level as high as some − 20 db below peak white. The calculations show that this rate occurs with peaks 10–20 db above r.m.s. noise. Thus the important conclusion appears that in systems with a signal/noise ratio* not better than 40 db, interference due to the inherent noise of the system may not be avoidable.

* Signal noise ratio = ratio of the "voltage black to peak white" to "the r.m.s. noise."

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DUAL CIRCUIT OF A FEEDBACK AMPLIFIER

By D. A. Bell, M.A., B.Sc.(Oxon.)

THE term 'dual circuit' is used to mean "that circuit which is derived from the given circuit by a set of rules governing the mutual interchange of impedances and admittances, voltage generators and current generators, meshes and nodes." When the circuits consist of arrays of 2-terminal branches, as in Fig. 1(a), the rules for transforming between the dual pairs are well known, [1,2,3] and can be carried through by transforming impedances into admittances even though each 2-terminal impedance branch may be made up of a number of separate elements as exemplified in Fig. 1(b). But the dual of the feedback-amplifier circuit of Fig. 1(c) cannot be drawn immediately, since one component (the amplifier) is a 3-terminal impedance for which there is no direct duality rule. (It might be thought that a balanced 4-terminal amplifier would provide a more general case, but a balanced amplifier would in fact be a 5-terminal network since we shall find that the 'earth' point cannot be suppressed when forming the dual of a circuit. Hence a balanced amplifier would usually be treated as a pair of unbalanced amplifiers, and the 3-terminal unbalanced amplifier is the basic type to examine.)

MS accepted by the Editor, March 1951

Figs. 2 and 3 show the dual circuits for the representation of a triode valve, and the strictly dual equations are respectively:

$$V_{out} = \frac{-Z}{Z + R_a} \mu V_g \qquad \ldots \quad \ldots \quad (1)$$

$$I_{out} = \frac{-Y}{Y + G_a} g_m V_g \quad \ldots \quad \ldots \quad (2)$$

The conventions of signs in equations and diagrams are apt to be troublesome, but two essential requirements are (i) that the dual circuits should invert into each other carrying the

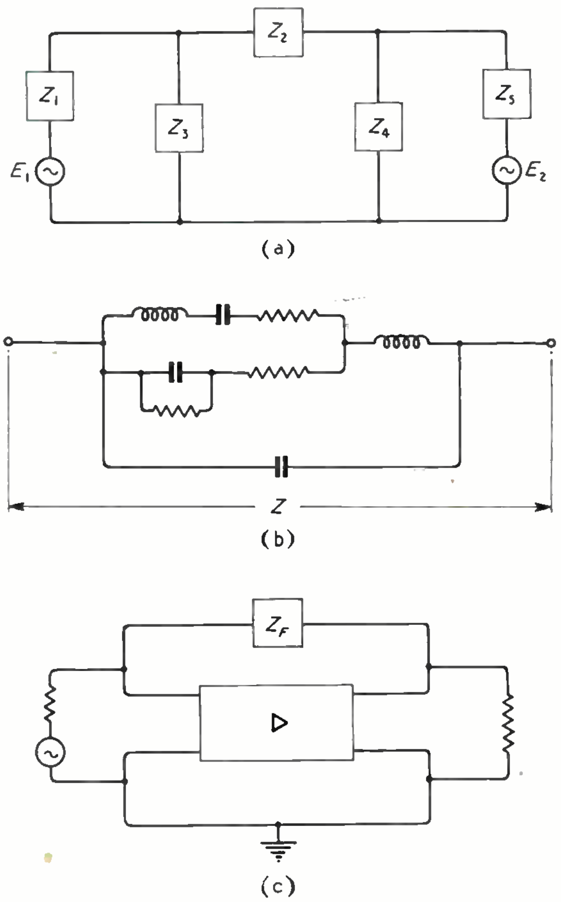


Fig. 1. *Network consisting of two-terminal branches (a) and an example (b) of a possible form of one of the branches. A network including a feedback amplifier (c) cannot be transformed by the same rules.*

correct sign, and (ii) that a single-stage amplifier of this type should show phase-reversal. In the circuit of Fig. 2 there is also the convention that the terminal p.d. of a generator is of the same sign as its e.m.f.; i.e., the voltage between A and C (anode and cathode) must be of the same sign as E. But the voltage A to C is the output voltage, so the overall phase-reversal requires E to equal *minus* μV_g. A possible convention in Fig. 3 is that the positive direction of current flow is upwards in all branches of the circuit. But it is obvious that the currents in the generator and in the load admittance must be one up and one down, and the output current is then proportional to *minus* the generator current. A similar result would be obtained for Fig. 2 if attention were directed to e.m.fs round the loop, instead of to potential differences from cathode.

It is common practice, however, to use a circuit with the configuration of Fig. 3 but with the valve and external load specified in terms of resistance and impedance respectively (instead of conductance and admittance) and with the output expressed as a voltage (instead of current). This destroys the similarity of the equations for the two diagrams, but the resulting expression

$$V_{out} = ZI_{out} = -g_m V_g Z \left[\frac{1}{1 + Z/R_a}\right] \quad (3)$$

is practically convenient since it readily allows of approximation when $Z \ll R_a$, and an output voltage is more often required than an output current.

The topological rules for constructing the dual of a circuit may be recapitulated as follows and are indicated by the dotted lines and lettered spots in the voltage-impedance versions of the circuits in Figs. 2 and 4.

(i) Within each mesh of the given circuit, place a spot. This will become a node in the dual circuit, as indicated in the diagrams by the same reference letter. For this purpose the free space outside the circuit is counted as a 'mesh space'.

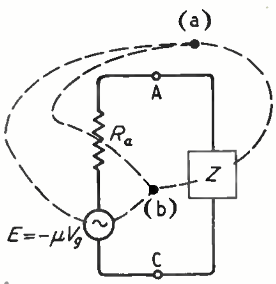


Fig. 2. *Equivalent circuit of triode valve.*

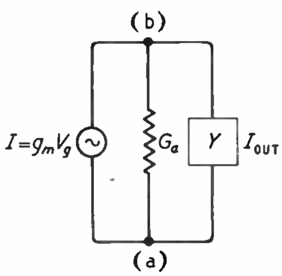


Fig. 3. *Dual of circuit of Fig. 2.*

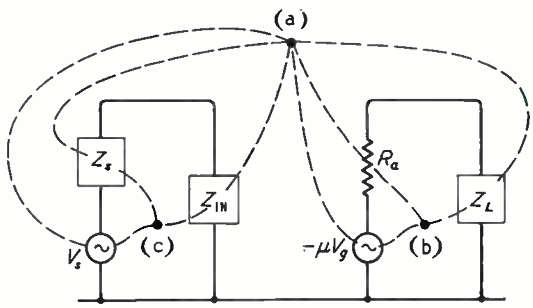


Fig. 4. *Equivalent circuit of a simple valve amplifier.*

(ii) Join pairs of spots by paths passing across all 2-terminal impedance and generator elements in the branches from which the meshes are constructed. (The pairs to be taken are all possible pairs for which the two meshes are adjacent; i.e., separated only by a branch of the circuit and not by intervening meshes.)

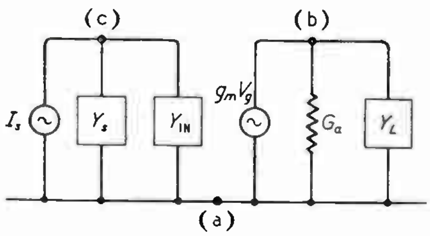


Fig. 5. *Dual of circuit of Fig. 4.*

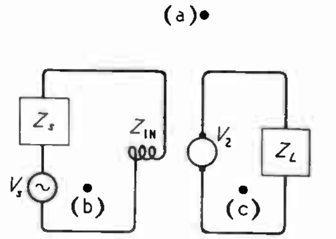


Fig. 6. *Field-controlled dynamo.*

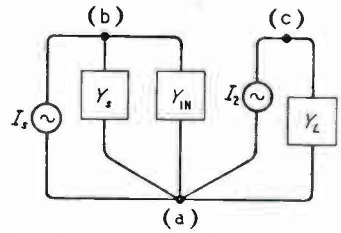


Fig. 7. *Dual of Fig. 6 showing the common connection (a).*

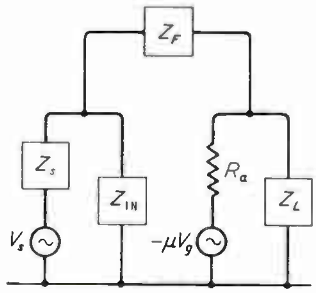


Fig. 8. *Single-stage feedback amplifier.*

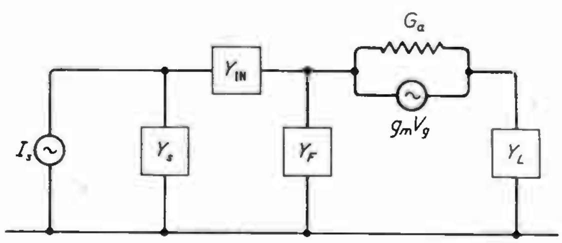


Fig. 9. *Dual of Fig. 8 for a feedback stage.*

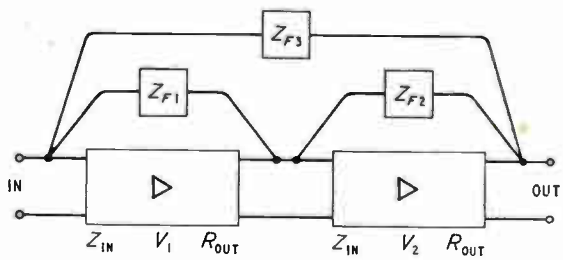


Fig. 10. *Two-stage amplifier with both local and overall feedback.*

(iii) For each path traced in (ii), connect in the dual circuit between the appropriate nodes an impedance which is the inverse of the one which was crossed in the given circuit.

Taking first a simple valve amplifier at low frequency, so that the anode current and voltage have no reaction on the input circuit, the voltage-impedance equivalent circuit is shown in Fig. 4. The input impedance Z_{in} takes account of both the input capacitance of the valve and any associated circuit such as a grid leak, and any series coupling impedance would be included in the source impedance Z_s. The only circuit connection between grid and anode circuits is that they have one common terminal, the cathode terminal of the valve. It may appear anomalous that a voltage V_g should be entered on the dual (current–admittance) circuit diagram given in Fig. 5, but this voltage is retained because the physical mechanism of the valve is of control by grid voltage. In computing the characteristic of the whole system via Fig. 5, one would perform the substitution $V_g = I_{in}/Y_{in}$. It will be seen that Fig. 5, like Fig. 4, has one terminal common to grid and anode circuits.

It is worth noting that if the given circuit had had no conductive connection between input and output (e.g., a magnetically-controlled deflection valve, a magnetic amplifier, or a field-controlled dynamo as indicated in Fig. 6) the dual circuit (Fig. 7) would still have had a node (a) common to both input and output circuits which corresponds to the spot (a) in free space in the given circuit. This is the diagrammatic representation of the fact that the system of Fig. 6 is not completely specified until the potential difference between some point on one circuit and some point on the other circuit is given. Conventionally, the potential of either circuit is referred either to 'earth' or to a 'point at infinity,' which is represented by the outer space in Fig. 6 containing the spot (a), a space which must necessarily be common to all parts of the system; and in the dual circuit (Fig. 7) the corresponding common node is the point which is very often the 'earth' terminal. It is this type of effect which usually requires the retention of the 'earth' terminal in a balanced system.

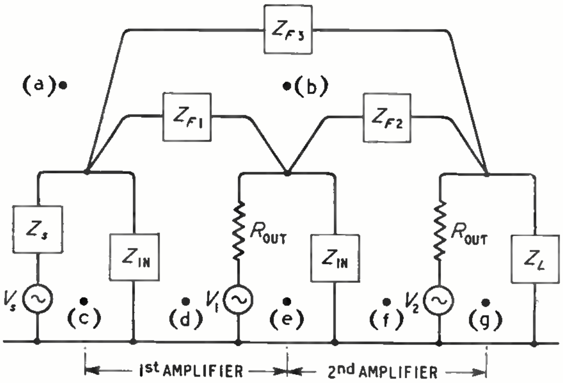
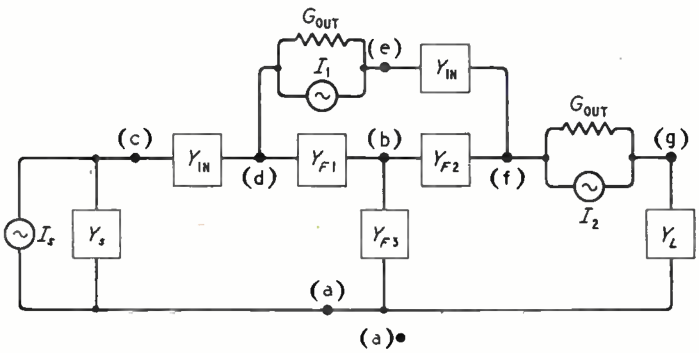


Fig. 11 (above). *Equivalent circuit of Fig.* 10.

Fig. 12 (right). *Dual of the circuits of Figs.* 10 *and* 11.

and an additional feedback over both stages (Fig. 10), which is troublesome to compute because one cannot treat the two stages separately.* In the equivalent circuit of Fig. 11 each of the amplifiers (which may in fact be multi-stage amplifiers) is represented by an input impedance, a generator and an output impedance. (The frequency restrictions imposed by the inter-stage couplings are ignored here, just as they are in using the ideal-amplifier symbol in Fig. 10.) In the dual circuit (Fig. 12) the original △ arrangement of feedback impedances has become, as one might expect, a Y arrangement of feed-back admittances.

The change from a given circuit to its dual cannot alter the characteristic behaviour of the system, and therefore there is no fundamental reason for assuming that this transformation will assist analysis. But there are cases in which one circuit is easier to 'read' than the other, and cases in which legitimate approximations or methods of compounding several circuit elements are more obvious in the transformed circuit than in the version originally given.



Having established the dual of a single amplifying stage, there is no difficulty in treating a single-stage amplifier which has added to it a voltage-feedback branch (Figs. 8 and 9). Note that if the positions in the diagram of two parallel branches such as V_s, Z_s and Z_{in} are interchanged, so that the one which was pre-viously adjacent to 'free space' is now adjacent only to internal meshes, the corresponding pair of series elements is interchanged in the dual circuit so that the one which was originally con-nected to the 'earth' line is now separated from it by the other one.

A more complicated example is the combination of two amplifiers with local feedback round each

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* cf. Fig. 8(c) in "Some Considerations in the Design of Negative-Feedback Amplifiers", by W. T. Duerdoth. *Proc. Instn elect. Engrs.*, 1950, Vol. 97 (Part III) p. 138.

A NETWORK THEOREM

By E. E. Zepler, Ph.D., M.Brit.I.R.E.

In a linear network containing a generator E it is frequently of interest to know the effect on the generator of connecting an impedance Z across two points of the network. While there are, naturally, many equivalent circuits showing the effect of the added impedance, there is one which seems particularly useful in many cases; this replaces the added impedance Z by another impedance Z_x in parallel with the generator. The magnitude of this new impedance is given by the following rule. We visualize that in order to find the current in Z the network has been replaced, in accordance with Thevenin's theorem, by a generator of e.m.f. E' and internal impedance Z_i. Then the power in this, in a sense fictitious, circuit is $E'^2/(Z_i + Z)$, where the term power is used for want of a better word; it may, therefore, be partly or wholly imaginary. Our rule then states that the impedance Z_x is such that the power in Z_x is equal to

$$\frac{E'^2}{Z_i + Z}\text{ ; i.e., } Z_x = \left(\frac{E}{E'}\right)^2 (Z_i + Z).$$

The proof of this is fairly simple, if we replace the network by a π-section (Fig. 1). Z_1 does not enter into the proof and will be disregarded. When an impedance Z is connected across Z_3 the power from the generator is

$$\frac{E^2}{Z_2 + \dfrac{ZZ_3}{Z + Z_3}} = \frac{E^2}{Z_2 + Z_3} \cdot \frac{(Z_2 + Z_3)(Z_3 + Z)}{Z_2Z_3 + Z_3Z + ZZ_2}$$

$$= \frac{E^2}{Z_2 + Z_3} + \frac{E^2}{Z_2 + Z_3} \cdot \frac{Z_3^2}{Z_2Z_3 + Z_3Z + ZZ_2}.$$

The power expressed by the second term on the right is equal to that in the circuit obtained from Thevenin's theorem, as can be seen at once by expressing the term in the form

$$\left(E\frac{Z_3}{Z_2 + Z_3}\right)^2 \frac{1}{Z + \dfrac{Z_2Z_3}{Z_2 + Z_3}}$$

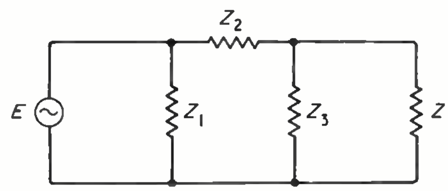

Fig. 1. *Network used in the proof of the theorem.*

MS accepted by the Editor, October 1950

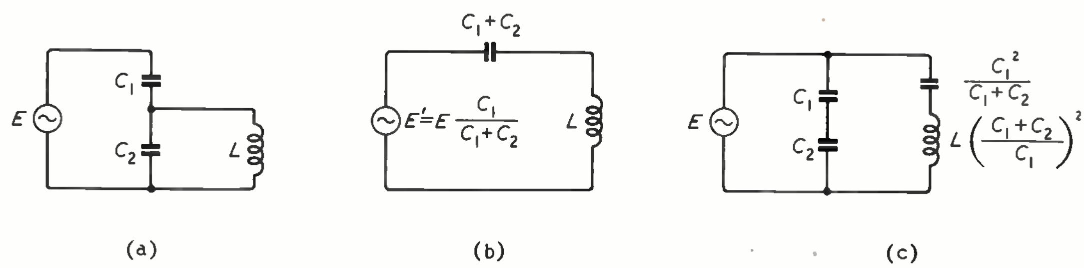

Fig. 2. *The circuit* (*a*) *is reduced to* (*b*) *by Thevenin's theorem and then to* (*c*) *by the new one.*

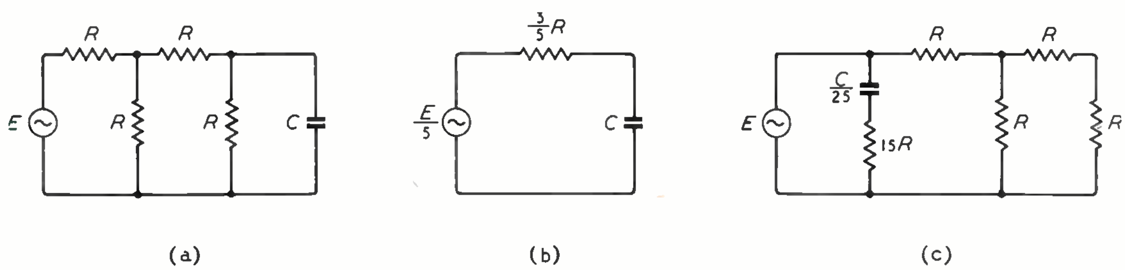

Fig. 3. *A further example of the application of the theorem to a network. The final circuit* (*c*) *is obtained from* (*a*) *through the intermediate step* (*b*).

Hence it is proved that the additional power taken from the generator E when Z is connected across Z_3 is the same as if the impedance Z_x, as defined above, were connected across E directly.

The practical use of the equivalent circuit may be seen from a few simple examples. In Fig. 2(a) the effect of connecting L across C_2 is to be replaced by an impedance across E. From the circuit of Fig. 2(b), obtained with the help of Thevenin's theorem, the power in the impedance to be placed across E and hence the impedance itself is found immediately. The equivalent circuit is shown in Fig. 2(c).

In Fig. 3(a) the effect of connecting C is found in a similar way. The circuit of Fig. 3(b) is the equivalent circuit as seen from C and obtained by Thevenin's theorem. Hence the power in Z_x is

$$\frac{E^2}{25}\,\frac{1}{\frac{3}{5}R + \frac{1}{j\omega C}}, \quad \text{and} \quad Z_x = 15R + \frac{25}{j\omega C}.$$

The equivalent circuit is shown in Fig. 3(c).

Fig. 4(a) shows a transformer with 'top capacitance'. If we wish to find an equivalent circuit from which the load across E may be readily seen it is best to replace the effect of connecting L_2 by an impedance Z_x across E. To render the circuit suitable for the use of Thevenin's theorem we employ the equivalent circuit of Fig. 4(b). When the switch is opened the p.d. between A and B is the difference between V_{AC} and V_{BC}. Now

$$V_{AC} = E\,\frac{C_1}{C_1 + C_2}, \quad V_{BC} = Ek\sqrt{\frac{L_2}{L_1}}, \quad \text{hence}$$

the e.m.f. of the equivalent generator is

$$E\left(k\sqrt{\frac{L_2}{L_1}} - \frac{C_1}{C_1 + C_2}\right).$$

Because of unity coupling between L_1 and k^2L_2 the latter inductance is short-circuited by a short-circuit across L_1. Therefore the inductance between B and C when E is replaced by a short-circuit is only $L_2(1 - k^2)$, while the impedance between A and C is given by the parallel combination of C_1 and C_2. The final circuit derived with the help of Thevenin's theorem is thus as shown in Fig. 4(c). The power in this circuit is

$$E^2\left(k\sqrt{\frac{L_2}{L_1}} - \frac{C_1}{C_1 + C_2}\right)^2 \frac{1}{j\left[\omega L_2(1 - k^2) - \frac{1}{\omega(C_1 + C_2)}\right]}.$$

Hence the equivalent circuit as seen from the generator E is that given in Fig. 4(d), where

$$p = k\sqrt{\frac{L_2}{L_1}} - \frac{C_1}{C_1 + C_2}$$

and where k is the coupling factor between the primary and the secondary of the transformer.

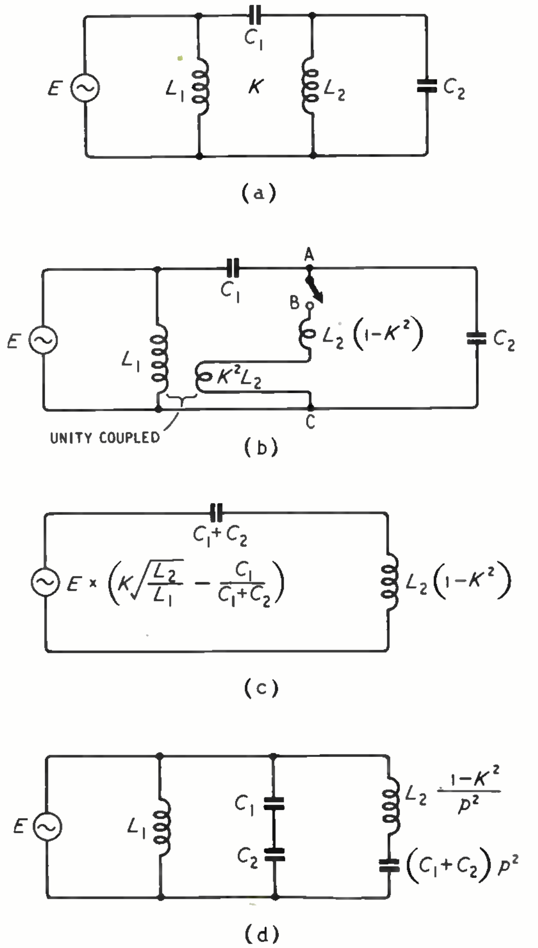

Fig. 4. *Equivalent circuits viewed from E of 'top-capacitance' coupled transformer.*

The term $k\sqrt{L_2/L_1}$ may be positive or negative, according to the winding sense of the transformer. With the usual type of inter-stage transformer the potential at B [Fig. 4(b)] will be large compared with the potential at A, so that the winding sense has little influence.

SOLAR ACTIVITY AND IONOSPHERIC EFFECTS

Comparison from May to November 1948

By R. E. Burgess, B.Sc., and C. S. Fowler, A.M.Brit.I.R.E.

(*Communication from the National Physical Laboratory*)

SUMMARY.—Observations of ionospheric propagation conditions on long waves (191 kc/s) and short waves (18·89 Mc/s) were compared with observations of solar noise bursts (on 30, 42, 73 and 155 Mc/s) and with the appearance of solar flares and sunspots. Detailed study of the correlation between these phenomena was made and statistical conclusions are presented.

1. Introduction

A PROGRAMME of recording solar radio-frequency radiation and ionospheric propagation was initiated at the Radio Research Station, Slough, in 1948. The objective was to determine the relation between ionospheric disturbances to long-wave and short-wave propagation and bursts of radiation at very high radio frequencies; the possibility was borne in mind that a statistical knowledge of any such relationship could be of use to either the communication engineer, in predicting ionospheric disturbances, or to the solar physicist, in observing solar activity particularly during times of visual obscuration.

The following phenomena were investigated:—

(a) short-wave ionospheric disturbances (s.w.d.) to high-frequency reception characterized by an abrupt fade-out,

(b) long-wave ionospheric disturbances (l.w.d.) characterized by a change of ionospheric reflection coefficient at low frequencies,

(c) bursts of solar radiation at radio frequency,

(d) solar activity in the form of flares or sunspots.

2. Observational Technique

Long-term daytime recording at Slough of propagational and noise effects was inaugurated in 1948 on the following basis:

(i) The strength of signals from the radio-telephony station at Klipheuvel, South Africa, operating on 18·89 Mc/s was recorded using a stable communication receiver with a.g.c. applied in order to compress the range of fading in the signal intensity. The station was selected in view of its consistent schedule during daylight hours and because the path crossed the equatorial regions and lay mostly in the illuminated hemisphere during daylight at Slough; thus it was particularly susceptible to fade-out conditions[1] and provided a reliable index of short-wave disturbances;

(ii) the signals from the long-wave broadcasting station at Königswusterhausen, Berlin, operating on 191 kc/s and at a distance of 975 km were used for recording long-wave ionospheric disturbances. This station transmitted for nearly all the day-

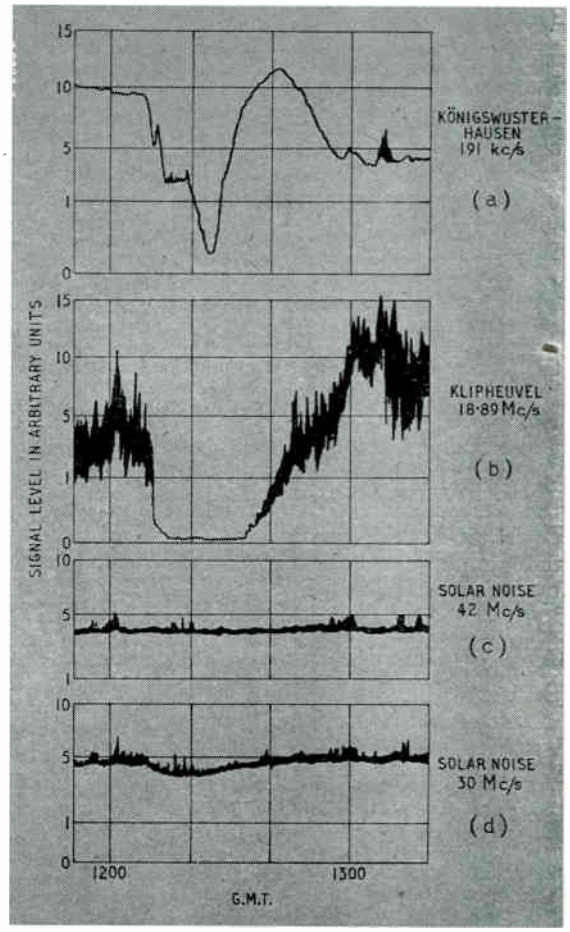

Fig. 1. *Ionospheric disturbances to long- and short-wave signals and related noise bursts (14th November, 1948).*

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time hours and its distance was such that, although the daytime ground-wave intensity exceeded that of the sky wave, it was possible to observe changes of the sky-wave component, in amplitude or phase, during an ionospheric disturbance. A horizontal aerial 200 ft long and 30 ft high was used for reception;

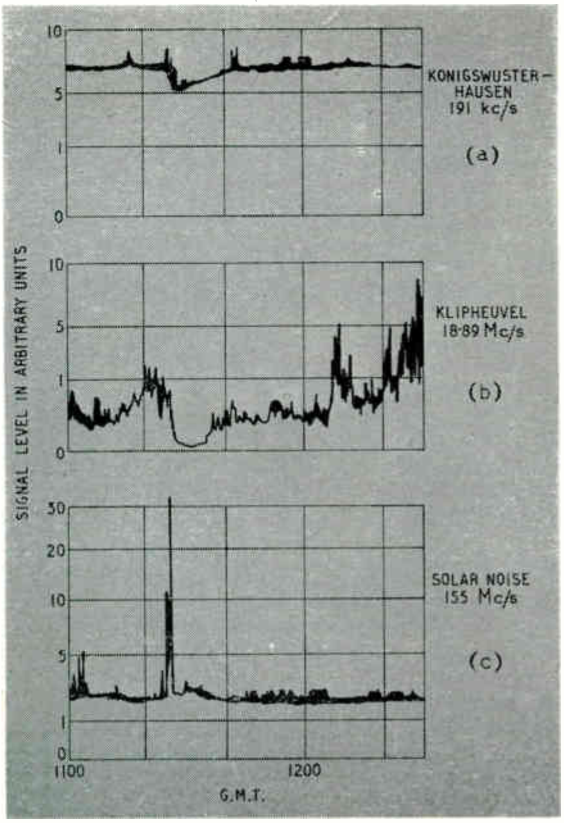

Fig. 2. *Ionospheric disturbances to long- and short-wave signals and related noise bursts; see also Fig. 3 for same occurrence (17th September, 1948).*

(iii) solar noise was recorded on frequencies of approximately 30, 42, 73 and 155 Mc/s. These frequencies were selected to be free from signal interference and sufficiently well separated to ensure the observation of a solar-noise burst on one or more channels. Low-noise receivers were used in conjunction with half-wave dipoles, erected at half a wavelength above the ground and lying perpendicular to the meridian. The time constants of the detector circuits and the damping of the recorder pen were adjusted to minimize the effect of interference produced by nearby ionospheric pulse transmitters and car ignition systems. It was found that the 'shapes' of the solar noise bursts recorded on the various channels were usually similar.

The rectified outputs from the six receivers were applied to two three-pen recorders using a paper speed of 3 in./hour. Time marks were applied to the charts every ten minutes from a crystal-stabilized clock and it is estimated that the accuracy of absolute timing on the records was $\pm$ 1 minute.

3. Data Analysed in the Study of Correlation

The following data were collected and examined for the period 14th May to 30th November 1948:

(i) *Short-Wave Disturbances (s.w.ds)*

A s.w.d. was taken as the occurrence of an abrupt decrease of level of the Klipheuvel signal with a subsequent slower recovery some minutes, or sometimes hours, later. Typical records are seen in Figs. 1 (b) and 2 (b). A comparison was also made with the s.w.ds reported by Messrs. Cable and Wireless, Ltd., which are based on interruptions to communications over world-wide circuits; viz., London-Europe, North America, South Africa, India and China. During the above-

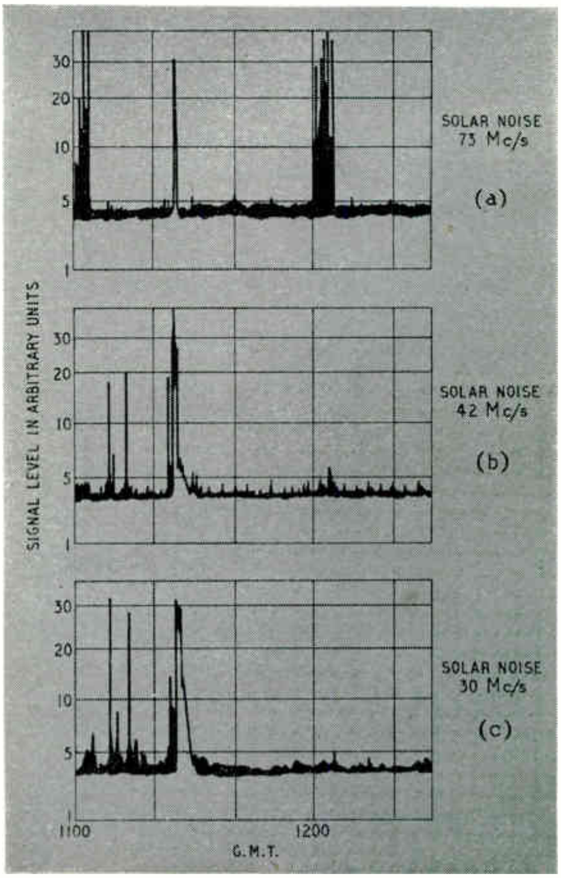

Fig. 3. *Comparison of noise bursts on different frequencies (17th September, 1948).*

mentioned period, a total of fifty-four s.w.ds were observed at Slough and by Cable and Wireless; of these only seven were not observed at Slough while twenty-two were not observed by Cable and Wireless. In view of the latter's extensive range of circuits these figures are regarded as a justification for the choice of path and recording technique adopted at Slough which, it is seen, enabled a substantially complete picture of sudden ionospheric disturbances to be obtained, since partial but abrupt fades could be taken into account, whereas these would probably not cause loss of commercial quality on a communication system.

the noise recording channels was taken to constitute a solar-noise burst. Examples of such a burst can be seen on the four frequencies in Figs. 2 (c) and 3 (a-c). In Fig. 1 (d) there is a general and prolonged decrease of the noise level on 30 Mc/s and this is due to the increased absorption of the galactic-noise background by the ionosphere during the 'fade-out' condition whose commencement preceded by about five minutes the occurrence of a solar-noise burst.

(iv) *Sunspots*

Only the largest of the sunspots listed in the Zürich "Heliographische Karten der Photosphäre"[2] were included in the analysis. They are classes E, F and G which have the following characteristics:

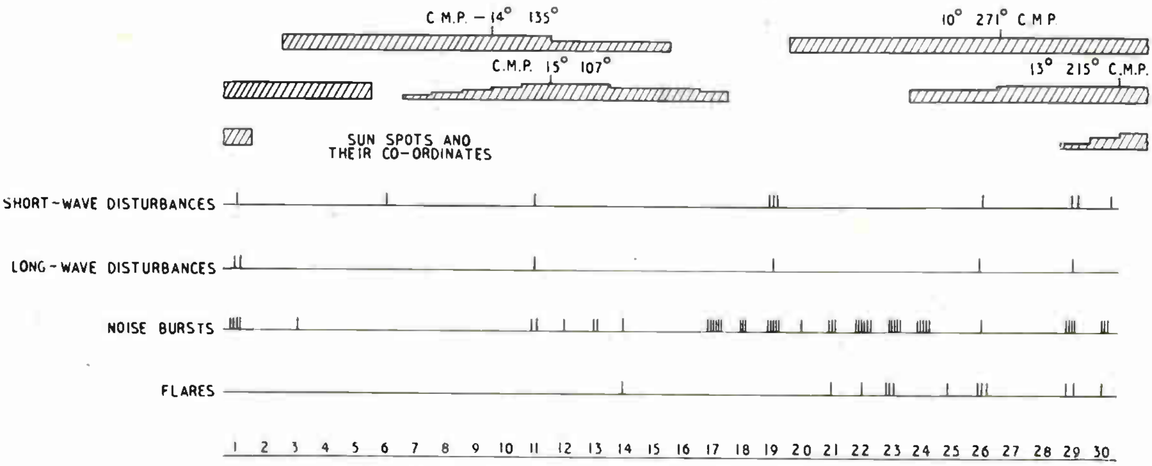

Fig. 4. *Radio and solar phenomena for July 1948.*

(ii) *Long-Wave Disturbances (l.w.ds)*

A sudden variation in the signal strength from Königswusterhausen was taken as constituting a long-wave disturbance. Usually during the daytime and especially in summertime the ground-wave signal exceeded the sky-wave by at least an order of magnitude but during a disturbance the change of low-frequency reflection coefficient of the ionosphere produced an observable change of resultant signal intensity. Any such sudden change followed by a relatively slow recovery (which is sometimes periodic) has been taken as constituting a long-wave disturbance. Examples are seen in Figs. 1 (a) and 2 (a) in which the enhanced sky-wave was in antiphase with the ground-wave and remained so during the disturbance, causing a depression in resultant signal level. Another type of disturbance has been described elsewhere[3] in which the signal varies in an oscillatory manner due to the change of amplitude of the sky-wave being accompanied by an appreciable change of phase path.

(iii) *Solar-Noise Bursts*

A sudden increase in level on more than one of

E—large bipolar sunspot group, both principal spots having penumbras, usually of complicated structure; intermediate region occupied by small sunspots. Total length of group at least 10°.

F—very large bipolar or complex sunspot group, length at least 15°.

G—large bipolar group without intermediate spots, length at least 10°.

(v) *Flares*

The times of occurrence of all the flares notified by Greenwich during the period of observation were included. Unfortunately, at the time of writing, the more complete Zürich data comprising world-wide observations are not available.

4. Correlation between the Solar and Radio Data

The times of commencement of an ionospheric disturbance on short waves (s.w.d.) and on long waves (l.w.d.) were found to be coincident within the accuracy of measurement. These phenomena also coincided in time with the commencement of a solar flare when observed. In some cases the commencement of the flare was not visible at Greenwich due to obscuration and the time quoted was that at which the flare was first seen.

The time of onset of a noise burst varied by up to 3 minutes from those of the other phenomena, and 68% of the bursts *preceded* the other phenomena by 1 to 2 minutes.

The table summarizes quantitatively the correlation between the phenomena described.

Fig. 4 shows a typical monthly summary of the time relationship between the phenomena described in the previous section. During periods of non-recording at Slough, a few s.w.ds and noise bursts were obtained from external sources but are not included in the statistical analysis.

5. Conclusions

The observed coincidence in time of solar flares and noise bursts with ionospheric disturbances shows that the former are not capable of providing a warning of fade-out conditions which would be useful to the communication engineer. However, the relative frequency of occurrence of solar-noise bursts in any period can be taken as one measure of solar activity which has possible application in the prediction of ionospheric storms and has obvious advantages over a parameter depending on visual observation.

It is seen that 86% of the noise bursts occurred without any accompanying ionospheric effect; of those noise bursts associated with ionospheric effects approximately three-quarters occurred with both s.w.ds and l.w.ds and approximately one-third of the s.w.ds and one-fifth of the l.w.ds occurred without noise bursts. It is concluded that there is a greater tendency for an active area of the sun to emit observable radio-frequency radiations than to emit ultra-violet radiation sufficiently intense to disturb the ionosphere significantly. This may be associated with a possible difference in directivity of the emissions of ultra-violet and radio-frequency radiation.

On all the occasions of an l.w.d., an s.w.d. was also observed whereas approximately 40% of the s.w.d. occurred without a corresponding l.w.d. This shows that a greater disturbance of ionization was needed to produce an observable l.w.d. than an s.w.d.

Half the observed flares were associated with some radio effect, and of those approximately one half were associated with an almost simultaneous l.w.d., s.w.d. and noise burst, and one-quarter with a noise burst alone. When a visible flare was accompanied by a noise burst the times of onset were coincident to within a minute and this fact may be of value to the astronomer during times of obscuration of the sun in fixing the exact time of commencement of a flare which later becomes visible.

No quantitative conclusions can be drawn regarding the correlation between sunspots and radio phenomena since during the seven-month

TABLE

1948	RADIO PHENOMENA							FLARES	
	Noise burst only	S.W.D. and noise burst	L.W.D. and noise burst	L.W.D. S.W.D. and noise burst	S.W.D. only	L.W.D. only	S.W.D. and L.W.D.	Total number of flares	Number of flares with no associated radio effect
May	18(1)	0	0	5(2)	0	0	0	3	0
June	7	1(1)	0	5(1)	4	0	0	6	4
July	66(2)	3	0	7(4)	1	0	1	13	7
August	30(1)	1(1)	0	4	2	0	1	5	3
September	26	0	0	2(1)	1	0	0	2	1
October	81	5	0	2	3(1)	0	1	1	0
November	22	0	0	1	0	0	4	0	0
Total	250(4)	10(2)	0	26(8)	11(1)	0	7	30	15

The numbers in parentheses are the number of flares associated with the phenomena listed.

period of observation there was hardly a day during which there were not one or more large spots on the sun.

Acknowledgment

The assistance of Mr. R. W. N. Mason in the construction and operation of the recording equipment is acknowledged. The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of the National Physical Laboratory and the Director of Radio Research of the Department of Scientific and Industrial Research.

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[2] Waldmeier, M., "Heliographische Karten der Photosphäre für das Jahr 1948" (Zürich, 1949).
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COMPOSITE LADDER FILTERS

Second-Order Image Impedances

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(B.B.C. Engineering Division)

SUMMARY.—The first part of the paper indicates a method of designing composite ladder-type filters with equal or inverse impedance functions of the second order which are independent of the propagation function. Complete design formulae are given for filters comprising one, two, and three half-sections.

The filters described in the second part have an impedance function of the second order at one pair of terminals only.

The advantage of these systems of design is that the filters so obtained have less components than those designed by any other method and are therefore more economical to construct.

1. Introduction

THE image impedance of a filter is a function of frequency. An impedance function of the first order or constant-k varies considerably over the pass band, whereas an impedance of the second order or m-derived can be made relatively flat over a large part of the pass band by a suitable choice of m. The number of elements in a filter section increases with the order of its image impedance. In designing a composite filter with second-order image impedances it is customary, therefore, for the sake of economy to design the bulk of the filter with cascaded sections having an image impedance of the first order, and to terminate the filter with half-sections having an image impedance of the first order on one side and of the required second order on the other side. An example of this is shown in Fig. 1.

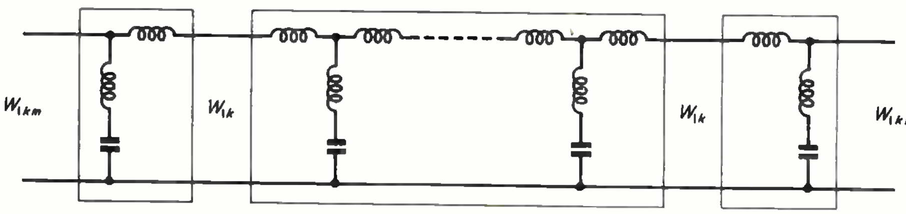


Fig. 1. *Composite filter of 1st-order sections terminated by 2nd-order half-sections.*

If the impedance function of the terminating half-sections is prescribed, this determines their attenuation function, as the two functions are interrelated, and unless this attenuation function is sufficiently alike to one of the sections in the bulk of the filter for the latter to be omitted, the process of terminating a filter in a second-order impedance function adds to the cost of the filter. This is particularly the case when there are only a few sections in the bulk of the filter because, first, there is less likelihood of employing the attenuation of the terminating half-sections to supersede a section in the bulk of the filter and, secondly, the smaller the bulk of the filter the larger is the termination in proportion to the whole.

Starting with a filter designed in this manner, however, it will be shown how the half sections, which are connected to each end of the filter to get the correct impedance characteristic, may be brought together at one end of the filter and then removed without altering the filter impedance.

MS accepted by the Editor, March 1951

2. Notation

The following notation will be used:—

$m_1, m_2, - - -$ = m values of sections used for their contribution to the attenuation
m_a = m value of half-sections used because of their impedance properties to terminate the filter
W'_{1k} = 1st order mid-series image impedance
W'_{2k} = 1st order mid-shunt image impedance
W'_{1km} = 2nd order mid-series image impedance
W'_{2km} = 2nd order mid-shunt image impedance
z = reactance of the series arm of a general *half*-section constant-k filter
y = susceptance of the shunt arm of a general *half*-section constant-k filter

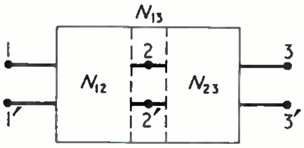
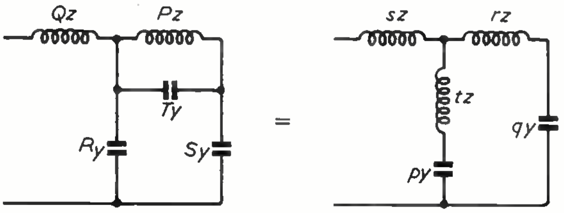


Fig. 2 (right). *Network* N_{13} *split into two* N_{12} *and* N_{23}.

Fig. 3 (below). *Two equivalent networks.*



3. Theorem and Transformation Formulae

Use will be made of a theorem and transformation formulae derived by the author in two previous papers.* The theorem states that if the 4-terminal network N_{13} of Fig. 2 can be split into two cascade-connected 4-terminal networks N_{12} and N_{23} in such a way that both N_{12} and N_{13} have equal image impedances at terminals 1,1′ then N_{12} and N_{23} have equal image impedances at terminals 2,2′.

The transformation formulae which have been slightly modified from the original are:

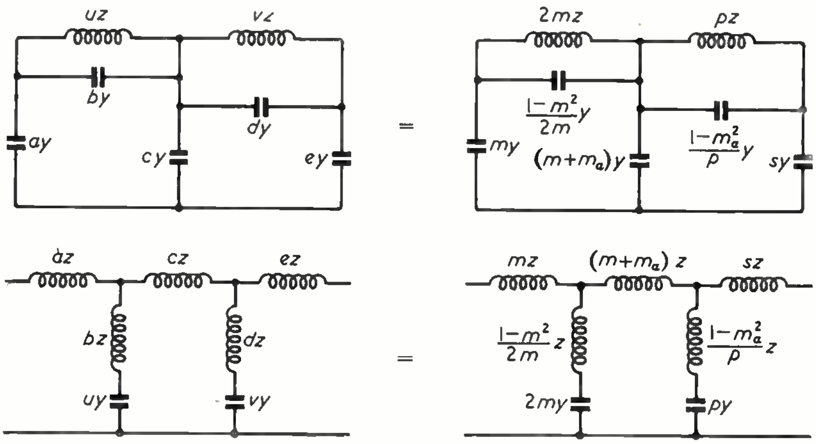


Fig. 4 (right). *Two pairs of equivalent networks.*

I.

The networks of Fig. 3 are equivalent if:

$$P = \frac{r^2}{r+t} \qquad Q = \frac{rt}{r+t} + s \qquad R = \frac{p\,(r+t)}{r}$$

$$S = \frac{qr - pt}{r} \qquad T = \frac{pt\,(r+t)}{r^2}$$

* "Double Derived Terminations," *Wireless Engineer*, February and November 1946.

In particular if

$q = 1$, $r = 1 + m$, and $pt = 1 - m^2$, then:

$$P = \frac{(1+m)p}{p+1-m} \qquad Q = \frac{1-m^2}{p+1-m} + s$$

$$R = p + 1 - m \qquad S = m$$

$$T = \frac{(1-m)\,(p+1-m)}{p}$$

The same formulae hold when small letters and capitals are interchanged.

II.

The networks of Fig. 4 are equivalent when:

$$a = m_a \qquad b = \frac{(1 - m_a{}^2)\,(2mp + (m_a - m)^2)}{p\,(m + m_a)^2}$$

$$c = \frac{2mp + (m_a - m)^2}{p + m - m_a}$$

$$d = \frac{1 - m^2}{2m} \cdot \frac{2mp + (m_a - m)^2}{(p + m - m_a)^2}$$

$$e = s + \frac{m^2 - m^2{}_a}{p + m - m_a}$$

$$u = \frac{p\,(m_a + m)^2}{2mp + (m_a - m)^2}$$

$$v = \frac{2m\,(p + m - m_a)^2}{2mp + (m_a - m)^2}$$

The equivalence of these networks is complete, so that any one may be replaced by the other irrespective of what is connected to their terminals. The process of transforming a network into its equivalent may therefore be carried out on a group of components in the middle of a filter.

In some circuits to be transformed, the component represented by Sy does not appear. This only means that the value of S to be substituted in the equation is zero.

For clarity in the diagrams z is represented by an inductor and y by a capacitor as in the previous papers. The formulae of course hold when z and y are any combination of reactors.

4. Derivation of Filters

Type 1.1

The simplest type of filter with second-order impedances is one having a propagation function equal to half a prototype section. This is derived as shown in Fig. 5.

The filter shown in Fig. 5 (a) consists of half a prototype section joined to half a series m-derived section on each side, the purpose of the latter being to produce the required image impedances. After applying transformation formulae I to the right-hand side of the circuit, the equivalent filter network of Fig. 5 (b) is obtained. A whole derived section may now be detached from the left-hand side of the filter as shown in Fig. 5 (c). The section on the right-hand side will then have a propagation function equal to the prototype half-section and, as may be seen, its image impedances are both of the second order.

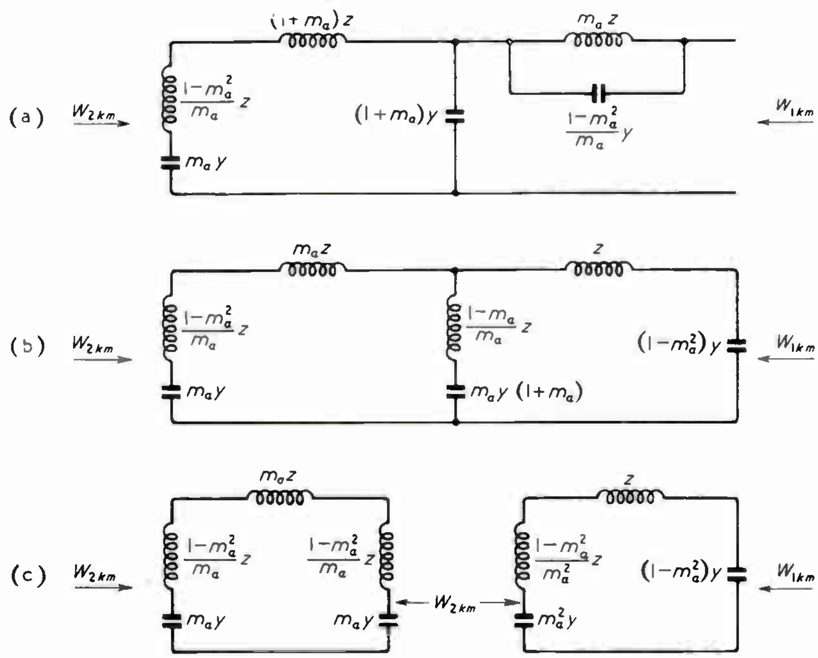


Fig. 5. *Derivation of the simplest 2nd-order impedance section of type* 1.1.

Type 1.2

The next type of filter to be considered is one with a propagation function equal to a full derived section. This filter is developed as shown in Fig. 6. Fig. 6 (a) shows the full section plus the terminating half-sections. Fig. 6 (b) shows the equivalent network obtained by applying transformation formulae II. The removal of a series-derived section from the left-hand side of the network leaves the filter shown in Fig. 6 (c) which is the one required. The values of the components in this filter are given below. A limitation is imposed by the fact that for e to be positive m_1 must be greater than m_a.

In the final network

$$w = \frac{2m_a^2 m_1}{m_a^2 + m_1^2}$$

$$v = \frac{2m_1^3}{m_a^2 + m_1^2}$$

$$f = \frac{(1 - m_a^2)(m_a^2 + m_1^2)}{2m_a^2 m_1}$$

$$c = \frac{m_a^2 + m_1^2}{m_1}$$

$$d = \frac{(1 - m_1^2)(m_a^2 + m_1^2)}{2m_1^3}$$

$$e = \frac{m_1^2 - m_a^2}{m_1}$$

In the second network

$$u = \frac{m_a (m_a + m_1)^2}{m_a^2 + m_1^2}$$

$$b = \frac{(1 - m_a^2)(m_a^2 + m_1^2)}{m_a (m_a + m_1)^2}$$

Type 1.3

A filter of this type designed in the usual manner is shown in Fig. 7 (a). After applying transformation formulae I to the components on the right-hand side, the network shown in Fig. 7 (b) is obtained. This may be transformed into Fig. 7 (c) by applying formulae II to all the components other than those in shunt with the input and output terminals. The removal of a shunt-derived section from the left-hand side gives the required filter as in Fig. 7 (d).

In the final network

$$w = \frac{m_a^2 (m_a^2 + 2m_1 + m_1^2)}{2m_a^2 m_1 + m_a^2 + m_1^2}$$

$$v = \frac{2m_1\,(m_a^2 + m_1)^2}{2m_a^2 m_1 + m_a^2 + m_1^2}$$

$$f = \frac{(1 - m_a^2)\,(2m_a^2 m_1 + m_a^2 + m_1^2)}{m_a^2\,(m_a^2 + 2m_1 + m_1^2)}$$

$$c = \frac{2m_a^2 m_1 + m_a^2 + m_1^2}{m_a^2 + m_1}$$

$$d = \frac{(1 - m_1^2)\,(2m_a^2 m_1 + m_a^2 + m_1^2)}{2m_1\,(m_a^2 + m_1)^2}$$

$$e = \frac{m_1\,(1 + m_1)}{m_a^2 + m_1}$$

In the third network

$$u = \frac{m_a\,(1 + m_a)\,(m_a + m_1)^2}{2m_a^2 m_1 + m_a^2 + m_1^2}$$

$$b = \frac{(1 - m_a)\,(2m_a^2 m_1 + m_a^2 + m_1^2)}{m_a\,(m_a + m_1)^2}$$

except that $g = c - m_a$

$$= \frac{m_a^2 - m_a m_1 + m_1^2}{m_1}$$

Denoting the tolerance by t we can proceed to evaluate the limit set upon m_1 in terms of m_a and t.

We have then that

$$w = \frac{2m_a^2 m_1}{m_a^2 + m_1^2} = m_a\,(1 - t)$$

i.e., $(1 - t)\,m_a^2 - 2m_a m_1 + (1 - t)\,m_1^2 = 0$

$$\text{giving } \frac{m_1}{m_a} = \frac{1 \pm \sqrt{1 - (1 - t)^2}}{(1 - t)}$$

$$= \frac{1 \pm \sqrt{2t - t^2}}{1 - t}$$

Filters, Type 2

In certain cases it may be desirable to have filters with a first-order impedance function at one pair of terminals and a second-order impedance function at the other pair of terminals, the latter impedance being independent of the propagation function. This is not strictly possible, but a solution may be obtained to a limited extent by taking into account the manufacturing tolerances that are allowed in practical filters.

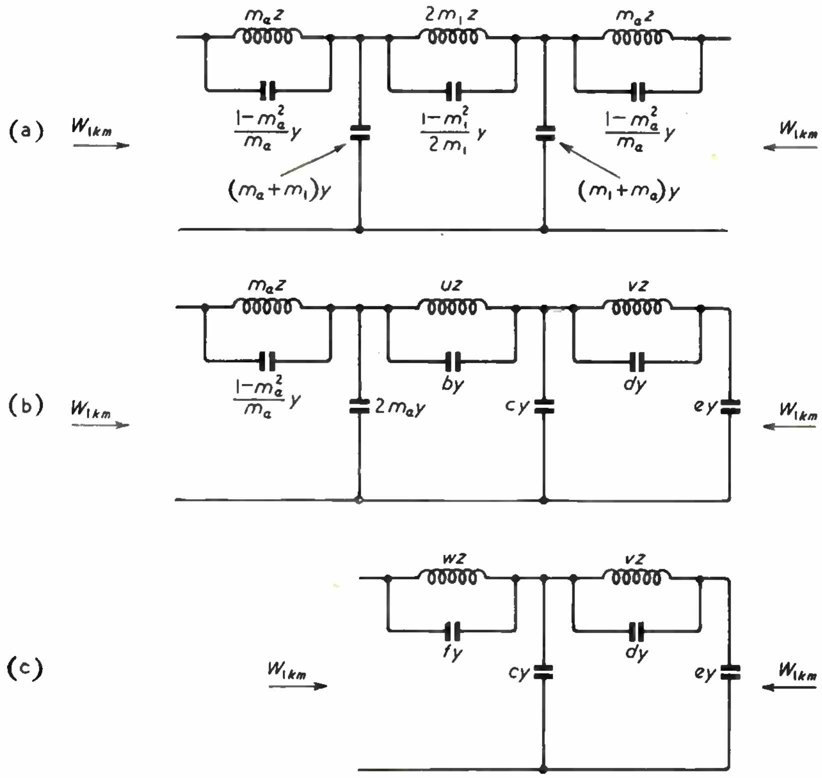


Fig. 6. *Derivation of type* 1.2 *section with 2nd-order impedance.*

Type 2.1

If in filter type 1.2 shown in Fig. 6 m_a is nearly equal to m_1, then w is nearly m_a. If the percentage by which it differs from m_a is tolerable from a practical standpoint a further half-section may be removed from the left-hand side of the filter. This has the effect of making the impedance on this side W_{2k}, which is of the first order. The filter will now be as shown in Fig. 8. The values of the components will be the same as in filter 1.2 Remembering the condition that obtained for filter type 1.2, namely, that $m_1 > m_a$ and approximating to the lowest power of t, we get that

$$m_1 = (1 + \sqrt{2t})\,m_a$$

Taking a practical value of t as up to 1% and m_a as 0·6, then m_1 may take any value between 0·6 and 0·685.

Type 2.2

A similar argument may be applied to the network of Fig. 7 (d). If the value of w is equal to $m_a(1 - t)$, we have this time

$$w = \frac{m_a^2(m_a^2 + 2m_1 + m^2_1)}{2\,m_a^2 m_1 + m_a^2 + m^2_1} = m_a(1 - t)$$

i.e., $m_a^3 + 2m_a m_1 + m_a m^2_1 = 2m_a^2 m_1(1 - t) + m_a^2(1 - t) + m^2_1(1 - t)$

or $m^2_1(1 - m_a - t) - 2m_a m_1(1 - m_a + m_a t) + m_a^2(1 - m_a - t) = 0$

giving

$$\frac{m_1}{m_a} = \frac{1 - m_a + m_a t \pm \sqrt{(1 - m_a + m_a t)^2 - (1 - m_a - t)^2}}{1 - m_a - t}$$

$$= \frac{1 - m_a + m_a t \pm \sqrt{t(1 + m_a)(1 - m_a)(2 - t)}}{1 - m_a - t}$$

and approximating to the lowest power of t

$$m_1 = \left(1 \pm \sqrt{2t\frac{1 + m_a}{1 - m_a}}\right) . m_a$$

The filter is shown in Fig. 9 and the values of the components are the same as in filter 1.3 except that

$$g = \frac{m_a^2(1 + m_1)}{m_a^2 + m_1} + m_1 - m_a$$

5. Applications

This type of filter has a rather wider application than that described in the introduction. Some of the circuits involving its use are shown in Fig. 10, where (a) shows a type-2 filter connected to an ordinary ladder filter, (b) shows a type-2 filter fractionally terminated and connected in series or in parallel with its complementary filter, and (c) shows two type-2 filters connected in tandem.

6. Conclusion

The type-1 filters described in this paper pro-

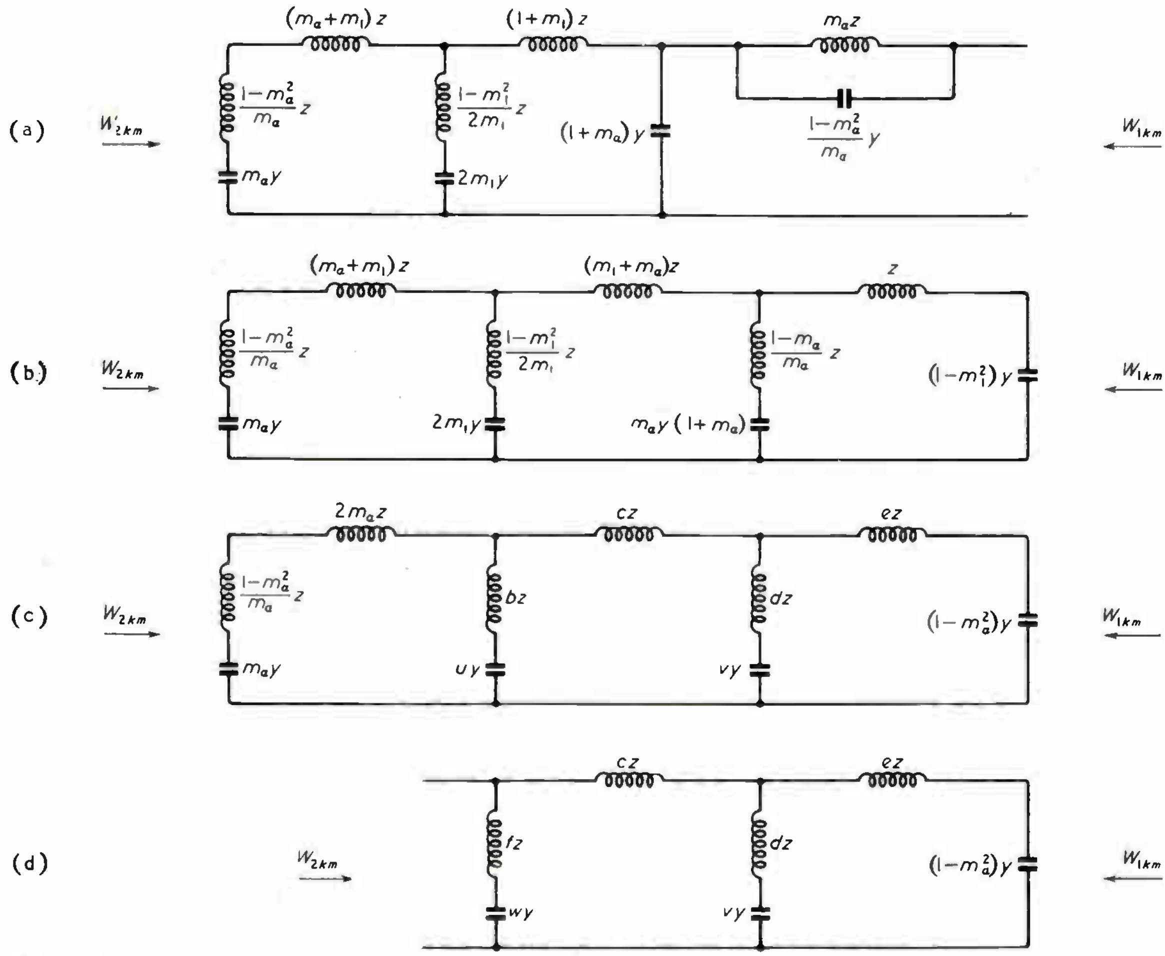


Fig. 7. *Derivation of type 1.3 filter. The usual form is shown at (a). After applying the transformation of formula 1 it takes the form (b), the application of formula 11 then brings it to (c) and the removal of the left-hand section gives the final form (d).*

vide attenuation and impedance functions equal to those of conventional design, but do so with a minimum number of components. The type-2 filters, however, are novel in that their attenuation function is not equivalent to that of conventional filters. In the latter the attenuation

Fig. 8 (right). *Type 2.1 filter section.*

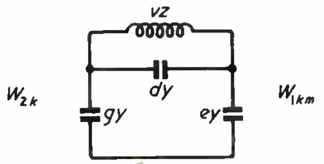
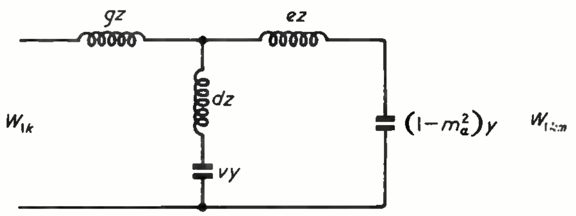


Fig. 9 (below). *Type 2.2 filter section.*



is the sum of that of a number of basic half-sections, whereas in the former it includes difference as well as sum. For example, filter type 2.1 was derived from a conventional filter consisting of an m_1 section and an m_a half-section by removing from it a whole m_a section. If the attenuation of the original filter is denoted by $m_1 + \frac{1}{2}m_a$ then that of the final filter will be $m_1 - \frac{1}{2}m_a$. Similarly the attenuation of filter 2.2 is that of an m_1 section plus a half prototype section minus an m_a half-section.

Each of the foregoing filters has, of course, a dual which may be obtained by the well-known method of interchanging inductors and capacitors,

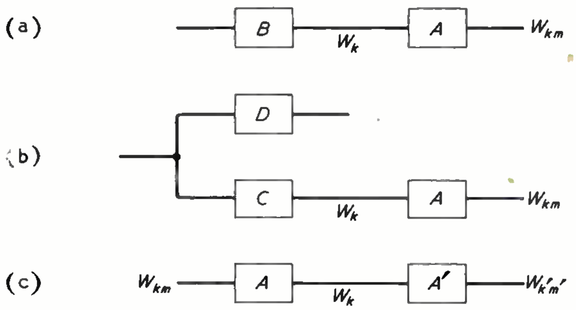


Fig. 10. *Methods of using type-2 filters. A and A′ = type-2 filters; B = ordinary ladder filter; C = fractional termination; D = complementary filter.*

and series and parallel connections. Again, although low-pass filters have been depicted throughout, the formulae apply equally well to high-pass or band-pass filters.

7. Acknowledgment

The author is indebted to Dr. Sturley for his valuable suggestions in the composition of this paper.

CORRESPONDENCE

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

Electrical and Magnetic Units

SIR,—In an Editorial, headed "Electrical and Magnetic Units", in the November issue of *Wireless Engineer*, G. W. O. H. critically questioned the wisdom of the new system of units which O. Löbl proposed in the *Elektrotechnische Zeitschrift* of 1st August, 1951.

Professor Löbl's proposed system of units is unlikely to affect the firmness of the m.k.s. (or Giorgi) system of units which took about 50 years to become established in the minds of the bulk of the electrical-engineering world, and which has already also been accepted by many physicists. However, his arguments to justify the proposal of a new system of units are very interesting because, implicitly, they bring to light an important point in favour of the Giorgi system which seems to have been overlooked in any discussion on systems of units.

The physicists of a hundred years ago found it convenient to use for the magnetic field the notations of magnetic flux density B and magnetic field strength H, and for the electric field the electric flux density D and electric field strength $\mathcal{E}$, on the understanding that in free space $B = H$ and $D = \mathcal{E}$. This convention was a mistake.

The mistake has now been rectified in the Giorgi system by introducing the conceptions of 'magnetic flux density of free space' and 'electric flux density of free space'. Hence, the conceptions of 'magnetic field strength (or magnetizing force)' (H) and 'electric field strength (or electric force)' ($\mathcal{E}$) are put into their proper places and they are given, as they are entitled to possess, their own and proper physical dimensions.

R. FEINBERG.

C.R.T. Laboratory,
Ferranti, Ltd.,
Moston, Manchester.
21st December, 1951.

Impedances in Parallel

SIR,—Referring to the graphical construction for the impedance equivalent to two impedances connected in parallel, may I bring to your notice a graphical solution further to those mentioned in the letters published in your journal, the January, April and July 1951 issues.

This construction is shown in the figure where first the points A and B are found at the intersections of the

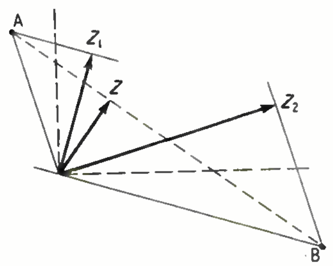
D

perpendiculars drawn respectively on Z_1 and Z_2, at their origins and ends. Thus the point A lies at the intersection of the perpendicular from the end of Z_1, and the perpendicular from the origin of Z_2.

The line AB is drawn next; the unknown vector Z is found to be the vector starting from the origin, and perpendicular to the line AB.

This simple solution was developed by me in a paper referring to various graphical solutions for the impedance of a network, and was published in Rumania, in *Buletinul Soc. Politehnice*, August 1929.

The proof of this construction can be easily found if reference is made to the construction based on two circles, mentioned in the Correspondence of your April 1951 issue.



These graphical constructions are useful, first, because they permit the impedance of a complicated network to be found by successive graphical transformations and, secondly, because they enable one easily to follow the modification of the impedance when a component varies. Thus, when a component varies so that the end of its vector describes a straight line or a circle, the end of the resultant impedance vector describes in the like manner a straight line or a circle.

T. TANASESCU.

Bucharest,
Rumania.
31st December, 1951.

SIR,—May I draw your attention to a construction for finding the impedance of two impedances in parallel, which has perhaps a little more elegance than that given by C. R. G. Reed (*Wireless Engineer*, January 1951, p. 32).

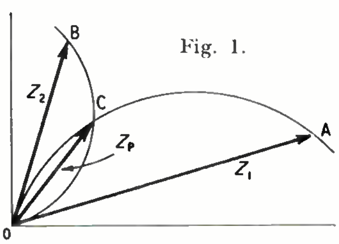


Fig. 1.

Let Z_1 and Z_2 be two given complex impedances, represented in Fig. 1 by OA and OB. Draw two circles through the origin, one through A and touching OB, the other through B and touching OA. Their intersection gives the required parallel combination Z_p.

Proof. We use some results from the geometrical theory of inversion. The points $r \exp i\theta$ and $r^{-1} \exp i\theta$ are said to be inverse with respect to a circle of unit radius centred at the origin. Well-known theorems in inversion state that the inverse of a straight line is a circle (or straight line) through the origin, and that the inverses of parallel straight lines are circles which touch at the origin.

In Fig. 2, A′ and B′ represent the inverse points of A and B. Since the parallel combination is the reciprocal of the sum of the reciprocals, the point C′ in Fig. 2 is the inverse of the required point C such that OC represents Z_p. Now C′ is the intersection of the lines A′ C′ and B′ C′, passing respectively through A′ and B′ and respectively parallel to OB′ and OA′. Therefore C is the intersection of two circles through the origin, passing respectively through A and B, and respectively touching OB and OA.

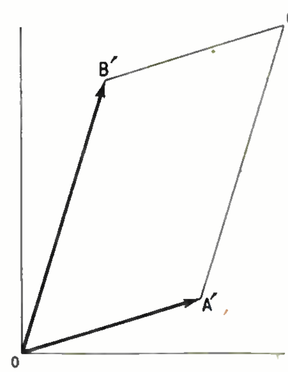


Fig. 2.

In the particular case when OA and OB are at right-angles, the solution is at the intersection of the semi-circles on OA and OB; i.e., at the foot of the perpendicular from O on to AB.

R. N. BRACEWELL.

Radiophysics Laboratory,
Commonwealth Scientific and Industrial Research Organisation,
Sydney, Australia.
1st November, 1951.

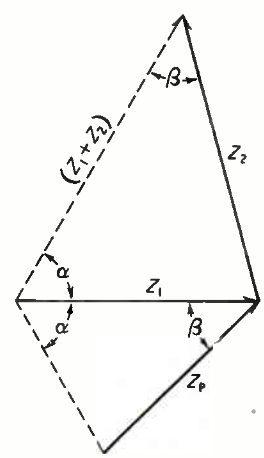
SIR,—I am referring to the correspondence regarding the geometric construction for the resulting impedance of the parallel combination of two impedances (*Wireless Engineer*, January, April, July, 1951). Unfortunately, I read this correspondence only recently.

It may interest your readers that a very simple construction had been published in *Electrical Review*, 19th and 26th December, 1947, pp.925–927 and 963–965, in a paper entitled "A.C. Network Analysis. Use of the Graphical Steady-State Method."

If Z_1 and Z_2 are the given impedances and Z_p is the impedance of their parallel combination (Z_1, Z_2 and Z_p are complex numbers) one can write

$$\frac{Z_p}{Z_1} = \frac{Z_2}{Z_1 + Z_2}$$

This equation states that the triangle formed by the vectors Z_p and Z_1 and the triangle formed by the vectors Z_2 and $(Z_1 + Z_2)$ respectively are geometrically similar and oriented in such a way that Z_p is turned with respect to Z_1 through the same angle as Z_2 with respect to $(Z_1 + Z_2)$. This leads to an extremely simple construction (see figure).



L. TASNY-TSCHIASSNY.

Electrical Engineering Department,
University of Sydney,
Sydney, Australia.
28th December, 1951.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research and published by arrangement with that Department.

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

ACOUSTICS AND AUDIO FREQUENCIES

016 : 534 293

References to Contemporary Papers on Acoustics.—R. T. Beyer. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 595–602.) Continuation of 2608 of 1951.

534.23-14 294

Effect of Temperature Inhomogeneities in the Ocean on the Propagation of Sound.—L. Liebermann. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 563–570.) Temperature measurements at different points at a given level exhibit variations of average magnitude 0·05°C over distances of the order of 60 cm. Reflection, scattering and focusing effects due to these inhomogeneities are investigated analytically, using the autocorrelation function, and also experimentally.

534.231 295

Discussion of Papers by Pachner and by Stenzel on Radiation from a Circular Emitter.—R. L. Pritchard. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, p. 591.) Stenzel's method of calculation (2707 of 1942), which was criticized by Pachner (1816 of 1951), is shown to be basically correct.

534.231 296

Measurements on an Acoustic Wave Propagated along a Boundary.—R. B. Lawhead & I. Rudnick. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 541–545.) Rudnick's theoretical analysis (3387 of 1947) was checked by measurements of amplitude and phase in the sound field of a point source located at a plane boundary between air and Fiberglas. The measured values agree well with calculated values based on measurements of the impedance and propagation constant of Fiberglas.

534.231 297

Acoustic Wave Propagation along a Constant Normal Impedance Boundary.—R. B. Lawhead & I. Rudnick. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 546–549.) An expression is obtained for the amplitude and phase in the sound field due to a point source on or near the boundary between air as upper medium and a lower nonisotropic medium with a constant normal impedance. The expression is developed in the same form as that for isotropic media (see 296 above and 3387 of 1947). A medium obeying the condition of constant normal impedance was constructed of tightly packed drinking straws. Measurements at various positions along and above the boundary agreed well with values given by the theory. The approximations introduced are such that the sound field is adequately represented at distances greater than one wavelength from the source.

534.231.3 298

On the Generalization of the Concept of Impedance in Acoustics.—O. K. Mawardi. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 571–576.) Present definitions of acoustic impedance are valid only when the specific impedance is constant on a wave front. This restriction is removed by extending the notion of vector fields to specific impedances. A definition based on energy concepts is proposed for acoustic impedance.

534.232 299

Synchronisation of Air-Jet Generators, with an Appendix on the Stem Generator.—J. Hartmann & E. Trudsø. (*Dan. mat. fys. Medd.*, 1951, Vol. 26, No. 10, 39 pp.)

534.232 300

The Air-Jet Generator as a Means for Setting Up Waves in a Liquid Medium.—J. Hartmann & F. Larris. (*Dan. mat. fys. Medd.*, 1951, Vol. 26, No. 11, 26 pp.)

534.232 301

On the Radiation Impedance of a Rectangular Plate with an Infinitely Large Fixed Baffle.—Y. Nomura & Y. Aida. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, March 1951, Vol. 1/2, No. 3, pp. 337–347.) Formulae for calculating the radiation impedance of a plate vibrating normally to the plane of the baffle are obtained from the pressure distribution on the plate. Numerical calculation indicates how the radiation resistance and reactance vary with the shape and size of the plate.

534.232 : 538.652 302

Performance Theory and Design Procedure for Laminated Magnetostriction Vibrators.—Y. Kikuchi & K. Fukushima. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 141–189.) Theoretical investigations are made of the performance of laminated magnetostriction vibrators for emitting ultrasonic waves in water and other liquids. Formulae for the calculation of the resonance frequency and the mechanical and electrical constants of rectangular- and ring-type vibrators are given. The effects of lamination space-factor and various material constants on the transduction efficiency are carefully studied, and design methods for this type of vibrator are determined.

534.232 : 538.652 : 621.3.017.32 **303**

On the Magnetic-Hysteresis Losses in Magnetostriction Vibration.—Y. Kikuchi & H. Shimizu. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, March 1951, Vol. 1/2, No. 3, pp. 365–379.)

534.26 **304**

The Diffraction of Sound by Circular Apertures.—T. Nimura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, March 1951, Vol. 1/2, No. 3, pp. 381–389.) Spheroidal wave functions are used in a rigorous theory of the diffraction of plane sound waves by a circular aperture with radius of the same order of magnitude as the wavelength. Numerical and graphical results are given for the sound fields in the neighbourhood of the aperture, the directional characteristics, acoustic impedances and the power transmitted through the aperture.

534.321.9 : 549.514.51 **305**

Mechanical Breakdown of Quartz Transducers at Resonance.—T. F. Hueter. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, p. 590.) Values given by Epstein, Andersen & Harden (2442 of 1947) for the maximum ultrasonic intensity attainable with a quartz crystal are discussed, and more recently determined values are given.

534.6 : 621.395.623 **306**

The Artificial Ear for Calibrating Telephony Receivers.—I. Barducci. (*Ann. Télécommun.*, June 1951, Vol. 6, No. 6, pp. 165–169.) See 1565 of 1950.

534.7 : 611.85 **307**

D.C. Potentials and Energy Balance of the Cochlear Partition.—G. v. Békésy. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 576–582.)

534.84 **308**

Wide [sound] Distribution from Radiator Groups.—F. Bergtold. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, p. 325.) Note complementary to the paper on sound reproduction in halls and open spaces (2890 of 1951).

534.845.1 **309**

A Free Field Method of Measuring the Absorption Coefficient of Acoustic Materials.—U. Ingård & R. H. Bolt. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 509–516.) The pressure and phase of an incident plane wave are measured first at a point on the absorptive surface and then at the same point in space with a perfectly reflecting material substituted for the absorptive material. Using the analysis for an infinite plane boundary, the absorption coefficient and the normal impedance for the particular angle of incidence are hence determined. Charts show the relations between these quantities and the measured phase difference and pressure ratio. Experimental results for materials of known properties confirm the validity of the method.

534.845.2 **310**

Absorption Coefficients of Fir Plywood Panels.—R. W. Kenworthy & T. D. Burnam. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 531–532.) Measurements made by the reverberation-chamber method are reported, using variously shaped panels, all of which exhibited greater absorption at 128 and 256 c/s than at higher frequencies.

534.845.2 **311**

Absorption Characteristics of Acoustic Material with Perforated Facings.—U. Ingård & R. H. Bolt. (*J. acoust. Soc. Amer.*, Sept. 1951, Vol. 23, No. 5, pp. 533–540.) An analysis of various sound-absorptive structures consisting basically of a layer of porous material separated from a rigid wall by an air cavity. Equations and design charts are given for the impedance and the absorption coefficient. Reverberation measurements give results in agreement with the calculated coefficients.

621.395.623.7 **312**

Amplitude and Phase Measurements on Loudspeaker Cones.—M. S. Corrington & M. C. Kidd. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1021–1026.) I.R.E. 1951 National Convention paper. Measurements of the motion of different points on a conical diaphragm were made at various critical frequencies. From these results the cause of various peaks and dips in the sound-pressure curve can be determined, making possible an improved design of loudspeaker.

621.395.625.3 **313**

On the True Frequency Response in Recording and Reproduction by Magnetic Methods.—R. Bierl. (*Z. angew. Phys.*, May 1951, Vol. 3, No. 5, pp. 161–165.) Consideration of reluctance in the magnetic circuit of the recording and reproducing head leads respectively to an inverse logarithmic and a hyperbolic function expressing the flux distributions. The flux in the reproducing head is then proportional to the differential of the magnetization, thus producing a frequency response which compensates that of the recording head. The sharp drop in high-frequency response previously attributed to a demagnetization effect, and the rise and final drop in the overall frequency response at low frequencies are easily explained by this theory.

AERIALS AND TRANSMISSION LINES

538.566 **314**

On the Propagation of Electric Waves from a Horizontal Dipole over the Surface of the Earth Sphere.—Nomura. (See 467.)

621.392 : 621.396.662 **315**

The Design of Transmission-Line Tuning Elements for Minimum Dissipation.—R. W. Klopfenstein. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1089–1094.) "A design procedure is described by which a coaxial transmission-line tuning stub may be designed for minimum energy dissipation when the input susceptance or reactance is specified. The problem has been formulated for air dielectric only where dielectric losses are ordinarily small compared to copper losses. The results are presented in curve form."

621.392.012.8 **316**

Transmission Line Equivalent Circuit.—L. A. Ware. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 287–288.) The development of a RLC circuit which simulates a $\lambda/4$ short-circuited line over a narrow band of frequencies is presented.

621.392.09 **317**

Propagation of Electromagnetic Waves along a Conducting Wire with Thin Dielectric Covering.—A. Fromageot & B. Louis. (*Bull. Soc. franç. Élect.*, June 1951, Vol. 1, No. 6, pp. 291–302.) The case is first considered in which the wire is perfectly conducting and the dielectric covering and surrounding medium are both without loss. The field and the propagation constants are determined and the field concentration and phase distortion due to the presence of the dielectric layer are calculated. The attenuations due to losses in the conductor, the dielectric, and the surrounding medium are then evaluated, and the complex constant of propagation is calculated, taking account of losses, so as to obtain an expression for the phase constant. The concentration of the field round the conductor facilitates matching of the line to terminal equipment and reduces the disturbing effects due to metal masses near the wire. The values found for attenuation and phase distortion compare favourably with those of coaxial lines and waveguides.

621.392.26† : 621.39.09 **318**

General Theory of Asymmetrical Waves in a Circular Waveguide with an Open End.—L. A. Vainshtein. (*Zh. tekh. Fiz.*, March 1951, Vol. 21, No. 3, pp. 328–345.) A rigorous solution is obtained of the problem of a circular waveguide within which an asymmetrical electric or magnetic wave travels towards the open end. Owing to diffraction effects the solution is more complex than in the case of symmetrical waves. Formulae are derived for the coefficient of reflection of an incoming wave from the open end and for the coefficients of transformation of this wave into others (including the transformation of electric waves into magnetic and vice versa). Approximate formulae are also derived for the radiation field. See also 2336 of 1951.

621.392.26† : 621.39.09 **319**

Numerical Results obtained from the Theory of Asymmetrical Waves in a Circular Waveguide with an Open End (E_1 and H_1 Modes).—L. A. Vainshtein. (*Zh. tekh. Fiz.*, March 1951, Vol. 21, No. 3, pp. 346–357.) For asymmetrical modes E_1 and H_1 in a circular waveguide graphs are plotted of the absolute values and phases of the coefficients of reflection from the open end and the coefficients of transformation from one mode to another. Graphs are also plotted for the radiation characteristics of modes E_{11} and H_{11}, and the results of the rigorous theory are examined in relation to Huyghens' principle. The excitation of various modes in the waveguide by a plane wave incident on the open end is also discussed.

621.392.26† : 621.39.09 **320**

The Propagation of Waves in Cylindrical Waveguides and the Hertzian Solution as Special Cases of the Propagation of Waves in Horns.—H. Kleinwächter. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, pp. 231–236.) Propagation in conical and pyramidal horns is analysed, using spherical coordinates. The Bromwich method is used for solving the wave equation, and the solution function is explained by examining the propagation of a particular mode. The method is further elucidated by deriving (*a*) the familiar solution for a cylindrical waveguide as the limiting case of a cone with zero aperture angle, and (*b*) the Hertzian solution for radiation from a dipole above a perfectly conducting plane as the limiting case of a cone with 180° aperture angle.

621.392.43 **321**

The Reactance of Mismatched Resonant Lines.—A. Ruhrmann. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, pp. 219–230.) The problem of matching an aerial to a feeder over a wide band of frequencies is treated analytically. For slight detuning of the operating frequency from the line resonance frequency the input impedance contains a reactive component proportional to the detuning; as a result, the input circuit can be represented by a damped oscillatory circuit with lumped parameters, whose reactance at resonance is easily determined. Various impedance-compensation and transformation circuits are analysed, and equivalent lumped-parameter circuits based on band-pass filter sections are established. The application of the theory for designing resonant-line circuits with predetermined properties is indicated.

621.392.43.012.3 **322**

Network Design Charts.—T. U. Foley. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 132, 134.) Sets of curves are provided which facilitate the determination of matching reactance limits of T, Π and L networks over a wide range of both phase shift and load resistance.

621.396.67 **323**

Theory of the Plane Annular-Slot Aerial.—J. Meixner & W. Kloepfer. (*Z. angew. Phys.*, May 1951, Vol. 3, No. 5, pp. 171–178.) The radiation pattern, radiated power and impedance of the aerial are determined as functions of λ, slot width and field distribution in the slot, for the case of a slot in an infinite plane. Numerical results are discussed. The radiation field plotted for a circular disk with a concentric annular slot in one face shows good agreement qualitatively with experimental results of Rhodes (1594 of 1949).

621.396.67 **324**

A Note on Super-gain Antenna Arrays.—N. Yaru. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1081–1085.) Numerical calculations have been made for linear broadside super-gain arrays. Using arrays having an overall length of a quarter wavelength as an example, it is shown that as the required directive gain is increased, tremendous currents are required to produce only a small radiated field, so that such arrays become quite impractical.

621.396.67 **325**

Radiation Patterns of Arrays on a Reflecting Cylinder.—J. E. Walsh. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1074–1081.) Simple approximate formulae are derived for the horizontal and vertical patterns of directional arrays of dipoles arranged on an arc of a circle or on the surface of a right circular cylindrical segment and backed by a reflecting cylinder. The dipoles may be aligned axially or circumferentially and the amplitude distribution on them may be uniform or cosinusoidal. The results are valid in the vicinity of the radiation maximum, but are less certain at wide angles unless the dipole spacing is substantially less than half a wavelength. The array patterns compare favourably with those of comparable linear arrays and in certain cases a greater gain is obtained.

621.396.671 **326**

Theoretical Calculation of the Input Impedances of Two Parallel Antennae.—S. Uda & Y. Mushiake. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 91–104.) Hallén's method for the calculation of the impedance of a single aerial is extended to a system of two linear aerials, and the results of numerical calculations and their applications are given.

621.396.677 **327**

Dielectric-Lens Aerial for Marine Navigational Radar.—D. G. Kiely. (*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, pp. 299–304.) "The design and performance of a dielectric-lens aerial for marine-navigational radar are described. The aerial has a fan-beam radiation pattern and is designed for horizontal polarization; its aperture and focal length are 4 ft and the maximum sidelobes are some 30 db below the main-beam level over the frequency band 9 320–9 500 Mc/s. This low sidelobe performance makes the aerial particularly suitable for marine-navigational radar application where the suppression of 'ghost' echoes due to sidelobes is important."

621.396.677 **328**

Aerials for Beam Stations.—K. O. Schmidt. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, pp. 313–315.) Supplement to a former paper (2352 of 1951, in which the title should be as above) giving a simple relation between aerial dimensions and beam angles.

621.396.677 : 621.397.6 **329**

Horn Antennas for Television.—D. O. Morgan. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 84–85.) An 8-ft equilateral horn aerial, with sides of wire mesh, has proved very effective in the reception of television and f.m. signals. Since for these signals horizontally polarized waves are of primary importance in the U.S.A., the top and bottom sides of the horn can be omitted. The aerial matches a 300-Ω line, one conductor being connected to each sector at the apex.

621.396.679.4 **330**

Nomographic Determination of Cable Efficiency in Feeding H.F. Energy from Transmitter to Aerial.—H. Geschwinde. (*Frequenz*, March 1951, Vol. 5, No. 3, pp. 67–69.) Formulae for the transmission of h.f. energy along a feeder cable are embodied in a nomogram whose use is illustrated by several numerical examples; the influence of the terminating impedance on the cable efficiency is shown.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.7† **331**

Variable Pulse Delay for Radar Ranging.—J. F. Gordon. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 100–103.) Based on paper presented at the 1950 National Electronics Conference (*Proc. nat. Electronics Conference, Chicago*, 1950, Vol. 6, pp. 94–102.) An electromechanical pulse-delay unit gives continuously variable or fixed pulse delays useful for radar ranging, navigation, propagation studies and similar techniques. Delay is obtained locally or remotely with range from a few microseconds to several milliseconds with a maximum error of 0·3 μs. A description is given of the circuits and their operation.

621.3.016.35 **332**

Note on Stability.—M. Parodi. (*J. Phys. Radium*, June 1949, Vol. 10, No. 6, pp. 200–201.) The sufficient conditions for a determinantal equation to have the real part of all its roots negative are stated, and the distribution of these roots in the neighbourhood of the origin is studied.

621.3.016.35 **333**

Supplement to a Note on Stability.—M. Parodi. (*J. Phys. Radium*, June 1951, Vol. 12, No. 6, pp. 665–666.) Generalization of the results obtained in 332 above.

621.3.016.35 **334**

Transmission Matrix Stability Criterion.—P. M. Honnell. (*Elect. Engng, N.Y.*, July 1951, Vol. 70, No. 7, p. 580.) Summary of 1951 A.I.E.E. Winter General Meeting paper. An analytical method is presented for predetermining the stability of a transmission network. The stability formulae apply to a network cut at any arbitary point, being thus more general than previous formulae.

621.3.016.35 **335**

Extension of Nyquist's Theory to the Case of Non-linear Characteristics.—A. Blaquière. (*C. R. Acad. Sci., Paris*, 30th July 1951, Vol. 233, No. 5, pp. 345–347.) The stabilized amplitude, the variation of frequency with amplitude, and the conditions for the frequency to be independent of amplitude to the first order, are calculated for oscillators with slightly nonlinear characteristics.

621.3.018.78† **336**

A New Method of Measuring and Analyzing Inter-modulation.—C. J. Le Bel. (*Audio Engng*, July 1951, Vol. 35, No. 7, pp. 18..31.) Two frequencies are mixed, without intermodulation, in a suitable circuit, passed through the system under test and then through a high-pass filter. The output of the latter is observed on a c.r.o. whose sweep is synchronized with the low-frequency tone. The resulting pattern can be analysed quantitatively to give the intermodulation percentage.

621.314.3† **337**

Analysis and Design of Self-Saturable Magnetic Amplifiers.—S. B. Cohen. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1009–1020.) I.R.E. 1950 National Convention paper. The theory of operation of the control element of a magnetic amplifier, the self-saturable reactor, is given and the concepts of extinction angle and firing angle are introduced. By assuming a characteristic for the core the two angles are related, thus making possible current and power calculations for the circuit. Applications of magnetic amplifiers are discussed and the merits of electronic and magnetic amplifiers are compared.

621.314.58 **338**

The Magnetic Modulator.—R. Feinberg. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 281–286.) A theoretical study of the modulator with sinusoidal transductor voltage and no-load or short-circuit output. Optimum performance is obtained with a core material whose magnetization curve has a narrow hysteresis loop, a sharp bend at the knee, and a high initial permeability, and with a transductor voltage giving a peak a.c. flux density in the core equal to the flux-density at the knee of the magnetization curve.

621.314.6.011.1 **339**

The Principles of Linear R.M.S.-Value Rectifiers.—O. Schmid. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, pp. 241–247.) The operation of the rectifier circuit described by Boucke (687 of 1951) is analysed. Methods are given for determining the appropriate values of circuit components. The magnitude of the residual waveform error is investigated and compared with that for area and peak rectifiers. For simplicity, consideration is restricted to operation with square-pulse, sawtooth-pulse and sine-wave voltages.

621.314.634 + 621.314.2 **340**

The New Siemens Selenium Rectifier for Broadcasting Equipment, and the Design of Suitable Transformers and Chokes.—Kühn. (See 505.)

621.316.86 : 621.396.822 **341**

Flicker Effect in Very Thin Carbon Films.—A. Blanc-Lapierre, M. Perrot & N. Nifontoff. (*C. R. Acad. Sci., Paris*, 16th July 1951, Vol. 233, No. 3, pp. 241–243.) Experimental results are given relating to the noise voltage occurring in resistors formed of thin carbon deposits, and in some commercial carbon resistors, due to the flow of a continuous current I. For a given resistance the spectral density, $e\nu^2$ of the noise voltage is a function of I and of the frequency, ν; the effect has been studied for I < 1 mA and ν between 600 and 800 c/s. Deviations of the results from Ohm's Law and from an I^2 law are discussed.

621.318.4 **342**

Internal Capacitance of a Multilayer Coil.—K. Jekelius. (*Frequenz*, March 1951, Vol. 5, No. 3, pp. 70–77.) Approximate formulae developed previously for calculating the uniformly distributed capacitance between two adjacent windings are reviewed, and a method is developed in which uncertainty due to estimating the mean dielectric constant is avoided, the calculation being based on the values of the dielectric constants for the winding insulation and the air spacing. A numerical example is worked out. The exact assessment and separation of internal and external coil capacitance is important in relation to the investigation of coil resonances.

621.319.4 : 517.35 **343**

A Solution for $\int e^{b(x + a\cos x)}dx$.—Emden: Rutishauser. (See 431.)

621.319.4.011.5 **344**

Harmonics of Current in [capacitor] **Dielectrics.**—B. Lavagnino. (*Alta Frequenza*, June/Aug. 1951, Vol. 20, Nos. 3/4, pp. 101–112.) Harmonic components of current passing through capacitors with imperfect dielectric are nearly always due to the presence of harmonics in the applied voltage; introduction or intensification of harmonics as a result of dielectric phenomena is to be expected only when the electric field is strong enough to produce ionization. The problem is discussed in relation to the sensitivity and accuracy of bridge measurements.

621.319.4.012.3 **345**

Temperature-Compensating Capacitor Nomograph.—T. T. Brown. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 132, 134.) An abac which gives, with one setting of a celluloid setsquare, the capacitance values required when two temperature-compensation capacitors are connected in parallel.

621.392 **346**

Extension of the Reciprocity Concept to Valve Circuits.—J. L. Bordewijk. (*Tijdschr. ned. Radiogenoot.*, May 1951, Vol. 16, No. 3, pp. 137–153.) Propositions of reciprocity theory previously shown to be valid for passive networks [3381 of 1948 (Bode)] are expressed so as to be valid also for active networks (i.e., including valves) by applying the concept of 'reversal'. The method is illustrated by comparison of the properties of the cathode-follower and the grounded-grid circuit and of oscillator arrangements with the resonant circuit connected in the one case to the anode and in the other case to the grid, and by examination of various amplifier arrangements.

621.392 **347**

The Fundamental Theorem of Electrical Networks.—J. L. Synge. (*Quart. appl. Math.*, July 1951, Vol. 9, No. 2, pp. 113–127.) An explanation for electrical engineers of Kron's method for the analysis of networks. The method of extracting 'meshes' and 'trees' from any network is discussed and Kirchhoff's laws are stated. The theorem relating the mesh currents to the branch currents is developed and the method of solution of a network using Kron's transformation matrix set out.

621.392 **348**

The Realization of a Transfer Ratio by means of a Resistor-Capacitor Ladder Network.—J. T. Fleck & P. F. Ordung. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1069–1074.) I.R.E. 1951 National Convention paper. A method is described for the solution of this problem, assuming that (*a*) poles of the transfer ratio $H(p)$ occur at real negative values of p and are simple, (*b*) zeros of $H(p)$ occur at real negative values of p, but may be multiple, (*c*) $H(p)$ is finite for $p = j\omega$, where $-\infty \leqslant \omega \leqslant \infty$. The resulting network usually has a gain factor considerably greater than can be obtained with a network synthesized on the lattice basis. To illustrate the method, the circuit parameters corresponding to a specified transfer function are determined.

621.392.5 **349**

The Dynamic Transfer Parameters of a Quadripole in Frequency Modulation.—G. Bosse. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, pp. 237–240.) Previously proposed methods for determining the dynamic transfer coefficient (i.e., the coefficient for the case of varying-frequency input voltage) are discussed. Spectral analysis of the input into components of constant frequency is laborious because of the large number of components involved. The input voltage can be considered as a series of pulses of infinitely short duration; the integral of the transient responses is related to the usual static transfer coefficient by a Laplace transform. The approximate evaluation of the dynamic coefficient may then be performed by graphical, numerical or analytical methods, depending on the particular frequency/time relation of the applied voltage.

621.392.5 **350**

The Transfer Function of an RC Ladder Network.—A. Fialkow & I. Gerst. (*J. Math. Phys.*, July 1951, Vol. 30, No. 2, pp. 49–72.) Analysis of four-terminal networks with two kinds of elements (mutual inductance excluded), including unbalanced networks having a common ground terminal. The topics considered are the zeros and poles of the transfer function, the realizability and the synthesis of the L-network and of the ladder network, with an illustrative example of the latter. The extension of the RC analysis to RL and LC networks is indicated.

621.392.5 : 517.432.1 **351**

Time-Dependent Heaviside Operators.—Zadeh. (See 432.)

621.392.52 : 621.317.3 **352**

Fundamental-Wave Filter for Instruments and Measurement Circuits.—H. Poleck. (*Frequenz*, March & April 1951, Vol. 5, Nos. 3 & 4, pp. 77–84 & 107–112.) Analysis of T-type and Π-type low-pass filters, particularly their properties at frequencies close to the fundamental and their attenuation of harmonics. The treatment is largely mathematical and the case in which the limiting frequency is $\sqrt{2}$ times the fundamental is particularly considered. Application of such filters in a rectifier-type voltmeter, a phase indicator with mechanical rectifier, and a wattmeter is described.

621.392.52 : 621.392.26† **353**

Tunable Waveguide Filters.—W. Sichak & H. Augenblick. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1055–1059.) I.R.E. 1951 National Convention paper. The bandwidth of a conventional microwave filter using resonant cavities tuned by irises is critically dependent on the resonance frequency. Another type of filter is discussed in which the cavities are tuned by changing the constants of the guide so as to maintain the wavelength in the guide constant at resonance. A filter using this principle has been built in which a strip of dielectric is inserted in a longitudinal slot in the broad face of the waveguide. The relation between resonance frequency and insertion depth was essentially linear between 4·4 and 5·0 kMc/s.

621.392.52 : 621.396.611.21 **354**

The Narrow-Band-Pass Crystal Filter for Short Waves.—Y. Watanabe & Z. Abe. (*Sci. Rep. Res. Inst. Tohoku Univ.*, *Ser. B*, March 1951, Vol. 1/2, No. 3, pp. 421–426.)

A method is described for eliminating adjacent frequencies in filter crystals by coating part of the crystal face with a damping medium such as Canada balsam.

621.392.6.012.8 **355**

Equivalent Circuits for 6-Pole and Higher-Order Multipole Networks.—G. Schmitt. (*Arch. Elektrotech.*, 1951, Vol. 40, No. 3, pp. 177–192.) Two types of 6-pole network are considered: (*a*) with three free pairs of terminals or (*b*) with two free pairs. For case (*a*) equations and matrices are given and, by a simple extension of the Π-equivalent for a quadripole, that for the 6-pole network is developed and from this the double-T equivalent. The physical significance of the elements of the matrices is explained. If a resistor is connected across one terminal pair of a 6-pole network, a quadripole results across each of the free terminal pairs. The matrices and equivalent circuits for this quadripole are investigated. It is also shown how the Π-equivalents of any higher-order multipole with more than three free terminal pairs can be derived from that of a 6-pole network. Case (*b*) is treated similarly.

621.396.611.3 **356**

Control Coupling between Two Oscillatory Circuits.—H. Wigge. (*Frequenz*, March 1951, Vol. 5, No. 3, pp. 63–67.) Various arrangements are illustrated and analysed in which two oscillatory circuits, connected to valves, are linked without feedback, each circuit acting on the other through an intermediate (control) element. In addition to three known coupling modes, each with two natural frequencies, two novel coupling modes are possible, each with three natural frequencies. The new couplings can be discussed in terms of classical theory by introducing imaginary values of coupling coefficients, inductances, resistances and capacitances.

621.396.611.4 **357**

Resonance Frequency of Spheroidal Cavity Resonator.—T. Nimura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 73–90.) Spheroidal coordinates and wave functions are used in analysis of the e.m. fields of oblate and prolate spheroidal cavity resonators. Resonance frequencies are determined for the case of rotationally symmetrical fields.

621.396.611.4 **358**

Coaxial Resonators Loaded by a Capacitance.—M. E. Zhabotinski. (*Zh. tekh. Fiz.*, March 1951, Vol. 21, No. 3, pp. 358–362.) Resonators of this type as used in u.s.w. wavemeters and high-Q circuits in oscillators, filters, etc., are discussed. An equivalent transmission line, shorted at one end and loaded by a capacitance at the other, is considered and the main relations determining the operation of the resonator are derived. From equation (5) the resonance frequencies can be determined, and formula (8) gives the Q. Using this formula the optimum configuration of a resonator can be found.

621.396.615 **359**

Wide-Range Variable-Frequency Oscillator.—A. Cormack. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 266–270.) The basic circuit of the phase-shift oscillator described comprises an amplifier stage and four cathode followers; frequency range is up to about 180 Mc/s. Bandwidths > 1 octave are obtainable; amplitude is substantially constant over most of the range. Frequency is controlled electronically.

621.396.615.029.3 **360**

Wide-Range Sweeping Oscillator.—L. A. Rosenthal. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 114–116.) A wobbled a.f. output or a steady tone of variable frequency is obtained from a Wien-bridge type of circuit including SiC nonlinear resistors. The oscillation frequency covers a range of over 20 to 1. A thermistor element in the bridge stabilizes the oscillation amplitude and ensures good sinusoidal waveform.

621.396.622.71 **361**

Numerical Treatment of the Limiting Effect of the Ratio Detector.—Behling. (See 485.)

621.396.645 **362**

Amplifiers with Slightly Variable Gain.—J. J. Zaalberg van Zelst. (*Tijdschr. ned. Radiogenoot.*, May 1951, Vol. 16, No. 3, pp. 117–133. Discussion, pp. 134–135.) Review of methods of stabilizing the gain of amplifiers, particularly the two types of circuit described in 3063 of 1947. The stability limit of amplifiers including elements which are not quite constant is discussed.

621.396.645.015.7† **363**

On Design Theory of Pulse-Amplifier.—K. Nagai. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 51–72.) The design of pulse amplifiers is described, particular attention being paid to the correction of response at high and low frequencies by inductive load compensation and RC circuits respectively. Experimental measurements on an amplifier designed for 1-μs pulses are discussed.

621.396.645.015.7† **364**

The Problem of the 'Best' Pulse Receiver.—L. Huber & K. Rawer. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, p. 254.) Correction to paper noted in 1866 of 1951.

621.396.645.029.3 **365**

Audio Amplifier Damping.—R. M. Mitchell. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 128–131.) Oscillations generated in the load of an amplifier can be damped out by controlling the output impedance by means of feedback. Methods of measuring the damping factor, defined as the ratio of the load impedance to the effective generator impedance, are described and its values for beam-tetrode and triode power valves are compared. Tests on the Williamson Type-W20 amplifier are described to illustrate the practical application of feedback.

621.396.645.37 **366**

Universal Direct-Coupled Differential Amplifier.—L. Goldberg. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 128–131.) Analysis is presented for a basic circuit capable of providing constant closed-loop gain and low output impedance, with or without sign inversion, for applications requiring a high-gain differential amplifier. Mathematical operations of many kinds, which can be effected with the aid of suitable external circuits, are summarized.

621.396.645.37 : 621.3.016.35 **367**

The Stability Problem in Feedback Amplifiers.—W. A. Lynch. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1000–1008.) The criterion for stability is formulated in terms of fundamental restrictions on amplifier characteristics at complex frequencies. From this, a simplified criterion placing equivalent restrictions on the characteristics at real frequencies leads directly to Nyquist's criterion. A survey of important relations between gain and phase shift in minimum-phase-shift networks leads to discussion of the principles by which stability can be effected.

621.396.645.371.011 **368**

The Exact Equations for Current and Voltage Negative-Feedback.—W. Reichardt. (*Frequenz*, May 1951, Vol. 5, No. 5, pp. 139–143.) Analysis of basic feedback circuits taking account of the alteration of the valve characteristic due to the addition of the feedback loop. The equations derived are of significance in a phase-reversal stage which can be regarded either as a voltage-feedback or a current-feedback circuit.

621.396.662 : 621.392 **369**

The Design of Transmission-Line Tuning Elements for Minimum Dissipation.—Klopfenstein. (See 315.)

621.397.645 **370**

R.F. Amplifier for U.H.F. Television Tuners.—B. F. Tyson & J. G. Weissman. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 106–109.) A concentric-line amplifier circuit is described which uses a disk-seal planar triode with grounded grid. The effect on gain and bandwidth of the position of the output loop along the concentric line is illustrated and the optimum position found. A gain of 12·5 db in the band 470–890 Mc/s is obtained with the amplifier ahead of a typical crystal mixer, and the power required from the local oscillator is very much reduced.

621.3.012.2 **371**

Die Ortskurventheorie der Wechselstromtechnik (The Theory of Circle Diagrams in Alternating-Current Technique). [Book Review]—G. Oberdorfer. Publishers: Deuticke, Vienna, 2nd edn 1950, 100 pp., 15 Swiss francs. (*Bull. schweiz. elektrotech. Ver.*, 16th June 1951, Vol. 42, No. 12, p. 456.) A comprehensive text-book for students and practising engineers.

621.3.016.35 : 621.3.012.2 **372**

Praktische Stabilitätsprüfung mittels Ortskurven und numerischer Verfahren (Practical Stability Testing using Circle Diagrams and Numerical Methods). [Book Review] —F. Strecker. Publishers: Springer, Berlin, 1950, 189 pp., cardboard covered 15 DM, bound 18 DM. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, p. 252.) Familiarity with modern ideas on network theory is necessary for an understanding of this book, but function theory and matrix calculations are not used.

GENERAL PHYSICS

53.05 : 518.4 **373**

The Analysis of an Experimental Curve is the Ideal Method of Determining a Physical Characteristic: Examples.—P. Vernotte. (*C. R. Acad. Sci., Paris*, 16th July 1951, Vol. 233, No. 3, pp. 230–232.)

535.422 : 538.56 **374**

On the Diffraction of Electric Waves by a Perfectly Reflecting Wedge.—Y. Nomura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 1–23.) Mathematical analysis of the diffraction of the spherical waves from an oscillating dipole by a perfectly reflecting wedge of arbitrary angle, with detailed discussion of the field in the geometrical shadow, and numerical examples.

536.491 : 621.3.032.44 **375**

Stationary Temperature of Current-Carrying Wires of Moderate Length.—J. Fischer. (*Arch. Elektrotech.*, 1951, Vol. 40, No. 3, pp. 141–171.) A comprehensive mathematical treatment for the case of a uniform wire extended between two relatively massive blocks in air or in a liquid or in vacuo. A solution for small temperature rises is first obtained and then the general solution is considered. Approximate solutions are derived and a numerical example is worked out for the case of a Pt wire, 4 cm long and 0·5 mm in diameter, carrying a current of 10 A in air.

537.311.31 : 539.23 **376**

Tentative Interpretation of the Mechanism of Electrical Conduction in Very Thin Metal Films.—N. Mostovetch. (*C. R. Acad. Sci., Paris*, 30th July 1951, Vol. 233, No. 5, pp. 360–364.)

537.311.33 **377**

Internal Barriers in Semiconductors.—H. K. Henisch. (*Phil. Mag.*, July 1951, Vol. 42, No. 330, pp. 734–738.) "The temperature dependence of conductivity is calculated for a semiconducting specimen containing internal barriers of various heights. This leads to a new interpretation of activation energies as deduced from conduction measurements."

537.525 **378**

High-Frequency Discharges: Part 1—Breakdown Mechanism and Similarity Relationship.—F. L. Jones & G. D. Morgan. (*Proc. phys. Soc.*, 1st July 1951, Vol. 64, No. 379B, pp. 560–573.) An experimental investigation covering the range of frequencies $f = 3{\cdot}5$–70 Mc/s is reported. Geometrically similar systems break down at the same potential when parameters ap and f/p are constant, a being a linear dimension and p the gas pressure. For small electron-cloud oscillations, breakdown between coaxial cylinders depends only on the inner cylinder radius. The significance of the similarity relation is discussed.

537.525 **379**

High Frequency Discharges: Part 2—Similarity Relationship for Minimum Maintenance Potentials.—F. L. Jones & G. D. Morgan. (*Proc. phys. Soc.*, 1st July 1951, Vol. 64, No. 379B, pp. 574–578.) The similarity relation applying to breakdown (see 378 above) applies also to the maintenance of h.f. currents, but the dependence of the minimum maintenance potential on the parameters $a\,p$ and f/p is not so well defined, due to discharge instability.

537.525.6 **380**

The Ion-Surge Discharge, a Cause of Arc-Back in the Mercury-Vapour Rectifier.—T. Wasserrab. (*Arch. Elektrotech.*, 1951, Vol. 40, No. 3, pp. 171–177.)

537.562 **381**

Theory of Ionized Media with Translational Symmetry.—M. Hoyaux. (*Rev. gén. Élect.*, July & Aug. 1951, Vol. 60, Nos. 7 & 8, pp. 279–291 & 317–328.) The analysis presented was made in relation to mercury-vapour-rectifier arcs, but is of more general application. The statistical properties of ionized media are studied; generalized equations of mobility are derived and extended to the case of bipolarized media. An experimental verification of the theoretical results was made using Ledrus' plasmograph.

537.562 : 537.311.37 **382**

Conductivity of Ionized Air in a High-Frequency Alternating Field.—A. Székely. (*Acta phys. austriaca*, June 1949, Vol. 3, No. 1, pp. 22–37.) Measurements at pressures down to 0·1 mm Hg and frequencies of 3–5 Mc/s show that with sufficiently low measurement voltages the conductivity increases considerably with increased frequency. The corresponding increase of the dielectric constant is only slight.

538.56 : 537.525.6 **383**

Longitudinal and Transverse Electrical Waves in Homogeneous Moving Plasma.—W. O. Schumann. (*Z. angew. Phys.*, May 1951, Vol. 3, No. 5, pp. 178–181.) For an e.m. oscillation propagated in the plasma in the direction of its motion, the charge density and velocity of the two types of wave generated are simply determined by using the relativistic transformation from the stationary state. For this method to be applicable, both plasma and phase velocity must be sufficiently low. See also 717 of 1951.

538.652 **384**

Magnetostriction in a Rotating Field.—T. Delourmel & A. Herpin. (*C. R. Acad. Sci., Paris*, 16th July 1951, Vol. 233, No. 3, pp. 239–241.) Theory previously developed for the case of a magnetizing field in a given direction is extended to cover the case of magnetization by a transverse rotating field, the distribution of magnetization of the individual domains being then asymmetrical and maximum distortion occurring in a direction at an angle to the mean direction of magnetization. This theory is illustrated by the Wiedemann effect, and explains the sensitivity of the effect to mechanical influences.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 621.396.9 **385**

The Summer Daytime Meteor Streams of 1949 and 1950: Part 1 — Measurement of the Radiant Positions and Activity.—A. Aspinall & G. S. Hawkins. (*Mon. Not. R. astr. Soc.*, 1951, Vol. 111, No. 1, pp. 18–25.) Radio echo observation of summer daytime meteor streams during 1949 and 1950 have been made with improved continuously recording equipment. Simultaneous photographic records were obtained using two narrow-beam aerials directed along different azimuths. From these observations the right ascension and declination of a major stream may be determined to within $\pm 1\cdot5°$. Three permanent streams, seen each year, have been identified; other streams of a sporadic or long-period character were observed. Part 2: 386 below (Davies & Greenhow). Part 3: 387 below (Almond).

523.5 : 621.396.9 **386**

The Summer Daytime Meteor Streams of 1949 and 1950: Part 2 — Measurement of the Velocities.—J. G. Davies & J. S. Greenhow. (*Mon. Not. R. astr. Soc.*, 1951, Vol. 111, No. 1, pp. 26–36.) Observations of meteor velocities made simultaneously with the directional observations described in part 1 are discussed and tables and histograms given. The velocities of meteors belonging to each stream have a considerable spread. Part 1: 385 above (Aspinall & Hawkins). Part 3: 387 below (Almond).

523.5 : 621.396.9 **387**

The Summer Daytime Meteor Streams of 1949 and 1950: Part 3 — Computation of the Orbits.—M. Almond. (*Mon. Not. R. astr. Soc.*, 1951, Vol. 111, No. 1, pp. 37–44.) The orbits of four daytime meteor streams have been computed from the observational data given in parts 1 and 2. The orbit of the main day-time stream indicates an association with a known visual night-time stream, while another orbit appears to be coincident with that of Encke's comet. Part 1: 385 above (Aspinall & Hawkins). Part 2: 386 above (Davies & Greenhow).

523.746 **388**

Forecasting Sunspot Variations to 1957.—A. F. Wilkins. (*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, pp. 298–299.) "Published forecasts of the variation of 12-month running averages of relative sunspot numbers for the period succeeding the 1947 maximum and up to the next maximum are presented. Although the trend of sunspot number in the immediate future has been obscured by an unexpected variation in the observed values following the 1947 maximum, it is considered that the forecasts should be suitable for the purposes of planning radio-communication services for most of the remainder of the present sunspot cycle."

523.75 : 550.38 **389**

Geomagnetic Solar-Flare Effects at Lerwick and Eskdalemuir, and Relationship with Allied Ionospheric Effects.—D. H. McIntosh. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 315–342.) Geomagnetic data for the period 1936–1949 are tabulated, and annual and daily variations are discussed in relation to solar flares and s.w. radio fade-outs.

523.854 : 621.396.822 **390**

The Origin of Galactic Radio-Frequency Radiation.—J. H. Piddington. (*Mon. Not. R. astr. Soc.*, 1951, Vol. 111, No. 1, pp. 45–63.) "The results of observations of the intensity and distribution of radio-frequency radiation from the galaxy at frequencies from 9·5 to 3 000 Mc/s have been collected. Some of these data are used to determine spectrum curves of the radiation from chosen regions of the galaxy. Using the equations defining the propagation of radio waves in an ionized gas, the forms of spectrum curves from distributions of gas are determined. It is shown that the power flux of radio energy from a thermally radiating gas cloud increases with frequency except under special conditions which are defined. From a comparison of the observed and the theoretical spectrum curves it is shown that galactic radiation at radio frequencies probably originates partly in hot ionized interstellar gas and partly in stellar atmospheres. The ionized gas provides most of the radiation at the higher frequencies and evidences itself by absorption at the lower. The properties of the stellar sources are discussed."

550.38 "1951.01/.07" **391**

Indices of Geomagnetic Activity of the Observatories Abinger (Ab), Eskdalemuir (Es) and Lerwick (Le), January to July 1951.—(*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 376–380.)

551.51 **392**

The Thermal Excitation of Atmospheric Oscillations.—M. V. Wilkes. (*Proc. roy. Soc. A*, 6th July 1951, Vol. 207, No. 1090, pp. 358–370.)

551.510.52 : 551.524.7 **393**

Mean Meridional Temperature Profile in the Troposphere as affected by Vertical and Horizontal Exchange Processes and the Heat of Condensation.—F. Defant. (*Arch. Met. Geophys. Bioklimatol. A*, 28th March 1950, Vol. 2, Nos. 2/3, pp. 184–206.)

551.510.52 : 621.396.11.029.62 **394**

The Effective Height of Tropospheric Inversion Layers in Ultra-Short-Wave Propagation.—Scholz. (See 476.)

551.510.535 **395**

Magnetic and Ionospheric Storms.—E. Appleton. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 113–119. In English.) The characteristic ionospheric-storm variation of F_2-layer ionization at a medium-latitude station is compared with corresponding effects for other latitudes. The well-known marked variability of the F_2-layer critical frequency is attributed in part to the operation of the ionospheric storm mechanism. Lack of correlation between variations of F_2 ionization at widely separated sites is due to variations of sign and magnitude of the ionospheric storm manifestation with geomagnetic latitude and to a certain dependence on local time.

551.510.535 **396**

A Critical Survey of Ionospheric Temperatures.—N. C. Gerson. (*Rep. Progr. Phys.*, 1951, Vol. 14, pp. 316–365.) The various methods available for the evaluation of temperature from an ionospheric parameter are discussed in detail, and the relative order of importance is given. The probable temperature/altitude relation is tabulated for the range 100–400 km. The electron and gas temperatures are essentially identical. Experimental evidence does not indicate a totally ionized atmosphere below 400 km. Thermal equilibrium, and therefore a Maxwellian velocity distribution, exists among the gaseous constituents from 100 km to 400 km, with no large diurnal temperature variations. It is considered that a temperature maximum, probably arising from the various absorption or heating processes together with a sufficient number density of particles, may occur in the ionosphere at about 400 km.

551.510.535 **397**

The Phase and Group Paths of Radio Waves Returned from Region E of the Ionosphere.—J. W. Findlay. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 353–366.) Equipment is described which has been used to measure the changes of phase path of radio waves of frequency 2·4 Mc/s returned from region E of the ionosphere. The daily variation of phase path cannot be explained quantitatively on the assumption that region E behaves as a simple Chapman region, although there is qualitative agreement with the theory. The seasonal changes of noon values of the group path are studied and found to disagree with those predicted for a Chapman region. Some evidence for the existence of a solar tide in the height of region E is given.

551.510.535 **398**

The Determination of the Collision Frequency of Electrons in the Ionosphere from Observations of the Reflection Coefficient of the Abnormal E Layer.—B. H. Briggs. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 345–348.) The reflection and transmission coefficients of the E_s layer are deduced from simultaneous measurements of the amplitudes of echoes returned from the E_s and F layers. The results are consistent with the assumption of a layer about 5 km thick in which the collision frequency is about 2×10^4 per sec at a well defined height.

551.510.535 **399**

Results of Ionospheric Observations made near Adélie Land.—M. Barré & K. Rawer. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 311–314. In French.) Photographic *h'f* traces recorded 470 km from the south magnetic pole are reproduced. Echoes from all regions, and particularly from the F_2 region, are often very scattered. 'High level' E_s echoes and other unusual records are probably due to oblique-incidence reflection from ionic clouds. See also 112 and 2997 of 1951.

551.510.535 **400**

Solar Tides in the Ionosphere over Calcutta.—A. P. Mitra. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 286–295.) The data for the F_2 region over the four-year period January 1946–January 1950 are examined, using an extension of Martyn's analysis which includes the effects of both recombination and vertical transport of ions. Numerical values of the resultant vertical drift velocity and of the recombination coefficients at Calcutta are deduced; the anomalous changes in F_2-layer ionization are explained.

551.510.535 **401**

Effects of Temperature Variations of the Upper Atmosphere on the Formation of Ionospheric Layers: F_2 Layer Anomaly: F_2 and F_1 Separation: F_2 Layer Behaviour during Magnetic Storms.—D. Lepechinsky. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 278–285.) An increase in temperature (T) causes a rise in the height of a Chapman layer, accompanied by a fall in the rate of ion production (I). Assuming a diurnal or seasonal temperature cycle, the I curve at constant height may show a maximum or a minimum at the time when T is highest, depending on the height chosen. The changes in height of the F_2 layer at Slough can be explained by a 46% rise in T at summer noon as compared with winter noon. The corresponding rise between 0600 and noon in summer would be 21%. The splitting of the F layer into F_1 and F_2 layers and certain storm characteristics can also be explained by this theory. See also 2995 of 1951 (Lejay & Lepechinsky).

551.510.535 **402**

Studies of the Tidal Effect in the Ionosphere.—O. Burkard. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 349–352. In German.) The F_2-layer critical frequency at Huancayo exhibits a luni-solar variation of the same nature as the variations of the geomagnetic components. The semidiurnal lunar variation is pronounced only during the daytime, and is practically nonexistent during the night.

551.510.535 : 537.562 **403**

Plasma Resonance in Ionospheric Irregularities.—N. Herlofson. (*Ark. Fys.*, 1951, Vol. 3, No. 15, pp. 247–297. Reprint.) A critical review of the experimental and theoretical deductions from meteor data. A wave theory of reflection is developed which includes the theory of Lovell & Blackett as a limiting case. The effects of induced currents are fully discussed and expressions for the scattering cross-section as a function of radius/wavelength are deduced for uniform, parabolic, exponential, and Gaussian distributions of electron density. The physical causes of the differences between parallel and sagittal scattering are discussed in detail and the conditions under which the scattering power can be greatly modified by plasma oscillations are found. Plasma oscillations can also be generated by small irregularities in the ionosphere at levels near the point of reflection of radio waves. The presence of these oscillations should be detectable experimentally. Suggestions for testing the full theory experimentally are included.

551.510.535 : 621.396.11 **404**

An Investigation of Sudden Radio Fadeouts on a Frequency near 2 Mc/s.—J. W. Findlay. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 367–375.) Experiments to determine the height at which the extra ionization occurs during fade-outs are described; observations were made of phase-path variations during 105 fade-outs. The collision frequency at the height of greatest extra ionization is estimated to be of the order of 4×10^5 per sec, the corresponding height being < 101 km.

551.510.535 : 621.396.6 **405**

Echo Sounding Equipment for Ionosphere and Weather Research.—R. Schrott. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 109–112.) Description of pulse transmitter and receiver developed at Graz for measurement of ionosphere critical frequencies and layer heights, etc.

551.510.535 : 621.396.822 **406**

Ionospheric Thermal Radiation at Radio Frequencies.—J. L. Pawsey, L. L. McCready & F. F. Gardner. (*J. atmos. terr. Phys.*, 1951, Vol. 1, Nos. 5/6, pp. 261–277.) Radio noise on medium frequencies includes a component due to thermal radiation from the ionosphere. Methods used to identify and measure this radiation are described; observations are made far from towns and during midday hours, on frequencies between 1 and 2 Mc/s, so as to avoid both man-made noise and atmospherics as far as possible. The magnitude of the thermal radiation observed corresponds to an effective temperature of about

300° K, or a field strength of about 10^{-9} V/m in a frequency band of 1 kc/s. The method is useful for studying the electron temperature in the absorbing (and emitting) region of the ionosphere, which is at a height of 70–80 km for the frequencies used. The measured temperatures, 240°–290° K, agree with other observations of temperatures at these heights.

551.594 **407**

Studies of the Atmospheric Potential Gradient: No. 4—Current Theory of the Electric Field in the Air.—H. W. Kasemir. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 84–97.) Benndorf's theory, according to which atmospheric electric currents are due to meteorological phenomena, is applied particularly to investigate the thunderstorm as energy source for these currents.

551.594 **408**

Air-Earth Current Observations in Various Localities.—A. R. Hogg. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 40–55. In English.) Observations at a number of stations indicate a statistical relation between the conductivity of the atmosphere near the ground and the air-earth conduction current. An estimate of the annual variation of air-earth current over the whole globe gives general support to the thunderstorm theory of the maintenance of the earth's electric charge.

551.594 **409**

The Development of Fundamental Ideas on Atmospheric Electricity.—H. Israël. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 1–16.)

551.594.12 **410**

Variations of Ion Concentration near the Ground.—H. Norinder & R. Siksna. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 29–39.) Measurements made at Uppsala indicate time variation of ion concentration unrelated to meteorological phenomena. In some cases they may be due to accumulation of radioactive substances near the ground.

551.594.13 **411**

The Control of the Electrical Conductivity of the Lower Atmosphere.—J. J. Nolan. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 17–28. In English.) Methods are described for counting the condensation nuclei and for determining their sizes and masses. The difficulty of establishing a simple relation between atmospheric conductivity, rate of ion production and nucleus concentration is enhanced by the tendency to an increase of airborne radioactive matter with increase of nucleus concentration.

551.594.14 **412**

New Studies on the Radioactivity of the Atmosphere.—V. F. Hess. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 56–63. In English.) Report of methods used and values found in an experimental investigation of the ionization due to the γ radiation from radioactive substances near the ground, and of the total ionization produced by these substances.

551.594.21 **413**

The Distribution of Electricity in Thunderclouds.—B. F. J. Schonland & D. J. Malan. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 64–69. In English.)

551.594.22 **414**

Measurements by Airplane of Electric Charge passing Vertically through Thunderstorms to Ground.—G. R. Wait. (*Arch. Met. Geoph. Bioklimatol., A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 70–76. In English.)

551.594.6 **415**

Progress in World Radio Meteorology and Weather Prediction.—J. Lugeon. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 120–138. In French.) Investigations of the relations between atmospherics and synoptical meteorology from 1920 onwards are summarized. An account is given of up-to-date methods of observation, and some results are discussed. Lines are indicated for the organization of radio-meteorological networks.

551.594.6 **416**

Analysis of Audio-Frequency Atmospherics.—R. K. Potter. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1067–1069.) Audio-frequency atmospherics have been reported by long-wave-radio and ocean-cable engineers and are known as whistlers, swishes, tweeks and rumblers. These have been studied by means of the sound spectrograph which displays a graph of frequency against time. The different types of frequency and overtone variation with time are described.

LOCATION AND AIDS TO NAVIGATION

621.396.9 : 523.5 **417**

The Summer Daytime Meteor Streams of 1949 and 1950.—(See 385–387.)

621.396.933 **418**

Miniature Radar Transponder Beacon.—R. S. Butts. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 104–107.) Description of aircraft equipment which, on receipt of radar signals, automatically transmits a reply to the interrogating station, thus extending the radar range obtainable with passive reflection alone and making possible the determination of aircraft azimuth as well as range. The equipment weighs under 4 lb and comprises receiver, temperature-compensated coaxial cavity resonator, and transmitter with pulse modulator. Potting of receiver and modulator in resilient casting resin reduces shock effects.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 : 535.61-31 **419**

Application of Optical Absorption in the Far Ultraviolet to the Detection of Leaks in Vacuum Apparatus and to the Measurement of Low Pressures.—J. Romand, V. Schwetzoff & B. Vodar. (*Le Vide*, July/Sept. 1951, Vol. 6, Nos. 34/35, pp. 1046–1051.) Devices are described in which a beam of ultraviolet rays, of wavelength in the range 1 200–1 850 Å, traverses the gas under examination before impinging on a photoelectric target. Preliminary test results are presented.

535.215 **420**

Energy Levels of F-Centres.—L. Pincherle. (*Proc. phys. Soc.*, 1st July 1951, Vol. 64, No. 379A, pp. 648–657.) The possible association between the existence of F centres and photoconductivity in ionic crystals is investigated.

535.37 **421**

Inhibiting Action of Iron on the Luminescence of Zinc Sulphide.—N. Arpiarian. (*C. R. Acad. Sci., Paris*, 30th July 1951, Vol. 233, No. 5, pp. 387–389.) Specimens of ZnS containing known amounts of Cu activator and of Fe were exposed to ultraviolet light at a temperature of 20°C. Curves are given showing the decay of the luminescence at various time intervals after removal of the excitation, for various compositions. The results are compared with those obtained using Ni or Co as inhibitor, and the theory of the effect is discussed.

537.228.1 : 548.0 : 547.476.3 **422**

Theory and Measurement of the Piezoelectricity of Rochelle Salt.—S. Honda. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 117–140.) In order to study the characteristics and applications of crystals such as Rochelle salt it is convenient to use the polarization theory. The general piezoelectric equations according to this theory are deduced. The electric, elastic and piezoelectric constants of Rochelle-salt X-cut plates were measured, the last by a newly developed direct dynamic method.

537.311.33 **423**

Change of Activation Energy with Impurity Concentration in Semiconductors.—L. Pincherle. (*Proc. phys. Soc.*, 1st July 1951, Vol. 64, No. 379A, pp. 663–664.) Analysis of the effect of screening by free carriers of the field around a trapping centre, resulting in a potential of the form $V = (e^2/\epsilon r) \exp(-r/r_0)$. The case of Si is considered quantitatively for various impurity concentrations and temperatures; comparison of theoretical and experimental values of activation energy suggests that although screening may not be the main cause of the variation it can be significant.

537.311.33 **424**

The Theory of Contact between Two Semiconductors of the Same Type of Conductivity.—A. I. Gubanov. (*Zh. tekh. Fiz.*, March 1951, Vol. 21, No. 3, pp. 304–315.) The theory of the rectifying effect of the contact between a metal and a semiconductor is extended to cover the case of two semiconductors with the same type of conductivity. The effect in the latter case is smaller, for the same contact potential difference, than that of the contact between a metal and a semiconductor. It is particularly small with a large current, when, under certain conditions, rectification in the opposite direction may also appear. In the limiting case when the conductivity of one of the semiconductors is considerably greater than that of the other, all the formulae of the proposed theory are reduced to the well known formulae for the case of the contact between a metal and a semiconductor. This enables the limits of applicability of the theory to be established.

537.311.33 : 546.816.221 **425**

Lead Sulphide — an Intrinsic Semiconductor.—E. H. Putley & J. B. Arthur. (*Proc. phys. Soc.*, 1st July 1951, Vol. 64, No. 379B, pp. 616–618.) Measurements of the Hall coefficient and conductivity of PbS over wide ranges of temperature and effective purity show that *p*-type and *n*-type specimens give the same results at sufficiently high temperatures. PbS shows intrinsic conductivity with an energy gap of about 1·2 eV, resembling PbTe.

538.221 **426**

Torque Curves and other Magnetic Properties of Alcomax.—K. Hoselitz & M. McCaig. (*Proc. phys. Soc.*, 1st July 1951, Vol. 64, No. 379B, pp. 549–559.)

538.221 **427**

Study of Macroscopic Magnetic Texture.—I. Épelboin. (*C. R. Acad. Sci., Paris*, 30th July 1951, Vol. 233, No. 5, pp. 358–360.) It is possible by thermal and electrolytic treatments to influence separately the various features determining the macroscopic magnetic texture, even of relatively thick specimens of high-permeability alloys. This affords a method of investigating the composition and the magnetic behaviour of such alloys in weak alternating fields.

621.316.993 **428**

Impulse Resistance of Different Earth Electrodes and Embedding Materials.—H. Norinder & O. Salka. (*Bull. schweiz. elektrotech. Ver.*, 19th May 1951, Vol. 42, No. 10, pp. 321–327. In German.)

621.318.4.042.15 **429**

Loss Angle of Manganese-Ferrite Cores.—A. Weis. (*Elektrotechnik, Berlin*, May 1951, Vol. 5, No. 5, pp. 213–216.) Curves are given showing the loss angle of Mn-ferrite and MnZn-ferrite as dependent on (*a*) frequency to 100 kc/s, (*b*) field strength to 90 millioersted. An ultrasonic resonance effect was observed in a Mn-ferrite ring core at 44·5 kc/s. This was investigated both with and without an e.s. screen round the core.

621.396.611.21 : 549.514.51 **430**

Crystal-Plates without Overtones.—J. J. Vormer. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1086–1087.) For a similar account see 1169 of 1951.

MATHEMATICS

517.35 : 621.319.4 **431**

A Solution for $\int e^{b(x + a\cos x)}dx$.—K. Emden: H. Rutishauser. (*Z. angew. Math. Phys.*, 15th July 1951, Vol. 2, No. 4, pp. 289–293.) The integral occurs in the expression for the charge on a capacitor, in series with a resistor and battery, when the plate distance is varied periodically. The integral $\int e^{b(y + a\sin y)}dy$ is easily derived by the substitution $y = x + \pi/2$. Rutishauser gives a simpler method of determining the Fourier coefficients in the solution.

517.432.1 : 621.392.5 **432**

Time-Dependent Heaviside Operators.—L. A. Zadeh. (*J. Math. Phys.*, July 1951, Vol. 30, No. 2, pp. 73–78.) Discussion of some of the basic properties. See also 1617 of 1950.

517.512.2 **433**

A Series Expansion of the Fourier Integral.—W. A. Whitcraft, Jr. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1087–1089.) "Evaluation of the Fourier integral by parts leads to two different series representations of a time function in the frequency domain. These series involve powers of the frequency variable and either derivatives or integrals of the function, evaluated at the upper time limit only."

681.142 : 531.764.5 **434**

A Digital-Computer Timing Unit.—R. M. Goodman. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1051–1054.) Description of the time-reference equipment for the EDVAC computer, and discussion of the application of electronic matrices to time-pulse production.

501 : 517.944.3 **435**

Integral Transforms in Mathematical Physics. [Book Review]—C. J. Tranter. Publishers: Methuen, London and J. Wiley & Sons, New York, 1951, 118 pp., 5s. (*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, p. 315.) One of Methuen's Monographs on Physical Subjects. "The object has been to show the reader what techniques are worth trying for the solution of various types of problem."

517.516 **436**

Fastperiodische Funktionen (Nearly Periodic Functions). [Book Review]—W. Maak. Publishers: Springer-Verlag, Berlin, 1950, 240 pp., 24.60 DM. (*Z. angew. Math. Phys.*, 15th July 1951, Vol. 2, No. 4, pp. 306–307.) Concerned exclusively with the generalizations of Bohr's theory of nearly periodic functions.

MEASUREMENTS AND TEST GEAR

621.317.001.5 : 519.2 437

An Outline of the Principles of Design and Analysis of Experiments.—D. L. Richards. (*P.O. elect. Engrs' J.*, July 1951, Vol. 44, Part 2, pp. 55–60.) Discussion of concepts underlying statistical techniques used at the P.O. Research Station in connection with investigations involving human subjects.

621.317.3 : 550.372 438

Ground Conductivity Measurements in Italy.—G. Galligioni. (*Alta Frequenza*, June/Aug. 1951, Vol. 20, Nos. 3/4, pp. 119–127.) A conductivity map of the whole of Italy has been prepared by Radio Italiana, based on about 3 000 measurements of the field strength of Italian broadcasting stations. The method is described and the reliability of the results is discussed; they can be used to predict field strength for different sites, frequencies, power values and aerial types.

621.317.3 : 621.392.52 439

Fundamental-Wave Filter for Instruments and Measurement Circuits.—Poleck. (See 352.)

621.317.321 440

The Measurement of High Alternating Voltages by means of Capacitive Voltage Dividers (C-Measurement).—P. Böning. (*Arch. tech. Messen*, May 1951, No. 184, pp. T49–T52.) The development of the basic circuits particularly considered is the insertion of an inductor in the meter circuit connected across one of the two capacitors comprising the divider. Three conditions, depending on choice of the value of the inductor, are considered and vector diagrams showing voltage and current distribution are constructed to determine the optimum values of the components in the measurement circuit.

621.317.335.3† : 621.317.755 441

Investigation of Dielectrics by Means of a Cathode Ray Oscillograph.—N. S. Novosil'tsev. (*Zh. tekh. Fiz.*, March 1951, Vol. 21, No. 3, pp. 369–374.) A description is given of an oscillograph with which the dielectric constant and leakage can be determined for substances with a single relaxation time. With a little practice it is possible to determine directly from the image on the screen the amount of absorption, the presence of several mechanisms and the speed of their operation, and the magnitude of losses. The effects of temperature and humidity on the dielectric properties are shown very clearly.

621.317.335.3.029.64† 442

New Method for Determination of Dielectric Constants in the cm-Wave Range.—E. Ledinegg & E. Fehrer. (*Acta phys. austriaca*, June 1949, Vol. 3, No. 1, pp. 82–110.) A comprehensive mathematical treatment of the method in which a plate of the material to be tested is inserted in a cylindrical cavity resonator. Field contours for fundamental and harmonic resonance are shown graphically and useful approximate formulae are given.

621.317.35 443

The Experimental Representation of Modulated Currents and their Comparative Examination by means of a Simple Test Assembly.—M. Kreuzritter. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, pp. 320–324.) Description of multi-purpose instrument for analysis of a.m. and f.m. signals. It is mains-operated and incorporates an electromechanical modulation device and a vibrating-reed indicator.

621.317.7 444

Moving-Coil Meters. External and Core-Type Magnetic System. Pivot Bearing and Tension-Strip Suspension.—A. Hug. (*Bull. schweiz. elektrotech. Ver.*, 2nd June 1951, Vol. 42, No. 11, pp. 385–389. In German.) Practical requirements are discussed and the advantages and disadvantages of the different types are compared. The possibility of combining their good features is briefly considered.

621.317.73 445

Impedance Measurements in the 50–2 000 Mc/s Range.—R. A. Soderman. (*Radio & Televis. News, Radio-Electronic Engng Section*, July 1951, Vol. 46, No. 1, pp. 3–6, 25.) A direct-reading admittance meter which can be balanced like a bridge and is suitable for the frequency range 50–1 000 Mc/s comprises a T arrangement of three coaxial lines fed at their common junction. When used together with a special constant-impedance adjustable-length line, the upper frequency limit is raised to 2 000 Mc/s.

621.317.733 446

The Wheatstone Bridge with Load-Dependent Resistances: Part 1—Fundamentals.—G. Nidetzky. (*Arch. tech. Messen*, May 1951, No. 184, pp. T55–T56.) Analysis of the effect of input voltage fluctuations in bridge circuits including resistors such as gas-filled tungsten-filament lamps, iron-in-hydrogen barretters, carbon-filament lamps, semiconductors, and gas discharge lamps.

621.317.75 : 621.397.5 447

Television Waveform Display.—K. R. Sturley. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 261–265.) The construction of a simple unit for the visual examination of television waveforms is described. The 50-c/s mains supply is used as a driver source, and 1 to 20 lines of both frames may be selected for separate and simultaneous display. Slow drift of the waveform, due to a.f.c. of the frame-synchronizing pulse at the transmitter, can be corrected manually; or the frame-scan generator of the receiver may be used to provide the driver source. The required additional equipment is described.

621.317.755 : 621.396.67.012 448

Oscillograph Field Plotter.—C. Susskind & A. R. Perrins. (*Electronics*, Sept. 1951, Vol. 24, No. 9, p. 140.) The position of a probe moving in the field of a microwave radiator is correlated by a potentiometer arrangement with that of the spot on the screen of a c.r.o. The intensity of the spot is controlled by the signal picked up by the probe and the field pattern is recorded photographically with a time exposure.

621.317.784 449

Power Meter and Mismatch Indicator.—A. F. Boff. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 278–281.) A simple directional power meter is described for the frequency band 150–250 Mc/s. The instrument is based on a directional coupler consisting of a loop inserted into a short coaxial line and loosely coupled to both the electric and magnetic fields in such a way that components cancel for waves in one direction and add in the other direction. Forward and reflected power in a transmission line are indicated simultaneously on two meters reading up to 10 W, with negligible loading of the line.

621.319.4(083.74) 450

R.F. Standard Capacitors for Minute Increments.—J. A. Conner. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 250–262.) The errors inherent in coaxial-cylinder capacitors are analysed, with emphasis on the design of incremental standards with variations of less than 10 pF per inch. Design details are given for a differential standard with a linear capacitance variation of 0·2 pF per inch of inner-electrode insertion, accurate to within 1% over the range of movement and with a residual uncertainty of only 0·000 15 pF.

621.396.615.029.3 451
Wide-Range Sweeping Oscillator.—Rosenthal. (See 360.)

621.396.677.5.089.6† 452
Checking Calibration of Loop Antennas.—P. C. Gardiner. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 140..250.) Description of a laboratory method of calibrating loop aerials for field-strength meters, in which the calibrating field is provided by a signal generator feeding a loop through a high impedance. This loop is concentric and coplanar with the receiving loop to be calibrated, has a smaller radius, and should have a smaller inductance. The number of turns in the receiving loop must be known and the mutual inductance between the loops must be calculated or measured. Design data are discussed.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

612.8 : 621.39.001.11 453
A Particular Problem of Signal Transmission [in the nervous system].—A Goudot. (*C. R. Acad. Sci., Paris*, 23rd July 1951, Vol. 233, No. 4, pp. 290–292.) The nervous system is compared with normal signalling circuits; various analogies exist. Selective transmission by certain nerve fibres is compared with propagation in waveguides.

620.179.14 454
Crawler detects Gun-Barrel Cracks.—R. D. Kodis & R. Shaw. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 92–95.) After the gun barrel has been circularly magnetized, its inner surface is scanned by a pickup coil that rotates at 900 or 1 800 r.p.m. round a central core. A servo system keeps the motion of the coil in step with the recording equipment.

620.179.16 455
Metal Wall Thickness Measurement from One Side by the Ultrasonic Method.—N. G. Branson. (*Elect. Engng, N.Y.*, July 1951, Vol. 70, No. 7, pp. 619–623.) 1951 A.I.E.E. Winter General Meeting paper. The method is based on setting up standing waves in the metal; the thickness is deduced from the measurement of two frequencies at which the metal is in thickness resonance. The circuit of a practical instrument, normally using frequencies in the range 1–20 Mc/s is given. Limitations of the method are discussed.

621.316.7.076.7 456
The Electro-analogue, an Apparatus for Studying Regulating Systems: Part 2—The Electrical Execution.—J. M. L. Janssen & L. Ensing. (*Philips tech. Rev.*, May 1951, Vol. 12, No. 11, pp. 319–335.) Detailed discussion of various parts, including the process analogues, the universal amplifier, the combining stages, the integrator and differentiator, models of discontinuously acting controllers, the oscilloscope, the supply and the mechanical construction. Part 1: 2782 of 1951.

621.365.54/.55† 457
Economics and Applications of High-Frequency Heating.—G. Lang. (*Bull. schweiz. elektrotech. Ver.*, 19th May 1951, Vol. 42, No. 10, pp. 328–342. In German.) Illustrated review of development of industrial heating techniques, with extensive bibliography.

621.365.55† : 621.396.615.141.2 458
Magnetrons for Dielectric Heating.—Nelson. (See 556.)

621.385.833 459
Resolving Power and Magnification Limit of the Magnetic Electron-Microscope as dependent on the Accelerating Voltage and Magnetic Field Strength.—W. Glaser. (*Acta phys. austriaca*, June 1949, Vol. 3, No. 1, pp. 38–51.)

621.385.833 460
New Formulae for Third-Order Aberrations in Electrostatic Lenses.—P. Sturrock. (*C. R. Acad. Sci., Paris* 16th July 1951, Vol. 233, No. 3, pp. 243–245.)

621.385.833 461
The Electrostatic Lens as Velocity Filter with High Resolving Power.—G. Möllenstedt & O. Rang. (*Z. angew. Phys.*, May 1951, Vol. 3, No. 5, pp. 187–189.)

621.385.833 462
Graphical Method for Approximate Determination of [charge] **Carrier Paths in Electrostatic Lenses, taking account of the Effects of Space Charge.**—W. Walcher. (*Z. angew. Phys.*, May 1951, Vol. 3, No. 5, pp. 189–190.)

621.385.833 463
Calculation of the Optimum Focusing of a Beam by an Electron Lens, taking account of Space Charge.—M. D. Gabovich. (*Zh. tekh. Fiz.*, March 1951, Vol. 21, No. 3, pp. 363–368.)

621.387.424† 464
A Directive Effect in Geiger-Müller Counters.—A. Rogozinski. (*C. R. Acad. Sci., Paris*, 30th July 1951, Vol. 233, No. 5, pp. 426–428.) The rate of rise of the pulse in a G-M counter is most rapid when the angle between the trajectory of the discharging particle and the axis of the counter is zero. Thus a counter circuit which selects the most sharply rising pulses will register only particles travelling parallel to the counter axis.

621.387.424† 465
Factors affecting the Stability of Geiger-Müller Counters with Thin End Window (Argon-Alcohol-Filled Bell-Shaped Counters).—A. Papineau. (*J. Phys. Radium*, June 1951, Vol. 12, No. 6, pp. 667–670.)

621.398 : 621.315 466
Electronic Equipment for Telecontrol in High-Voltage Networks.—F. Burlando. (*Bull. schweiz. elektrotech. Ver.*, 19th May 1951, Vol. 42, No. 10, pp. 352–356. In French.) Description of a system in which a 900-VA Wien-bridge oscillator is inductively coupled to a 25-MVA line, thus superimposing a control signal, variable from 50 to 100 kc/s, on the mains frequency. The oscillator output is automatically controlled by a thyratron circuit suitably adjusted to the impedance of the line.

PROPAGATION OF WAVES

538.566 467
On the Propagation of Electric Waves from a Horizontal Dipole over the Surface of the Earth Sphere.—Y. Nomura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, Jan. 1951, Vol. 1/2, No. 1, pp. 25–49.) The electromagnetic field generated by a horizontal electric or magnetic dipole on or over a finitely conducting spherical surface is considered and formulae for the magnitude of each of the spherical-coordinate components of the electric and magnetic forces are obtained. The field of the horizontal dipole is calculated from that of the vertical dipole. An approximate expression based on optical ray theory is also given.

538.566 **468**

Surface Waves 'ad infinitum'.—H. Ott. (*Arch. elekt. Übertragung*, July 1951, Vol. 5, No. 7, pp. 343–346.) A reply to criticism of the author's theory by Kahan & Eckart (2517 of 1951), pointing out that in their treatment they have extended to all values of ± and any loss angle a result which is only valid for propagation over an absorption-free earth as $r \rightarrow \infty$. With correct treatment, results previously given by the author are confirmed.

538.566 **469**

Surface Wave in Dipole Radiation over Plane Earth.—T. Kahan & G. Eckart. (*Arch. elekt. Übertragung*, July 1951, Vol. 5, No. 7, pp. 347–348.) A reply to Ott (468 above) maintaining the correctness of their conclusions, which are shown to be confirmed by the results of several other investigators.

621.396.11 **470**

Investigation of Ionospheric Propagation and R.F. Interaction by means of Electrical Models.—M. Carlevaro. (*Alta Frequenza*, Aug. 1950, Vol. 19, No. 4, pp. 185–210.) On the basis of the well-known analogy between the propagation of plane e.m. waves in a homogeneous medium and the propagation of voltage or current waves along a transmission line, models for demonstrating propagation phenomena in ionized and magnetoactive media are constructed from transmission lines or symmetrical ladder networks. The limits of validity of the method are examined, and various examples are presented. A detailed description is given of a proposed ladder network for demonstrating the Italian experiments on gyro-interaction carried out during the last three years. The method may be useful in eliminating algebraic calculations not only in ionospheric propagation problems but also in the study of electromagnetic screens.

621.396.11 **471**

The Effective Velocity of Propagation of Short Radio Waves.—J. Fuchs. (*Arch. Met. Geoph. Bioklimatol. A*, 15th Nov. 1950, Vol. 3, Nos. 1/2, pp. 139–152.) For making the necessary corrections to time signals transmitted by short waves, their propagation time between transmitter and receiver must be known accurately. A theory is developed according to which the effective propagation velocity v_{eff} is a function of angle of incidence and apparent height of reflection; values of v_{eff} corresponding to angles of 30°, 20°, 10° and 0° are respectively 246 000 km/s, 268 000 km/s, 281 000 km/s and 290 000 km/s. The principles of methods used for calculating v_{eff} for any distances and ionization conditions are discussed with the aid of world maps of critical frequencies, and a new method of calculating apparent height of reflection is proposed.

621.396.11 **472**

Wave Propagation in the Ionosphere at Oblique Incidence, taking account of the Earth's Magnetic Field: Part 2.—K. Försterling. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, pp. 209–215.) Continuation of 3518 of 1949. The equations of propagation are derived and discussed for large angles of incidence, assuming dielectric constant to be a known function of height.

621.396.11 : 551.510.535 **473**

Time-Delay Measurements on Radio Transmissions. Results on Medium Frequencies.—R. Naismith & E. N. Bramley. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 271–277.) The time delay of the first ionospheric echo with respect to the ground wave, and the variation of this delay, were measured at distances from zero up to 1 200 km, at frequencies between 0·7 and 2·0 Mc/s. The equivalent height of the main reflecting region was in the range 90–97 km, but other regions at 70–76 km, 105–110 km and 120–130 km were observed under certain conditions. These results indicate that stratification and partial-reflection phenomena have an important influence on oblique propagation at medium frequencies.

621.396.11 : 621.317.353.3† **474**

Ionospheric Cross-Modulation.—G. R. Mather. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 252..260.) Controlled experiments on the Luxemburg effect were carried out on five consecutive nights, using two broadcasting transmitters with frequencies around 1·1 Mc/s, the disturbing power radiated being over 100 kW. Measurements at a distance of 555 miles from the wanted transmitter, using a wave analyser in conjunction with a receiver, showed that the carrier of the wanted transmission was modulated to a depth of 0·6 to 0·75% by the unwanted transmission. This was a smaller transfer of modulation than had been expected from theory.

621.396.11.029.55 : 523.745 **475**

Variation of Angle of Arrival of Short Waves in Transatlantic Communications due to the Influence of the Sunspot Cycle.—H. Neyer & K. Rawer. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, pp. 215–218.) Angle-of-arrival measurements made in 1939–1940 by Kotowski, Schüttlöffel & Vogt (434 of 1951) and repeated in 1944 are discussed in relation to opposing views on the mechanism of long-distance propagation, viz., whether the waves are guided or suffer zig-zag reflection. The observations were made at Brück a.M. on transmissions from 15 North American stations operating on frequencies from 6·10 to 17·83 Mc/s. The observed variation of angle of arrival with frequency supported the theory of zig-zag reflections. Compared with the earlier measurements, the 1944 figures show a marked decrease of angle of arrival for a given frequency; this is due to the reduction of F_2-layer critical frequency corresponding to the changed phase of the sunspot cycle.

621.396.11.029.62 : 551.510.52 **476**

The Height of Tropospheric Inversion Layers Effective in Ultra-Short-Wave Propagation.—W. Scholz. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, pp. 287–293.) Report and discussion of measurements made from 1938 to 1945 over a 64-km path in Germany. A 50-W transmitter radiated a signal on 206 Mc/s, with 800-c/s a.m., from an aerial 135 m above ground. Received field strength 43 m above ground and temperature gradient up to a height of 240 m, measured by twelve thermometers at different heights at about the mid-point of the path, were recorded.

621.396.81 **477**

Reception of N.B.S. Standard Signals.—C. Egidi. (*Alta Frequenza*, June/Aug. 1951, Vol. 20, Nos. 3/4, pp. 113–118.) Field-strength measurements made at Turin of the WWV 10-Mc/s standard signals are reported. The measurements were made between 0700 and 0900 C.E.T. over the period 1947–1950.

621.396.11.029.6 **478**

Propagation of Short Radio Waves. [Book Review]—D. E. Kerr (Ed.). Publishers: McGraw-Hill, New York, 1951, 728 pp., $10.00. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 320..324.) M.I.T. Radiation Laboratory Series, Vol. 13. ". . . covers its field in a comprehensive fashion."

RECEPTION

519.272.15 : 621.39.001.11 : 621.396.621 **479**

Application of Correlation Analysis to the Detection of Periodic Signals in Noise.—Y. W. Lee, T. P. Cheatham, Jr, & J. B. Wiesner. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1094–1096.) Discussion on paper abstracted in 730 of 1951.

621.396.621 : 621.396.712.2 **480**

Remote-Pickup Broadcast Receiver.—A. A. Kelley. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 102–103.) A circuit diagram and details are given of a station crystal-controlled 26-Mc/s receiver for programmes from a remote transmitter. Simplicity ensures reliability and ease of operation and servicing.

621.396.621.015.7† **481**

Superregenerative Receivers for Pulse Reception.—S. Marmor. (*Ann. Télécommun.*, June 1951, Vol. 6, No. 6, pp. 170–180.) Three superregenerative receivers are described for reception of pulses of the order of 2 μs duration, on a frequency of 170 Mc/s. They illustrate alternative methods for damping the trains of oscillations of the detector valve on application of the quench voltage, and for effectively increasing the quench frequency.

621.396.621.54 **482**

Tracking of Superheterodyne Receivers.—H. S. de Koe. (*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, pp. 305–315.) Tracking frequencies are calculated for various criteria defining the best tracking-error curve. The capacitances and inductances in the oscillator circuit are then calculated by means of formulae containing dimensionless variables; wiring capacitances, etc., necessitate the use of a domain of possible combinations of inductance and capacitance represented by nearly linear curves. Calculations are illustrated by examples. A curve published by M. Wald (3313 of 1941) is shown to be wrong.

621.396.622 **483**

Characteristics of A.M. Detectors.—W. E. Babcock. (*Audio Engng*, July 1951, Vol. 35, No. 7, pp. 9–11..27.) The operation and distortion characteristics of various diode and triode detectors are discussed. The 'infinite-impedance' triode detector combines the low distortion of the diode with the high input impedance of the triode, and minimum distortion is produced even with large grid signals. A disadvantage of the infinite-impedance detector, and of all the other detectors considered except the diode, is that if a.v.c. is desired a separate channel must be used.

621.396.622 **484**

The Dynamic Characteristics of Frequency Detectors.—G. Bosse. (*Arch. elekt. Übertragung*, July 1951, Vol. 5, No. 7, pp. 314–320.) Simple approximate formulae are derived for the dynamic characteristics of the conventional phase discriminator, from which the distortion arising from transient effects can be calculated. Means of reducing even-harmonic distortion occurring with greater relative bandwidth are described. Experimental results confirm the theory.

621.396.622.71 **485**

Numerical Treatment of the Limiting Effect of the Ratio Detector.—H. Behling. (*Frequenz*, April 1951, Vol. 5, No. 4, pp. 89–97.) Making use of a graphical representation of the diode characteristics and equivalent circuits, a method is developed for determination of the amplitude limitation effected in the detector circuit. A complete analysis is presented in a numerical example showing how coupling, damping, diode load resistance, etc., may be chosen for optimum limiting effect.

621.396.662 **486**

'Physiological Volume-Control' in Broadcast Receivers.—F. Bergtold & S. Sawade. (*Telefunken Ztg*, March 1951, Vol. 24, No. 90, pp. 48–50.) Discussion of the volume adjustments for the different audio frequencies required to obtain faithful reproduction. Tests carried out on 10 observers indicated that with reduced output level, not only should the lower frequencies be boosted but the higher frequencies should be reduced in comparison with the values acceptable at a normal output level.

621.396.82 **487**

A Theoretical Investigation of the Elimination of Whistle Interference in Superheterodyne Receivers.—F. Wetzorke. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, pp. 315–319.) A method is developed for determination of the optimum i.f. for avoidance of harmonic whistles and image signals. From a classification of interfering stations as local, near, or distant according to an empirical formula, and from a series of equations derived, optimum values of i.f. can be calculated. A diagram is constructed representing the number of whistle interferences as a function of i.f. over the range 350–530 kc/s. The optimum i.f. can be found directly from inspection of the contour of the diagram.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **488**

Basis of Information Theory.—S. Malatesta. (*Alta Frequenza*, June/Aug. 1951, Vol. 20, Nos. 3/4, pp. 128–159.) The fundamental principles of modern communication theory are outlined. Essential elements of statistical calculus are presented and used to examine the time functions representing signal and noise. Quantity of information is defined; the magnitude of the quantity of information contained in a signal in the presence of noise is discussed. The generalized Hartley Law is established which relates quantity of information to signal bandwidth, signal duration and signal-power/noise-power ratio.

621.39.001.11 **489**

On the Word 'Cybernetics'.—A.A. (*Onde élect.*, May 1951, Vol. 31, No. 290, p. 257.) The word 'cybernetics' has been used before in senses quite different from that now given to it by Wiener. A passage from Plato is quoted and translated, in which κυβερνητικη clearly means 'the art of navigation'. Ampère, in his 'Essai sur la philosophie des sciences, ou exposition naturelle de toutes les connaissances humaines', published in 1834, defines cybernetics as the 'study of how it may be made possible for citizens to enjoy secure peace.'

621.394/.396 **490**

Evolution of the Technique of Long-Distance Lines during the Last 15 Years.—R. Sueur. (*Ann. Télécommun.*, June 1951, Vol. 6, No. 6, pp. 146–164.) Developments in overhead lines, underground and submarine cables and radio links are described, together with regulating apparatus, repeaters, filters and related apparatus. A comprehensive bibliography, referring mainly to the French literature, is appended.

621.395/.396].5 **491**

New Long-Distance Telephony Systems.—J. Schniedermann. (*Frequenz*, May 1951, Vol. 5, No. 5, pp. 125–137.) Analytical review of the development of Siemens & Halske equipment for the German telephone service. Block diagrams of cable and radio systems are shown. Choice of frequency and channel capacity are discussed. S.w. s.s.b. transmission and reception apparatus for transcontinental and oversea services, and directional v.h.f. and u.h.f. multichannel equipment are described, with illustrations. Unit-type construction and miniaturization of components are discussed in considering the economics of the service.

621.396.5 **492**

A Subscriber's Battery-Operated Single-Channel V.H.F. Radio-Telephone Equipment.—J. H. H. Merriman, R. Hamer & B. R. Horsfield. (*P.O. elect. Engrs' J.*, July 1951, Vol. 44, Part 2, pp. 75–80.) "This article describes experimental equipment operating in the 70–90 Mc/s range, using phase modulation and powered by primary batteries, which, together with signalling units, provides a duplex radio channel for connection in a subscriber's line circuit. The characteristics of this equipment which make it suitable for providing telephone service to isolated communities are explained and mention is made of the performance obtained on experimental links using such equipment."

621.396.619.11 **493**

Transient Response of Asymmetrical Carrier Systems.—G. M. Anderson & E. M. Williams. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1064–1066.) A time-domain study is made of the transient response of asymmetrical a.m. carrier systems. A vector integral method is used for determining the system response to arbitrary modulation. Nonlinearity of the envelope transfer function is examined as dependent on modulation depth. The time delay associated with the response to an applied step voltage is found to be always less than the corresponding delay on cutting off the voltage. Improvement of the transient response of asymmetrical systems as a consequence of detuning appears to be accompanied by a decrease in signal noise ratio.

621.396.65 **494**

940–960-Mc/s Communications Equipment for Industrial Applications.—F. B. Gunter. (*Elect. Engng, N.Y.*, July 1951, Vol. 70, No. 7, pp. 573–578.) Factors influencing the design of a multichannel system are discussed, particularly the choice and stability of frequency, method of modulation, and aerial and receiver design. Details of one system are given.

621.396.65 **495**

Radio Relay Design Data, 60 to 600 Mc/s.—R. Guenther. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1027–1034.) I.R.E. 1950 National Convention paper. General design data for a complete radio-relay transmission system are shown graphically. From these data, signal and noise levels can be determined. The ratio of the theoretical to the practical wide-band gain is a measure of the deficiency of conversion in the demodulator. Comparison of different modulation systems must be based on measured values of wide-band gain. Numerical data for f.m. and p.t.m. systems are derived and tabulated.

621.396.65 : 621.396.41 **496**

U.H.F. Multiplex Radiotelephony Communications.—H. Chireix. (*Ricerca sci.*, July 1951, Vol. 21, No. 7, pp. 1105–1126. In French.) The subject is examined from the theoretical viewpoint and a calculation is made of the threshold value of receiver input power above which the modulation system used can provide an improvement in signal/noise ratio. The insertion loss due to radio links with or without passive relays is calculated, and hence the required transmitter power. System gain is determined for the case of f.m. on carrier and for the case of p.t.m. with f.m. or a.m. transmission. Equipment for terminal and relay stations for f.m. and p.m. systems is considered. Problems relating to systems with a large number of channels and to television relaying are discussed.

621.396.65 : 621.396.41.029.64 **497**

Multiplexing Klystrons.—W. L. Firestone. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 112–116.) Possible methods of combining several modulated r.f. signals in a single microwave transmission line involve the use of hybrid-T junctions, directional couplers, line stretchers, tunable stubs, multiple-feed aerials, band-pass or band-stop cavities and other special devices. These are discussed with particular reference to crosstalk, signal loss, bandwidth, simplicity of tuning, and interference in receiver circuits. In the final system developed, three transmitters and three receivers are connected to a common main line by means of suitably tuned cavities.

621.396.65 : 629.135.4 **498**

Siting Microwave Antennas by Helicopter.—B. I. McCaffrey. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 82–84.) Two machines, one with a transmitter and the other with a receiver, hover over proposed sites for microwave relay stations and, by means of observations at gradually decreasing heights, determine optimum heights for the aerials that are to be used.

621.396.712 **499**

German Broadcasting Transmitters.—(*Funk-Technik, Berlin*, May 1951, Vol. 6, No. 10, p. 256.) A list of existing and proposed l.w., m.w., s.w. and u.s.w. transmitters showing location, frequency and power.

621.396.7 **500**

Radio Installations. [Book Review]—W. E. Pannett. Publishers: Chapman & Hall, London, 444 pp., 45s. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 289–290.) ". . .concerned with the practical design, construction and maintenance of modern radio transmitting and receiving installations for broadcasting and commercial communications purposes."

SUBSIDIARY APPARATUS

621-526 **501**

Graphical Synthesis of Networks for A.C. Servos.—G. A. Bjornson. (*Elect. Engng, N.Y.*, July 1951, Vol. 70, No. 7, p. 571.) Summary of 1951 A.I.E.E. Winter General Meeting paper. The performance of a servo-mechanism is often improved by modifying the data signal. A method is given for determining the transfer function of a compensation network which modifies the signal in the desired manner.

621-526 : 621.3.016.35 **502**

A General Criterion for the Stability of Servo-mechanisms which are Centres of Hereditary Phenomena.—J. Loeb. (*C. R. Acad. Sci., Paris*, 30th July 1951, Vol. 233, No. 5, pp. 344–345.)

621.314.1.082.72 **503**

Electrostatic D.C. Transformer.—G.W.O.H. (*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, pp. 291–293.) Discussion of the principles, construction and possible applications of the apparatus described by Malpica (270 of January).

621.314.5 **504**

Comparative Representation of Various D.C./A.C. Converters.—H. Tigler. (*Arch. elekt. Übertragung*, July 1951, Vol. 5, No. 7, pp. 335–342.) Formulae are derived from which the performance of a converter can be assessed. Application to various types commonly used in power and communication engineering shows that the best results are obtained with arrangements with parallel oscillatory circuits, d.c. chokes and square-wave modulation.

621.314.634 + 621.314.2 **505**

The New Siemens Selenium Rectifier for Broadcasting Equipment, and the Design of Suitable Transformers and Chokes.—R. Kühn. (*Funk u. Ton*, Sept. 1951, Vol. 5, No. 9, pp. 449–465.) Details of the various types available in these sealed rectifiers are given; their power, current and voltage characteristics are tabulated and shown graphically, so that the technical data required for the design of low-power rectifier transformers and chokes can be determined. Four examples illustrate the method of calculation.

621.316.722.078.3 **506**

Characteristics of Glow Lamps in Voltage-Stabilization Circuits.—B. Schumacher. (*Frequenz*, May 1951, Vol. 5, No. 5, pp. 121–124.) Analysis of simple, compensated and bridge-type circuits and discussion of the operating characteristics of different lamps.

621.316.933 **507**

Protection of Low-Voltage Systems and Equipment.—M. Witzig. (*Brown Boveri Rev.*, April 1951, Vol. 38, No. 4, pp. 115–119.) An account of methods and equipment developed to reduce overvoltage effects on consumers' premises.

621.316.933 **508**

The Overvoltage Protection of Electrical Systems.—C. Degoumois. (*Brown Boveri Rev.*, April 1951, Vol. 38, No. 4, pp. 92–95.) The mode of operation of various types of protector is discussed, with particular reference to conversion of surge energy to heat, conversion of e.s. to e.m. energy, screening, and surge-energy storage.

621.316.933.3 **509**

The Function of Lightning Arresters in Modern Overvoltage Protection Schemes.—H. Bossi. (*Brown Boveri Rev.*, April 1951, Vol. 38, No. 4, pp. 96–105.) Discussion with particular reference to 'Resorbit' arresters, which include quenched spark gaps in series with a resistor having a nonlinear characteristic, so that a low resistance is offered to surge voltages while the resistance to normal voltages is much greater.

621-526 **510**

An Introduction to Servomechanisms. [Book Review]—A. Porter. Publishers: Methuen & Co., London, 1950, 154 pp., 7s. 6d. (*Ann. Télécommun.*, June 1951, Vol. 6, No. 6, p. A362.) One of the Methuen series of Monographs on Physical Subjects. The fundamental nature of closed control systems is discussed, and modern techniques for the analysis and synthesis of servomechanisms are presented.

621-526 **511**

Dynamik selbsttätiger Regelungen (The Dynamics of Automatic Regulation): Vol. I. [Book Review]—R. C. Oldenbourg & H. Sartorius. Publishers: R. Oldenbourg, Munich, 2nd edn 1950, 258 pp., 26 DM. (*Arch. elekt. Übertragung*, May 1951, Vol. 5, No. 5, p. 252.) The broad presentation and the mathematical introduction to regulating processes make this book particularly useful for private study.

TELEVISION AND PHOTOTELEGRAPHY

621.397.262 **512**

Multiplexed Broadcast Facsimile.—J. V. L. Hogan & J. W. Smith. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 97–99.) Facsimile newspaper pictures and print are relayed from New York to Ithaca by ultrasonic modulation superposed on normal audio modulation in f.m. transmissions. Filters ensure no interference between the facsimile reproduction and the audio programme.

621.397.5 **513**

Dot Arresting improves TV Picture Quality.—K. Schlesinger. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 96–101.) A method of receiving dot-interlace television without using a separate synchronous detector is described. The method involves a process of dot arresting, or deflection sampling. An anastigmatic deflection yoke and dot-arrestor coil assembly is described. Figures are given for the brightness and sampling merit of sine-wave arresting and gating, and practical circuits and comparison photographs are shown. Selective dot-coverage circuits remove the objectionable dot structure from large areas and leave the dots in regions of fine detail.

621.397.5 **514**

Problems in Mobile TV.—E. B. Pores. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 136..168.) Transmission of a 7-kMc/s signal to a fixed receiver from a ship moving at 18 knots was effected by manual tracking of both the transmitting and the receiving aerial. Methods of reducing to a minimum the frequency variations of the engine-driven generator are described and equipment installation problems in television broadcasts from remote locations are discussed.

621.397.5 : 535.62 **515**

Crispening Circuit for Color TV.—D.G.F. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 85–87.) Designed for improved resolution in field-sequential colour television, the 'crispening' technique sharpens the vertical edges of extended objects by combining the video waveform and its derivative, modified nonlinearly. Details are given of practical circuits for use with domestic field-sequential receivers and at the end of a coaxial cable network. Oscillograms and photographs show results obtained.

621.397.5 : 621.317](083.74) **516**

New American Standard for the Measurement of Television Transmission Quantities.—W. Reichel. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, pp. 311–313.) Summary of I.R.E. Standard 23.S1, noted in 2035 of 1950, and comment on its applicability to prospective television in Germany.

621.397·5 : 621.317.75 **517**

Television Waveform Display.—Sturley. (See 447.)

621.397.6 **518**

Television Apparatus for a 625-Line Service, shown in Belgium.—J. C. Nonnekens. (*HF, Brussels*, 1951, No. 9, pp. 237–254.) The general principles involved in the provision of a television service are stated and apparatus and techniques demonstrated at the Salon de la Radio et de la Télévision, September 1950, are described.

621.397.6 : 535.62 **519**

Picture Generator for Color Television.—R. P. Burr, W. R. Stone & R. O. Noyer. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 116–119.) A simple colour-pattern generator is obtained by modifying the circuit of an existing black-and-white-pattern source. The hue and saturation of each bar of the pattern are under the operator's control. The equipment is suitable for testing either simultaneous, dot-sequential or field-sequential systems.

621.397.611.2 **520**

[Television] **Camera Equipment.**—R. Aschen. (*Radio franç.*, March 1951, No. 3, pp. 8–15.) Technical description of apparatus ranging from Zworykin's original iconoscope to the most modern equipment.

621.397.611.2 **521**

Impedance Changes in Image Iconoscopes.—J. E. Cope & R. Theile. (*Wireless Engr*, Aug. 1951, Vol. 28, No. 335, pp. 239–247.) To obtain a good signal/noise ratio a high value of load resistance must be used, and because of the capacitance across it this gives rise to signal integration, which must be compensated by differentiation in the following amplifier. The degree of compensation must be varied according to the average brightness of the televised scene, otherwise 'streaking' effects occur due to change in impedance of the camera tube. This impedance varies with the secondary-emission current from the storage surface, which in turn depends on the changing stream of photoelectrons. Two methods of automatic compensation of this effect are discussed; a suitable feedback to the signal plate provides a practical solution.

621.397.611.2 522

Television Camera Tubes.—L. Bedford: J. D. McGee & E. L. C. White. (*Wireless Engr*, Sept. 1951, Vol. 28, No. 336, pp. 288–289.) Further discussion on 1264 and 2556 of 1951.

621.397.611.21 523

Marconi Orthicon Camera.—J. Sánchez-Cordovés. (*Rev. Telecomunicación, Madrid,* June 1951, Vol. 6, No. 24, pp. 42–49.) Detailed description of the construction and properties of this image tube, with particular reference to sensitivity, coefficient of merit, light flux required, relative output, response to colour, resolution, effect of spurious signals, and limitation due to image persistence.

621.397.62 524

Modified-Butterfly U.H.F.-TV Converter.—M. W. Slate, J. P. Van Duyne & E. F. Mannerberg. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 92–96.) Highly stable non-contacting resonators of the semi-butterfly type constructed from silvered glass-bonded mica are used for the oscillator and mixer tuned circuits. Coupling from the oscillator and aerial to the mixer circuit is by bifilar coiled transmission lines giving balanced to unbalanced transformation. A cascode i.f. amplifier feeds the v.h.f. television receiver.

621.397.62 525

Flywheel Synchronizing Circuit for Television-Receiver Time-Bases.—A. B. Starks-Field. (*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, pp. 293–297.) In the normal triggering method of synchronizing television line time-bases, random variations in the firing time may be produced by the effect of noise on the synchronizing signal, giving rise to irregularities in any vertical edges in the picture. A 'flywheel' synchronizing circuit is described in which the timebase phase is compared with that of the incoming synchronizing signal, the difference being used to generate a control voltage which governs the time-base recurrence frequency.

621.397.62 : 621.385.2 526

A New Damper Diode.—Bareiss. (See 551.)

621.397.621 527

Efficiency Line-Scan Circuits.—W. T. Cocking. (*Wireless World*, Aug.–Oct. 1951, Vol. 57, Nos. 8–10, pp. 302–305, 347–350 & 425–430.) To design 'efficiency' line-scan circuits successfully, proper allowance must be made for departures from ideal characteristics of the transformer. The directly fed deflector-coil circuit is examined first, as being not only simpler to analyse but also a practical alternative to the transformer-coupled circuit. Control of the 'efficiency' diode by means of a damped tuned circuit with a phase-reversing winding is explained, and the relative merits of operation with and without such control are discussed. The reduction of efficiency inherent in the use of a transformer is outweighed when h.v. boost is used by recovering energy from the diode autobias circuit. Differences in U.S. and British practice are illustrated and explained.

621.397.621 528

Theory of Deflection Systems for Television Tubes.—F. Kirschstein. (*Funk u. Ton*, April 1951, Vol. 5, No. 4, pp. 177–189.) Description, with circuit diagrams and performance curves, of the theory and operation of magnetic deflection circuits used currently in receivers.

621.397.621.2 529

General Characteristics of Cathode Ray Tubes [for television].—(*Radio franç.*, March 1951, No. 3, p. 24.) Characteristics of six tubes are tabulated, one using electrostatic deflection, with maximum anode voltage 6 kV, and the others having electromagnetic deflection, with maximum anode voltages ranging from 12 to 19 kV.

621.397.645 530

R.F. Amplifier for U.H.F. Television Tuners.—Tyson & Weissman. (See 370.)

621.397.645.37 531

Television Picture Amplifier with Negative Feedback for the Reduction of Microphony and Input-Circuit Noise.—W. Dillenburger. (*Funk u. Ton*, March 1951, Vol. 5, No. 3, pp. 113–123.) The use of a high-resistance load for the camera tube and subsequent selective feedback in the amplifier circuit is discussed. The design of the circuit is considered taking account of interelectrode and stray capacitances. The circuit and typical response curves of a single-stage and three-stage amplifier are shown. While the design of the former is simple to calculate, the latter is best designed empirically; component values are critical and the circuit tends to self-oscillation; an advantage is that nonselective feedback can be applied to the individual stages, hence valve replacement is simplified.

621.397.8 532

Long-Distance Reception of Television.—(*Radio-Revue TV*, June 1951, No. 6, p. 231.) Dates and times in May 1951 are quoted when good reception of television broadcasts was achieved in Belgium over ranges as great as 2 200 km, mainly from Leningrad. Such reception is attributed to propagation by reflection from sporadic-E clouds.

621.397.5(43) 533

Einführung in die neue deutsche Fernsehtechnik (Introduction to Modern German Television Technique). [Book Review]—W. Dillenburger. Publishers: Schiele & Schön, Berlin, 1950, 210 pp., 12.50 DM. (*Z. angew. Phys.*, May 1951, Vol. 3, No. 5, p. 200.) Written for technicians and engineers, but excluding difficult mathematical treatment.

TRANSMISSION

621.396.61 : 621.396.619.13 534

High-Power F.M. Broadcasting Transmitters.—L. Rohde, H. Nitsche & A. Pfefferl. (*Frequenz*, Feb. 1951, Vol. 5, No. 2, p. 52.) Corrections to 764 of 1951.

621.396.619.13 535

Characteristics of Frequency Modulation with Reactance Valves.—G. Bronzi. (*Alta Frequenza*, Aug. 1950, Vol. 19, No. 4, pp. 175–184.) A theoretical-experimental method of investigation is proposed based on measurement of two frequencies, viz., the centre (carrier) frequency and that obtained with the modulator valve off. The applicability of the method is confirmed by experimental results.

621.396.619.13 : 621.396.61 536

Arrangement for Frequency Modulation of Transmitters by means of Premagnetized H.F. Iron-Cored Coils.—H. Boucke. (*Fernmeldetech. Z.*, May 1951, Vol. 4, No. 5, pp. 201–206.) A method was investigated in which the inductance of a frequency-determining h.f. iron-cored coil was varied in accord with the modulation frequency. The centre frequency used was 80 Mc/s with a frequency swing of about 1 Mc/s and the highest modulation frequency was 160 kc/s. Details are given of the magnet winding and core and of the oscillator coil, and the various possible sources of nonlinear distortion are investigated.

621.396.619.2 **537**

'F M Q'.—W. S. Mortley. (*Wireless World*, Oct. 1951, Vol. 57, No. 10, pp. 399–403.) The Marconi 'frequency-modulated quartz' system was devised to give a reliable circuit with adequate centre-frequency stability and low distortion. The modulator is of balanced susceptance type and a crystal with a quarter-wave network attached is used in the oscillator. A circuit is described which gives the required frequency modulation with audio input.

621.396.67 **538**

V.H.F. Common-Aerial Working.—E. G. Hamer. (*Electronic Engng*, July 1951, Vol. 23, No. 281, pp. 244–249.) The problems involved are reviewed briefly and the design of suitable filters using lengths of cable or lumped elements is discussed. The special difficulties encountered in common-aerial working using transmitters whose frequency difference is between 5% and 0·6%, or $<0·6\%$, are also dealt with.

621.396.97 : 621.396.61 **539**

Improving Program-Limiter Performance.—D. W. Howe, Jr. (*Electronics*, Sept. 1951, Vol. 24, No. 9, pp. 108–111.) Circuit development, adjustment, operation and performance data are given for a delay limiter which prevents sideband spread on a.m. broadcasting channels. Fringe-area reception is improved, overmodulation peaks are reduced in number and greater total modulation energy is obtained.

VALVES AND THERMIONICS

621.383.42 **540**

New Photocells.—J. Stuke. (*Elektron Wiss. Tech.*, April 1951, Vol. 5, No. 4, pp. 112–114.) Discussion of improvements obtained in sensitivity and durability, particularly under tropical conditions, in the new design of Se photocell in which a semiconducting layer with enhanced conductivity replaces the thin semitransparent metallic coating usually applied over the Se layer.

621.385 : 621.317.723 **541**

Receiving Valves suitable for Electrometer Use.—G. A. Hay. (*Electronic Engng*, July 1951, Vol. 23, No. 281, pp. 258–261.) Ordinary receiving valves of certain types can be used as electrometer valves after selection and special treatment. The characteristics of these are compared with those of special valves.

621.385.017.72 **542**

Cooling of Electronic Valves by Vaporization of Water.—C. Beurtheret. (*Onde élect.*, June 1951, Vol. 31, No. 291, pp. 271–281.) Anode cooling of high-power valves is greatly improved by the use of thick anodes, with correctly shaped fins, enclosed in a water jacket. The water is vaporized, thus abstracting heat from the anode, and the steam is condensed and the water re-used. With a smooth anode, small bubbles of vapour adhere to the surface and form a nonconducting layer, leading to the destruction of the valve; by using the finned surface the power dissipated by the anode may be increased by a factor of three without risk. Valves are illustrated having anode dissipations of 40–50 kW.

621.385.029.63/.64 **543**

Amplification and Generation of Microwave Oscillations with Travelling-Wave Valves.—H. Schnitger. (*Funk u. Ton*, March 1951, Vol. 5, No. 3, pp. 143–145.) Data on the noise factor of the helix-type valve and the double-beam valve, concluding the paper noted in 3004 of 1950.

621.385.029.63/.64 **544**

The Calculation of Travelling-Wave-Tube Gain.—C. C. Cutler. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 914–917.) A summary of available relevant information presented in the form of graphs and nomograms.

621.385.029.63/.64 **545**

Production of Direction-Dependent Electronic Attenuation in Travelling-Wave Valves.—H. Schnitger. (*Fernmeldetech. Z.*, July 1951, Vol. 4, No. 7, pp. 301–305.) Investigation of the attenuation of the normal reflected wave effected by interaction with a second wave injected into the valve and travelling in the opposite direction. This is achieved by suitably decreasing the beam-current intensity and making the beam voltage 10–20% lower than for optimum amplification. Quantitative measurements show general agreement with calculated values of attenuation. For a helix 13 cm long and mean wavelength 30·5 cm the attenuation was >2 neper over 9% of the waveband.

621.385.029.63/.64 : 621.396.822 **546**

Noise in Traveling-Wave Tubes.—F. N. H. Robinson & R. Kompfner. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 918–926.) Gain, noise factor and attenuation in travelling-wave valves are discussed; the design and performance of a valve with a dispersive helix are described. Much of the noise observed may be due to beam partition effects; these are analysed, and approximate expressions are given for their magnitude. Noise due to electron-wave-type amplification and thermal noise originating in the resistive part of the helix are also discussed, and experiments are described.

621.385.029.64/.65 **547**

A Spatial-Harmonic Traveling-Wave Amplifier for Six-Millimeters Wavelength.—S. Millman. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1035–1043.) An amplifier is described in which the electron beam interacts with a spatial harmonic of an e.m. wave propagated along an array of resonator slots. This construction makes possible a considerable reduction of beam voltage for a given physical separation of the circuit elements. With a valve of this type operating with a beam voltage of 1 200 V, net power gains of about 18 db have been obtained at wavelengths of about 6 mm. A magnetic field of 1 600 gauss is sufficient for proper beam focusing. Apart from small variations of gain with frequency, caused by internal reflections, the bandwidth is of the order of 3%.

621.385.032.21 **548**

New Cathode Design improves Tube Reliability.—D. R. Hill. (*Electronics*, Aug. 1951, Vol. 24, No. 8, pp. 104–106.) See also 773 of 1951.

621.385.032.216 **549**

Thermionic Emission from (Ba-Sr)O Cathodes Illuminated by the Incandescent Lamp.—T. Hibi & K. Ishikawa. (*Phys. Rev.*, 1st Aug. 1951, Vol. 83, No. 3, pp. 659–660.) The emission current was examined as a function of intensity and duration of illumination. From the experimental results it is concluded that the trapping energy of an electron in an isolated impurity centre is greater than 1·32 eV.

621.385.032.216 **550**

Mechanism of Thermionic Emission from an Oxide Cathode in Pulsed Operation.—Ya. E. Pokrovski. (*Zh. eksp. teor. Fiz.*, March 1951, Vol. 21, No. 3, pp. 423–428.) The main phenomena of pulsed electron emission from an oxide cathode are discussed on the basis of the energy-level theory of semiconductors. Reasons for inaccuracies in earlier theories of electron emission from semiconductors are pointed out, and a formula is derived representing the process of fatigue of the oxide cathode. The physical meaning of the constant characterizing the speed of the fatigue process is established, and the effect of a barrier layer on the emission properties of the oxide cathode is taken into account. The theoretical results obtained are in agreement with experimental data.

621.385.2 : 621.397.62 **551**

A New Damper Diode.—M. Bareiss. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 262..270.) Description of the design of an experimental valve similar to Type 6W4GT, but incorporating a new heater assembly and with an increased anode-cathode spacing. Heating time is 16 sec. Sample valves withstood a short inverse peak-voltage test of 6 000 V.

621.385.3 **552**

The Amplification Factor of a Triode: Part 1—A Parallel Plane Triode with Arbitrary Electrode Dimensions.—M. Wada. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. B*, March 1951, Vol. 1/2, No. 3, pp. 399–420.) A general theory is developed for determining the characteristics of planar triodes. This is used to determine the amplification factor of u.h.f. triodes with small grid-cathode spacing.

621.385.3 : 546.289 **553**

The Germanium Triode.—H. Salow. (*Z. angew. Phys.*, June 1951, Vol. 3, No. 6, pp. 231–239.) Discussion of the construction, characteristics, operation and applications of the transistor and related devices.

621.385.832 **554**

Charge Storage in Cathode-Ray Tubes.—C. V. Parker. (*Proc. Inst. Radio Engrs*, Aug. 1951, Vol. 39, No. 8, pp. 900–907.) "The charging process in cathode-ray tubes used for static storage of information is analyzed for both a stationary spot and a linear scan, with and without redistribution of secondary electrons. Approximate equations are derived for the surface potentials and charging current as functions of time and other parameters, such as primary beam current, writing speed, and initial potentials. The results are presented graphically in special cases for comparison with photographs of experimental wave forms."

621.396.615.14 **555**

New U.H.F. Resnatron Designs and Applications.—D. B. Harris. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 86–89.) Resnatrons have been developed for u.h.f. amplifiers with high power output. Gain, bandwidth, efficiency, frequency stability and noise level are satisfactory for television transmission.

621.396.615.141.2 : 621.365.55† **556**

Magnetrons for Dielectric Heating.—R. B. Nelson. (*Elect. Engng, N.Y.*, July 1951, Vol. 70, No. 7, pp. 627–633.) The use of centimetre wavelengths for dielectric heating leads to a high rate of heating, but entails disadvantages in respect of the difficulty of screening the equipment, nonuniform heating due to the production of interference patterns, and attenuation of waves as they penetrate the dielectric. A suitable compromise is achieved at the commonly used frequencies of 915 and 2 450 Mc/s. A 5-kW magnetron and associated circuits for use at 915 Mc/s are described.

621.396.615.142.2 **557**

Wide-Band 5-kW Klystron Amplifier.—P. Guénard, B. Epsztein & P. Cahour. (*Ann. Radioélect.*, April 1951, Vol. 6, No. 24, pp. 109–113.) The inherently narrow passband of the klystron amplifier is expanded, at the expense of overall gain, by using a three-cavity klystron in which the resonance frequencies of the first two cavities are appropriately spaced. A c.w. power of 5 kW at 900 Mc/s is obtained, with a gain of 20 db and a variation of less than 1 db in a bandwidth of 6 Mc/s.

621.396.615.142.2 **558**

High-Power U.H.F.-TV Klystron.—Engineering Staff of Varian Associates. (*Electronics*, Oct. 1951, Vol. 24, No. 10, pp. 117–119.) A two-stage triple-cavity klystron amplifier for use in the band 470–890 Mc/s is described. The cathode, of tantalum, is heated by electron bombardment and can be replaced to extend the life of the valve. The cavities are stagger tuned to provide a bandwidth of 5·6 Mc/s between 1-db points. By using a water-cooled collector, c.w. output powers of 5 kW are obtained, the gain being 24 db. See also (A) 2304 of 1951, No. 46 (Abraham et al.).

621.396.822 **559**

Induced Grid Noise and Noise Factor.—R. L. Bell. (*Proc. Inst. Radio Engrs*, Sept. 1951, Vol. 39, No. 9, pp. 1059–1063.) An approximate treatment of narrow-band triode noise factor is developed, taking account of induced grid noise by a new method. Minimal values of noise factor predicted theoretically are compared with experimental values; from a knowledge of circuit losses, shot noise, mutual conductance, space-charge input capacitance, and d.c. grid current on the valve, noise factor and optimum source conductance can be predicted with accuracy over a wide range of operating conditions. Lead inductance effects and normal induced grid effects (grid noise, space-charge capacitance, transit-time damping) have no influence on noise performance, but merely affect tube admittances. Direct compensation in the grid circuit of components of input admittance to which these effects give rise, leads to a deterioration of noise factor usually attributed to grid noise. The optimum grid-circuit detuning may be provided automatically by the input space-charge capacitance.

MISCELLANEOUS

001.891 **560**

Scientific Research of Philips' Industries from 1891 to 1951.—W. de Groot. (*Philips tech. Rev.*, July/Aug. 1951, Vol. 13, Nos. 1/2, pp. 3–48.) An account written to commemorate the jubilee of the Philips organization at Eindhoven. It was not possible to celebrate this event suitably at the due time, in 1941. Research is organized under five broad headings: (*a*) light and its production, including gas discharges; (*b*) electrotechnics, acoustics and radio; (*c*) chemistry, including metallurgy; (*d*) X-ray investigations; (*e*) mathematics and mathematical physics. The historical development in all these fields is outlined.

621.396 : 061.4 **561**

18th National Radio Exhibition.—(*Wireless Engr*, Oct. 1951, Vol. 28, No. 337, pp. 316–320.) A number of developments in evidence at the exhibition are discussed, chief attention being paid to television receivers. Methods of station selection and various improved scanning circuits are described. Other items discussed include new models of test equipment, and trends in sound-broadcasting receiver design. Component developments are briefly reviewed.

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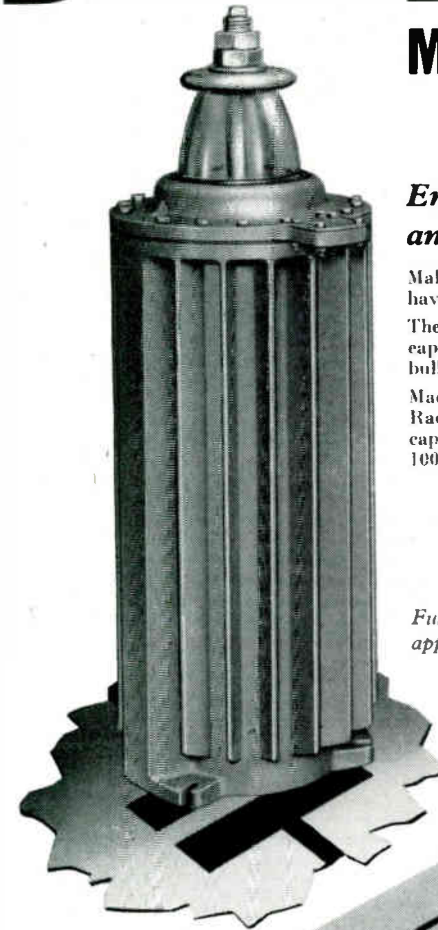

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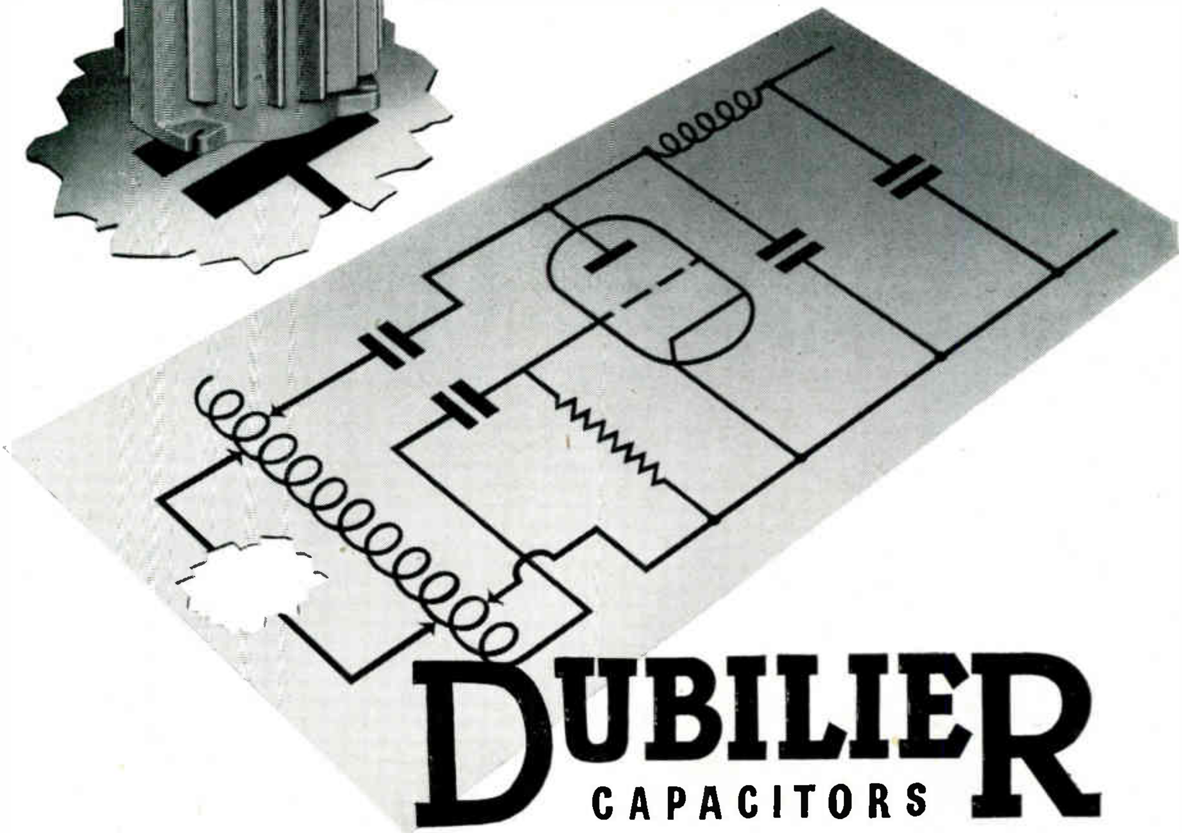

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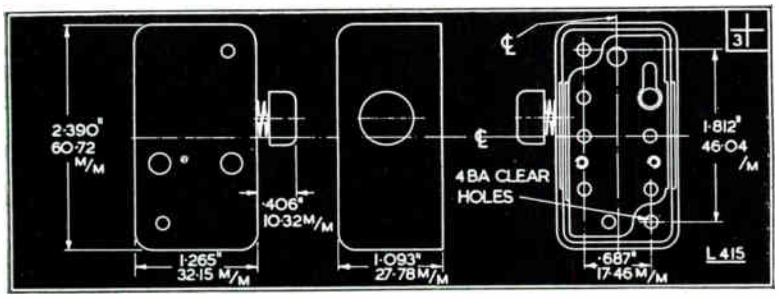

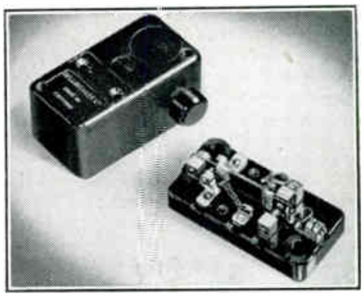

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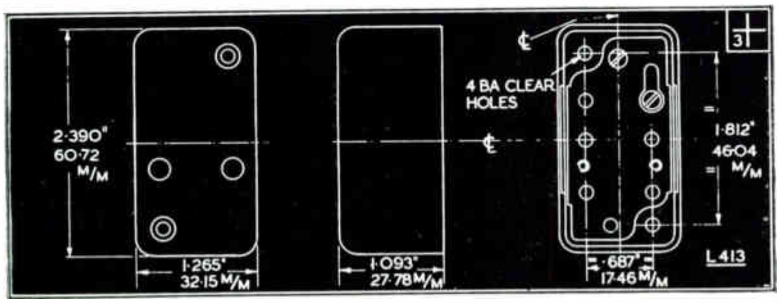

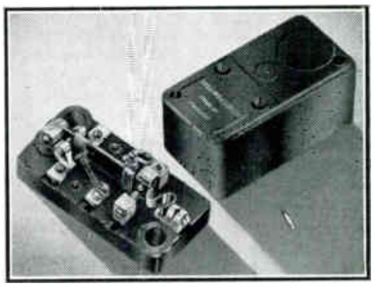

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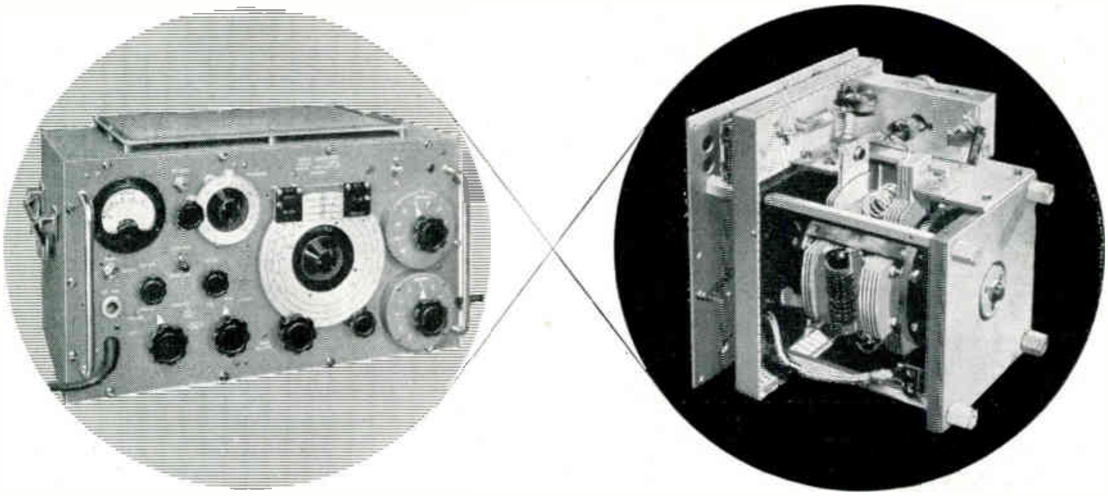

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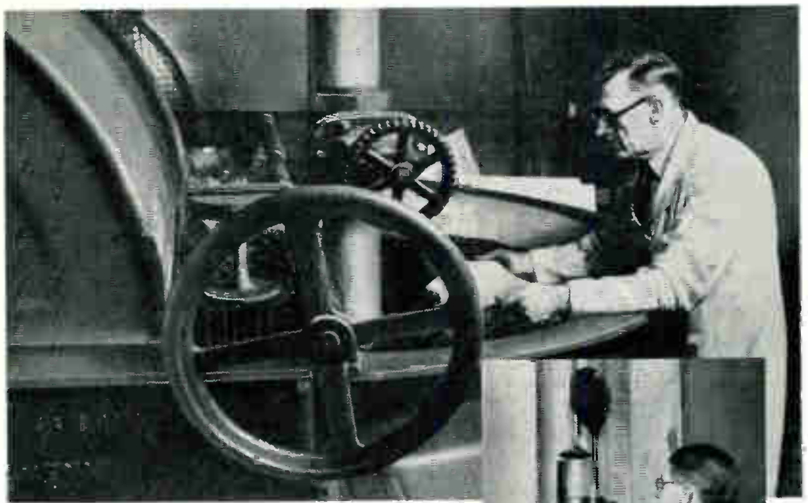
Pulp is very important. It is a key stage in the production of a component which forms a vital part of a loudspeaker—the cone. Cone, and hence speaker quality, are dependent on the pulp, and its manufacture is a rigidly controlled process.

The "beaterman" in our illustration is feeding raw pulp board into the "beater", where the cellulose fibres are first "brushed out"—stroked and straightened under a precision roller—then "beaten"—cut to an accurately controlled length.

At this stage the pulp is impregnated with the chemicals which render the cones proof against humid conditions and the attacks of termites and bacteria. The pulp is then ready to be formed into "raw" cones or "felts", the second stage of its progress from crude paper mill board to sensitive speaker diaphragm: the subject of the next announcement in this series.



Testing at every stage is the rule at Goodmans. With the Schopper-Reigler apparatus (right) our chemist makes frequent tests for pulp consistency.

AUDIOM 60 12″ 15w P.M. Loudspeaker

Limited supplies of this speaker are now becoming available again on the home market. If you are unable to obtain supplies, please let us know your dealer's name.

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- ★ **No tapped matching transformer employed.**
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- ★ **Seventeen input impedances from 2 to 5,000 ohms available by switching.**
- ★ **Cannot contribute to the distortion when testing high-power amplifiers.**

Illustrated descriptive leaflets of any or all Airmec instruments will be forwarded upon request

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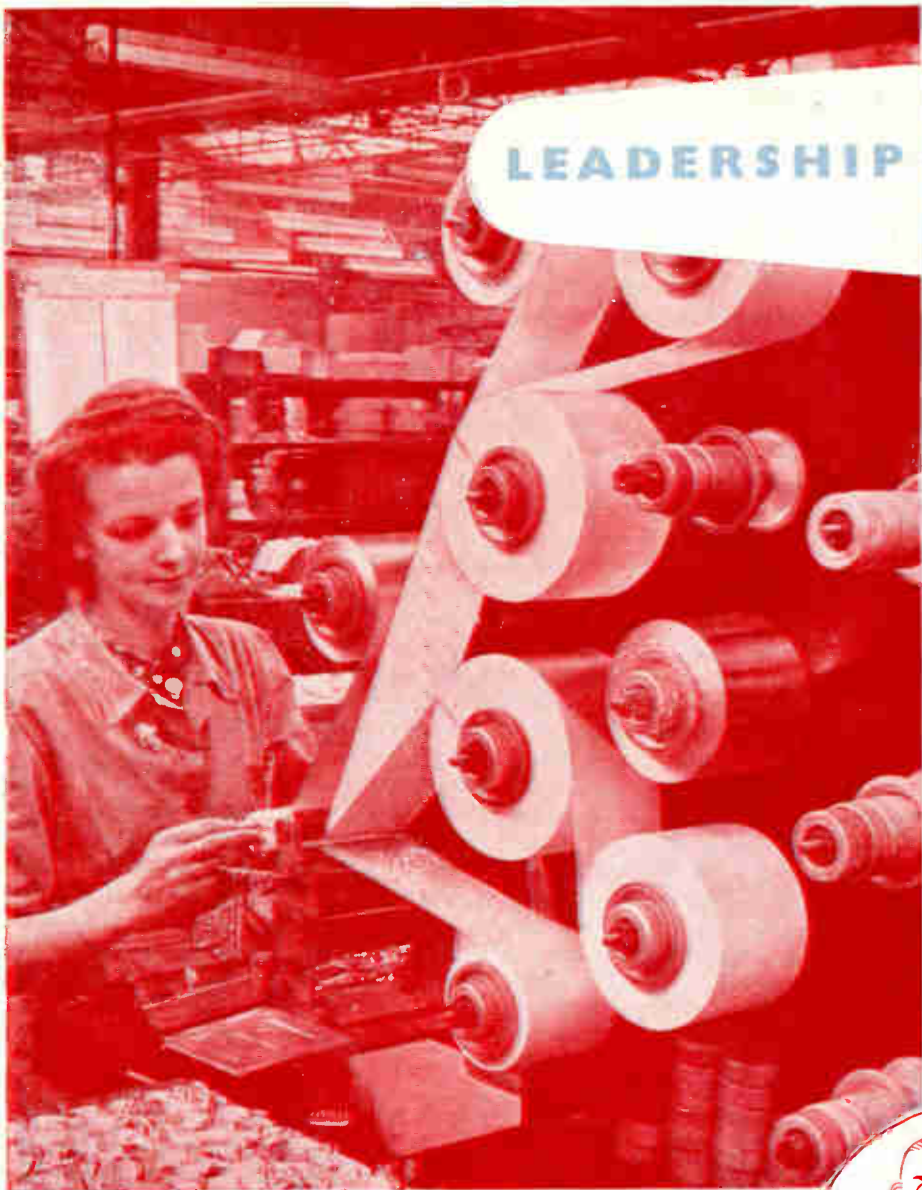
Printed in Great Britain for the Publishers, Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1.
at The Baynard Press by Sanders Phillips & Co., Ltd., Chryssell Road, London, S.W.9.



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The illustration shows one of the multi-reel paper dielectric winding machines designed to simplify the winding of many ranges of condensers.

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RADIO. An operative using Ersin Multicore Solder in the factory of Bang & Olufsen, Struer—one of Denmark's largest radio manufacturers.

TELEVISION. This single unit of the famous Pye Portable Television Transmitting Equipment has more than 3,000 Ersin Multicore soldered joints.

ELECTRONIC COMPONENTS. Ersin Multicore is used in large quantities by H. Karayannis & Co., Athens, the leading electronic distributors in Greece.

ELECTRIC MOTORS. At the Normand Electrical factory, London, an operative using Ersin Multicore to solder connections to commutator bars on lap-wound armatures for Neco fractional horse-power motors.

TELEPHONE EQUIPMENT. Ersin Multicore being used to solder switchboard apparatus at Autelca Mediterranea, S.A.T.A.P.—one of Italy's largest manufacturers of telephone equipment.

ELECTRIC LAMPS. The Johannesburg factory of African Lamps Ltd., where Ersin Multicore is used for soldering the cap connections by hand.

DRY BATTERIES. A section of an H.T. battery showing Ersin Multicore soldered joints on the top of the cells. In addition, each wire is soldered to the side of each battery can.

TECHNICAL INFORMATION

Laboratory Engineers and Technicians are invited to write for samples of alloys and a copy of "Modern Solders" containing useful tables of melting points, fully illustrated chapters on the Constitution of Alloys, Use of Solder, Standard Gauges, Flux Facts, etc.

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