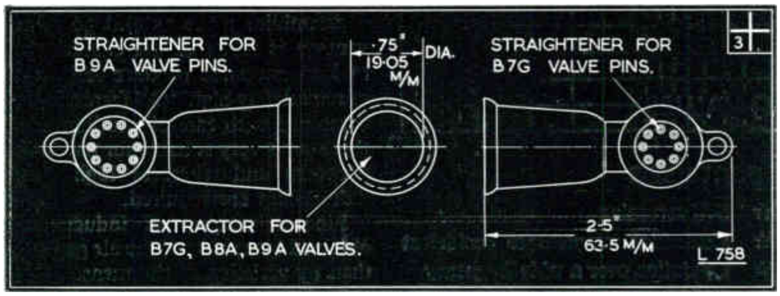
WIRELESS ENGINEER

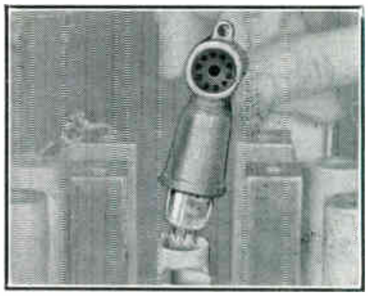
THE JOURNAL OF RADIO RESEARCH & PROGRESS

JUNE 1952

VOL. 29 No. 345 THREE SHILLINGS AND SIXPENCE

The "Belling-Lee" page for Engineers

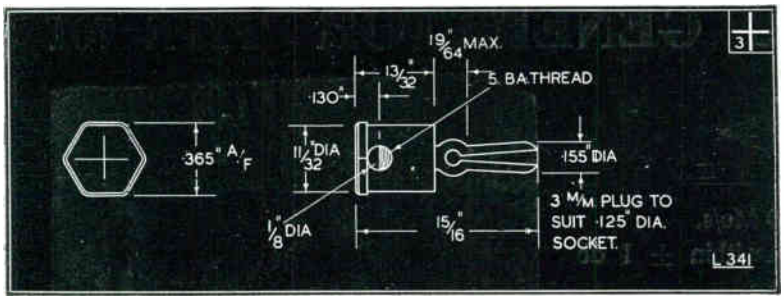


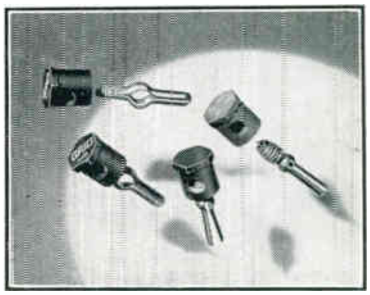


VALVE EXTRACTOR

LIST NUMBER L 758

Suitable for B7G, B8A, and B9A valves, this novel valve extractor will be found extremely useful for removing valves from equipment made up of closely packed components. The extractor is moulded in rubber, and a pin straightener for B7G and B9A valves is incorporated in the handle. This is accurately moulded in a hard phenolic material, and will obviate the damage to valves and/or holders caused through trying to force insertion with bent pins.





"BOWSPRING" WANDER PLUG

LIST NUMBER L 341

The "bowspring" design of the turned brass pins gives a remarkably strong contact. Moulded in black or red phenolic material, and fitted with hexagonal cap; side loading for flexible leads.
Can be supplied with standard letterings, subject to special quotations dependent on quantities.

The above items and many more are fully described in our 1952 General Catalogue. Please write quoting P365/WE.

BELLING & LEE LTD
CAMBRIDGE ARTERIAL RD., ENFIELD, MIDDX., ENGLAND

ERIE

Reg. Trade Mark

Super-k-miniature

Insulated

Tubular Ceramicons

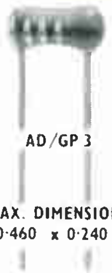
TYPE GP 3

Higher capacitances, in standard sizes, than ever before.

GP 3 Ceramicons go a stage further than GP 2 by offering the designer an even higher capacitance (tube-size for tube-size) than ever before. They are manufactured from a new ceramic of high dielectric constant, designated Hi-K 35, which, after patient research has been developed from the well known Erie Hi-K 15 material, which has enjoyed Government type approval ever since its introduction in 1943, and is still being used with satisfaction, in tens of millions of Ceramicons all over the world.

Having been proved by normal and accelerated life tests since 1949, and in overseas markets since 1950, this new material is now being made available for home applications.

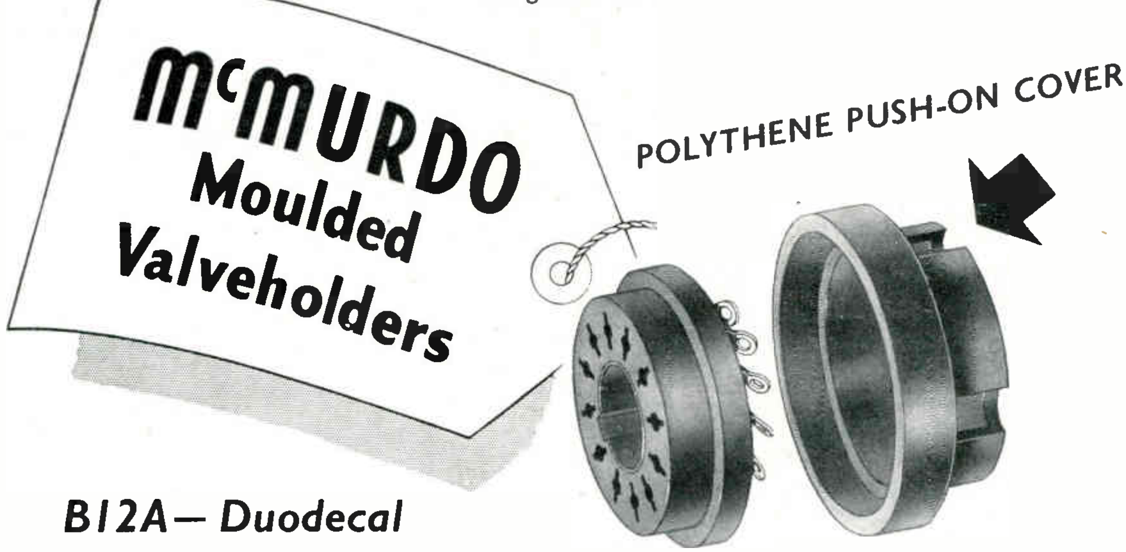
FLASH TEST: 1150 volts D.C.
INSULATION RESISTANCE: Greater than 7500 megohms.
LIFE TEST: 700 volts D.C. at 71°C 1000 hours.
WORKING VOLTAGE: 350 D.C. at 71°C.
POWER FACTOR: Not greater than 2·5%, when measured at room temperature, and at 300 Kc/s, with an applied potential not greater than 5 volts R.M.S.
TOLERANCE ON CAPACITANCE: +80%–20%.
TEMPERATURE CHARACTERISTICS: The capacitance of GP 3 Ceramicons shall not decrease more than 25% from their value at room temperature, as temperature is varied from +10°C to 75°C.

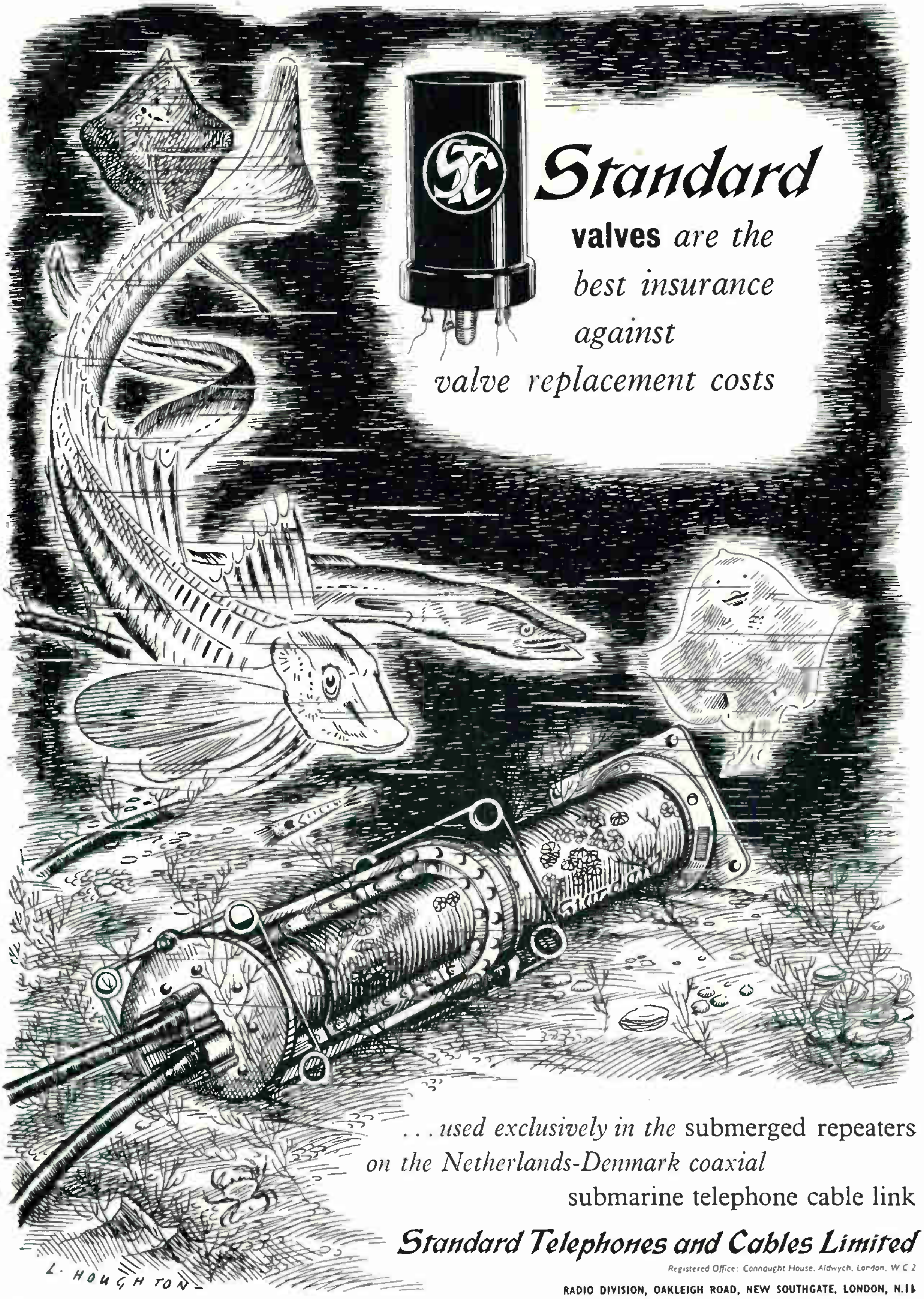
STYLE	CAPACITANCE
 AD/GP 3 MAX. DIMENSIONS 0·460 x 0·240	1000 to 2000 mmfd (0·002 MFD)
 BD/GP 3 MAX. DIMENSIONS 0·710 x 0·240	2001 to 5000 mmfd (0·005 MFD)

ERIE *Resistor Limited*

CARLISLE ROAD, THE HYDE, LONDON, N.W.9., ENGLAND
Telephone: COLindale 8011. Cables: Resistor London

Factories: London and Gt. Yarmouth; Toronto, Canada; Erie, Pa., U.S.A.


McMURDO
Moulded
Valveholders
POLYTHENE PUSH-ON COVER


STC
Standard
valves are the best insurance against valve replacement costs
. . . used exclusively in the submerged repeaters on the Netherlands-Denmark coaxial submarine telephone cable link
Standard Telephones and Cables Limited
Registered Office: Connaught House, Aldwych, London, W.C.2
RADIO DIVISION, OAKLEIGH ROAD, NEW SOUTHGATE, LONDON, N.11
L. HOUGHTON


OFF
ON
SPARK GAP
TESVAC
FERRANTI
FERRANTI
Tesvac
HIGH FREQUENCY VACUUM TESTER
HIGH FREQUENCY Approximately 4 megacycles per second.
HIGH VOLTAGE Approximately 25 kV. maximum.
Simple and safe to use

WIRELESS ENGINEER

The Journal of Radio Research and Progress

Managing Editor: HUGH S. POCOCK, M.I.E.E.
Editor: W. T. COCKING, M.I.E.E. Technical Editor: Professor G. W. O. HOWE, D.Sc., LL.D., M.I.E.E.

Volume 29 · Number 345

CONTENTS

JUNE 1952

Published on the sixth of each month

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

Editorial, Advertising and Publishing Offices: Dorset House, Stamford Street, London, S.E.1
Telephone: Waterloo 3333 (60 *lines*) · *Telegrams: Wirenger, Sedist, London*

BRANCH OFFICES AT BIRMINGHAM · MANCHESTER AND GLASGOW

Radar Display Tubes with High Resolution

The most important requirements of a radar display tube are that it shall be capable of a resolution at least equal to that of the radar system employed; that it shall present a picture of high brilliance and contrast; and that it shall operate reliably under the continuous and arduous conditions of service encountered in navigational radar applications.

The Mullard range of radar display tubes, developed with the advantage of the vast experience gained in the production of television picture tubes, has been specially designed to meet these requirements.

Characterised by very high definition, high brilliance and contrast, low deflection defocusing, low astigmatism, and long after-glow, these tubes are now being extensively used in a wide variety of navigational radar equipments. Brief descriptive details of these tubes are given below; more comprehensive information will gladly be supplied on request.

	MF41-15	MF31-55	MF13-1	
DESCRIPTION	A 16" flat-faced, wide angle, radar display tube with a metallized fluoride screen. Ideal for use in harbour radar systems.	A 12" flat-faced radar display tube with a metallized fluoride screen. This tube is on the Government list of Preferred Types (CV429).	A 5" compact, flat-faced, radar display tube with a metallized fluoride screen. Designed for small marine and airborne radar displays where high performance coupled with saving in space is required. This tube is the equivalent of the 5FP7A.	
HEATER				
V_h	6.3	6.3	6.3	V
I_h	0.3	0.3	0.3	A
LIMITING VALUES (absolute ratings)				
V_{a2} max.	16	15	11	KV
V_{a2} min.	6	7	5.5	KV
V_{a1} max.	450	†600	450	V
V_{a1} min.	200	250	200	V
$-V_g$ max.	200	200	200	V
DIMENSIONS				
Max. bulb diameter	406	307	127.5	mm
Max. overall length	515	520	289	mm
Useful screen diameter	360	260	108	mm
BASE	B12A	B12A	Octal.	

† Beam current not to exceed 50 μA

Mullard

MULLARD LTD., COMMUNICATIONS AND INDUSTRIAL VALVE DEPT., CENTURY HOUSE, SHAFTESBURY AVE., LONDON, W.C.2

MVT 114

WIRELESS ENGINEER

Vol. 29 JUNE 1952 No. 345

The Gyrator

IN the *Philips Research Reports* for April 1948, B. D. H. Tellegen suggested that network synthesis would be considerably modified, and even simplified, if, in addition to the four recognized circuit elements, viz., resistors, inductors, capacitors, and transformers, we had a fifth one, viz., a four-pole which violated the reciprocity relation. In the ordinary four-pole shown in Fig. 1 the relations between currents and voltages are given by the equations

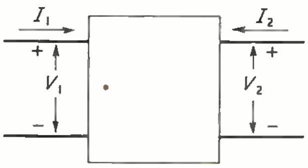


Fig. 1.

$$\left.\begin{aligned} V_1 &= Z_{11}I_1 + Z_{12}I_2 \\ V_2 &= Z_{21}I_1 + Z_{22}I_2 \end{aligned}\right\} \quad .. \quad .. \quad (1)$$

From these we have

$$\left.\begin{aligned} I_1 &= (1/Z_{11})V_1 - (Z_{12}/Z_{11})I_2 \\ V_2 &= (Z_{21}/Z_{11})V_1 + (Z_{22} - Z_{21}Z_{12}/Z_{11})I_2 \end{aligned}\right\}(2)$$

In the ideal transformer $Z_{11} = \omega L_1$, $Z_{22} = \omega L_2$ $Z_{12} = Z_{21} = \omega M$ and $L_1 L_2 = M^2$. Making these substitutions in (2) we have

$$\begin{aligned} I_1 &= V_1/\omega L_1 - uI_2 \\ V_2 &= uV_1 \end{aligned}$$

where $u = M/L_1 = L_2/M$ is the transformation ratio. The term $V_1/\omega L_1$ is the open-circuit magnetizing current and can be neglected, giving the equations

$$\left.\begin{aligned} I_1 &= -uI_2 \\ V_2 &= \quad uV_1 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (3)$$

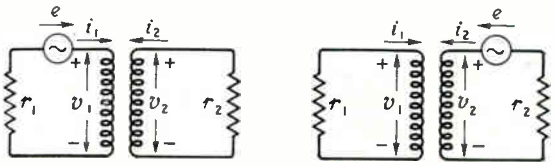
The minus sign is due to the symmetrical assumption of positive current and voltage directions.

It is perhaps not generally realized that the reciprocity relation holds in such a case; that is, that an e.m.f. inserted in the primary circuit produces the same current in the secondary circuit as would be produced in the primary circuit if the same e.m.f. were inserted in the secondary circuit. In Fig. 2(a) the e.m.f. e is inserted in the primary circuit and we have

$$v_1 = e - i_1 r_1,\ v_2 = uv_1 = ue - ui_1 r_1$$

$$-i_2 = \frac{v_2}{r_2} = \frac{ue}{r_2} - \frac{ui_1 r_1}{r_2} = \frac{ue}{r_2} + \frac{u^2 r_1}{r_2} i_2$$

$$\therefore -i_2 = \frac{ue}{u^2 r_1 + r_2}$$



(a) (b)

Fig. 2.

In Fig. 2(b) the e.m.f. e is inserted in the secondary circuit and we have

$$v_2 - e = -i_2 r_2,\ v_1 = \frac{v_2}{u} = \frac{e}{u} + \frac{r_2}{u^2} i_1 = -i_1 r_1$$

$$-i_1 = \frac{e}{ur_1} + \frac{r_2}{u^2 r_1} i_1,\ i_1\left(1 + \frac{r_2}{u^2 r_1}\right) = -\frac{e}{ur_1}$$

$$\therefore -i_1 = \frac{ue}{u^2 r_1 + r_2}$$

Hence, a positively directed e.m.f. in either case produces a negative current, or expressed in another way, a right-handed e.m.f. in either circuit produces a right-handed current of the same magnitude in the other circuit.

Up to this point we have confined our attention to the ordinary four-pole. The special four-pole which Tellegen conceived and to which he gave the name "gyrator" has properties defined by the equations

$$\left.\begin{aligned} v_1 &= -si_2 \\ v_2 &= si_1 \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

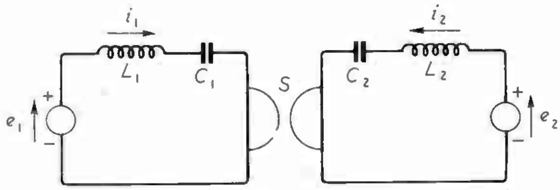


Fig. 3.

Whereas the transformation ratio n is a mere number, s has the dimensions of a resistance, and is called the *gyration resistance*; as Tellegen says, it 'gyrates' a current into a voltage and vice versa. From (4) one obtains the following strange results. If the secondary circuit is open $i_2 = 0$ and therefore $v_1 = 0$, hence the primary is effectively short-circuited by opening the secondary. Connecting an inductance L between the secondary terminals is equivalent to connecting a capacitance $C = L/s^2$ between the primary terminals, and conversely, connecting a capacitance C between the secondary terminals is equivalent to connecting an inductance $L = s^2C$ between the primary terminals. More generally, an impedance Z connected in series or in parallel with the secondary load is equivalent to an impedance s^2/Z respectively in parallel or in series with the primary terminals. As an example, in the case of an inductance L across the secondary terminals, we have from (4)

$$i_2 = -\frac{v_2}{j\omega L} = -\frac{si_1}{j\omega L} \text{ and } v_1 = -si_2 = \frac{s^2 i_1}{j\omega L}$$

Hence $i_1 = \dfrac{j\omega L}{s^2} v_1$ equivalent to $j\omega C v_1$ if $C = L/s^2$.

In the case of a resistance R and an inductance L in series between the secondary terminals,

$$v_2 = si_1 = -i_2(R + j\omega L) = \frac{v_1}{s}(R + j\omega L)$$

$$\therefore i_1 = \frac{v_1}{s^2}(R + j\omega L) = \frac{v_1}{s^2/R} + jv_1\left(\frac{\omega L}{s^2}\right)$$

which is equivalent to a resistance s^2/R in parallel with a capacitance $C = \omega L/s^2$ between the primary terminals. Fig. 3 shows two tuned circuits coupled by an ideal gyrator; e_1 and e_2 are the e.m.fs acting in the two circuits. The symbol employed in Fig. 3 to represent the gyrator is that suggested and used by Tellegen. From equations (4) we obtain by putting $i = dQ/dt$

$$\left.\begin{aligned} L_1 \frac{d^2Q_1}{dt^2} + \frac{Q_1}{C_1} - s\frac{dQ_2}{dt} &= e_1 \\ L_2 \frac{d^2Q_2}{dt^2} + \frac{Q_2}{C_2} + s\frac{dQ_1}{dt} &= e_2 \end{aligned}\right\} \quad \ldots \quad (5)$$

These two equations with the third terms of opposite sign are exactly similar to those for two mechanical systems with gyroscopic coupling; L_1 and L_2 replace the masses, Q_1 and Q_2 the displacements, e_1 and e_2 the forces, and $1/C_1$ and $1/C_2$ the stiffnesses; $-sdQ_2/dt$ and sdQ_1/dt with equal and opposite coefficients are the exact equivalents of what are called the gyrostatic terms. It was for this reason that Tellegen chose the name "gyrator" for the new network element.

To realize a system in which the reciprocity relation is violated Tellegen started from the equations

$$\left.\begin{aligned} Q_1 &= Cv_1 + Ai_2 \\ \Phi_2 &= Av_1 + Li_2 \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

On comparing these with equations (2) it is seen that the sign of the coupling factor A is now the same on both lines. The symbol Φ stands for flux-turns. On differentiating these equations we have

$$\left.\begin{aligned} i_1 &= C\frac{dv_1}{dt} + A\frac{di_2}{dt} \\ v_2 &= A\frac{dv_1}{dt} + L\frac{di_2}{dt} \end{aligned}\right\} \quad \ldots \quad \ldots \quad (7)$$

Fig. 4 shows a four-pole described by Tellegen which embodies these equations. The shaded rectangle represents a space between two plane electrodes filled with a medium, the properties of which are represented by the equations

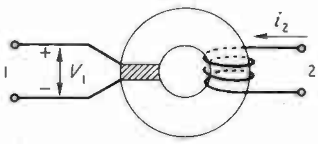


Fig. 4.

$$\left.\begin{aligned} D &= \epsilon E + \gamma H \\ B &= \gamma E + \mu H \end{aligned}\right\} \quad \ldots \quad \ldots \quad (8)$$

The yoke of magnetic material is assumed to have a very large permeability and to be wound with a coil the current in which is i_2. If the area of cross-section is S, the charge on the electrodes will be

$$Q_1 = SD = \epsilon SE + \gamma SH$$

and the flux-turns will be

$$\Phi_2 = nSB = \gamma nSE + \mu nSH$$

where n is the number of turns of the coil.

Although Tellegen does not mention it, the yoke must be assumed not only to have a large permeability, but also to be a dielectric of very small dielectric constant, otherwise it would form a short-circuit or a very large capacitance across the terminals 1.

If l is the distance between the electrodes, then $v_1 = lE$ and $i_2 = Hl/n$; hence

$$\left.\begin{aligned} Q_1 &= \frac{\epsilon S}{l} v_1 + \frac{\gamma n S}{l} i_2 \\ \Phi_2 &= \frac{\gamma n S}{l} v_1 + \frac{\mu n^2 S}{l} i_2 \end{aligned}\right\} \quad \ldots \quad \ldots \quad (9)$$

By putting $D = \epsilon_0 E + P$ and $B = \mu_0 H + J$, $\epsilon - \epsilon_0 = \kappa$ and $\mu - \mu_0 = \chi$ we obtain for the electric polarization P and magnetic polarization J

$$\left.\begin{aligned} P &= \kappa E + \gamma H \\ J &= \gamma E + \chi H \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

So what is required to carry out this scheme is a magnetic dielectric which is polarized both electrically and magnetically by either an electric or a magnetic field. The coefficient γ is a measure of the cross susceptibility; its dimension is the reciprocal of a velocity. If the molecules of a dielectric were both electric and magnetic dipoles, then, on applying either an electric or magnetic field, they would be oriented and P and J would both be increased. The device shown in Fig. 4 would then function as a gyrator. Tellegen calls $\gamma^2/\epsilon\mu$ the coupling coefficient and states that for the best results it should be as near unity as possible; in other words, γ should be the geometric mean of ϵ and μ, or perhaps more correctly of κ and χ. If P and J were due entirely to combined electric and magnetic dipoles, the ratio P/J would be independent of E and H and therefore from (10) $\kappa/\gamma = \gamma/\chi$ and $\gamma^2 = \kappa\chi$.

Up to this point we have discussed the gyrator as it existed in the mind of B. D. H. Tellegen in 1948. We propose to postpone to our next number the consideration of its subsequent development and practical realization.

G. W. O. H.

ANTI-RESONANT H.F. TRANSMISSION LINES

Input Impedance Characteristics

By Professor H. M. Barlow, Ph.D., M.I.E.E.

So many aspects of this subject have already been covered with such admirable thoroughness that it is surprising to find an exception justifying further discussion. The evaluation in a convenient form of maximum input resistance and reactance values for anti-resonant lines has apparently not been given all the attention it deserves and the purpose of this paper is to deal with that problem in the particular case of the short-circuited quarter-wavelength line.

Line Characteristics

A short-circuited line of length l with uniformly-distributed constants, characteristic impedance Z_0 and propagation coefficient $P = \alpha + j\beta$ has an impedance Z_s at input given by:—

$$Z_s = Z_0 \tanh Pl \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

For high frequencies we can assume with sufficient accuracy that Z_0 is purely real and (1) can therefore be re-written in terms of resistive and reactive components of Z_s as follows:—

$$Z_s = R_s + jX_s = \left(\frac{Z_0 \sinh \alpha l \,.\, \cosh \alpha l}{\sinh^2 \alpha l + \cos^2 \beta l}\right) + j\left(\frac{Z_0 \sin \beta l \,.\, \cos \beta l}{\sinh^2 \alpha l + \cos^2 \beta l}\right) \quad \ldots \quad (2)$$

MS accepted by the Editor, September 1951

Now for comparatively short lines whose length is comparable with the wavelength $\alpha l \ll 1$, so that $\sinh \alpha l \approx \alpha l$, and with slightly less accuracy $\cosh \alpha l \approx 1$. Thus we have:—

$$R_s/Z_0 = \frac{\alpha l}{\alpha^2 l^2 + \cos^2 \beta l} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

$$\text{and } X_s/Z_0 = \frac{\sin \beta l \,.\, \cos \beta l}{\alpha^2 l^2 + \cos^2 \beta l} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

These normalized resistive and reactive components of the impedance at input when plotted in terms of line length as a fraction of wavelength give curves, Fig. 1, which are well known and can be delineated over the greater part of the range neglecting the $\alpha^2 l^2$ term compared with $\cos^2 \beta l$ in the denominator of (3) and (4). When, however, l approaches $\lambda/4$ both terms in the denominator become significant.

It is convenient to plot the results for a line of given Q defined as

$$2\pi\left[\frac{\text{Maximum energy stored in the magnetic field}}{\text{Energy dissipated per cycle}}\right]$$

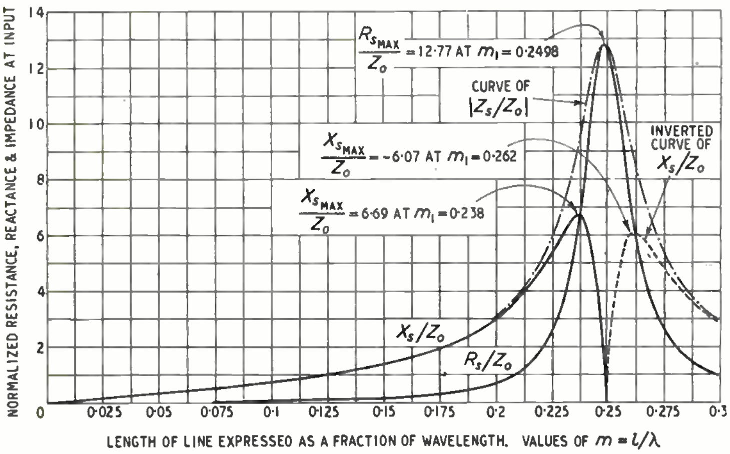
Neglecting any loss in the short-circuit at the end of the line we find:*

* See for example 'High Frequency Transmission Lines,' by Willis Jackson, Methuen Monograph, p. 100.

$$Q = \frac{\omega L}{R + GZ_0^2} \quad \cdots \quad (5)$$

where R, G and L represent resistance, leakance and inductance per unit length of line respectively.

We also know that at high frequencies $\alpha = \frac{(R + GZ_0^2)}{2Z_0}$ so that, using the values $Z_0 = \sqrt{\frac{L}{C}}$



and $\lambda f = \frac{1}{\sqrt{LC}}$ we have

$$Q = \frac{\pi}{\alpha\lambda} = \frac{\beta}{2\alpha} \quad \cdots \quad (6)$$

Maximum Value of Resistive Component of Input Impedance

Differentiating (3) with respect to l and equating to zero gives:—

$$\cos^2 \beta l + 2\beta l \sin \beta l . \cos \beta l = \alpha^2 l^2$$

and since for our present purpose we are only interested in values of βl closely approaching $\pi/2$ we can write $\sin \beta l = 1$ and neglect the term $\cos^2 \beta l$ in comparison with the others.

Thus we get:—

$$2\beta l . \cos \beta l = \alpha^2 l^2 \quad \cdots \quad (7)$$

Let $\quad l = m\lambda \quad \cdots \quad (8)$

where m approximates to 0·25 then $\beta l = 2\pi m$ and using (6) we can re-write (7) as:—

$$\cos 2\pi m = \frac{\pi m}{4Q^2} \quad \cdots \quad (9)$$

In solving equation (9) it is convenient to plot both sides for various values of m round about 0·25. This has been done in Fig. 2, and it will be seen that, since the left-hand side is independent of Q and is represented by a single curve of slope $\frac{d}{dm}(\cos 2\pi m) = -2\pi$ very nearly, within the range of interest, while for the right-hand side we have a family of curves corresponding to different Q values, the appropriate point of intersection for which $m = m_1$ is readily obtained.

Inserting in (3) the condition given by (7) for the maximum value of R_s and using (6) we have:—

$$R_{s,max}/Z_0 = \frac{1}{\alpha l(1 + 1/16Q^2)} \quad \cdots \quad (10)$$

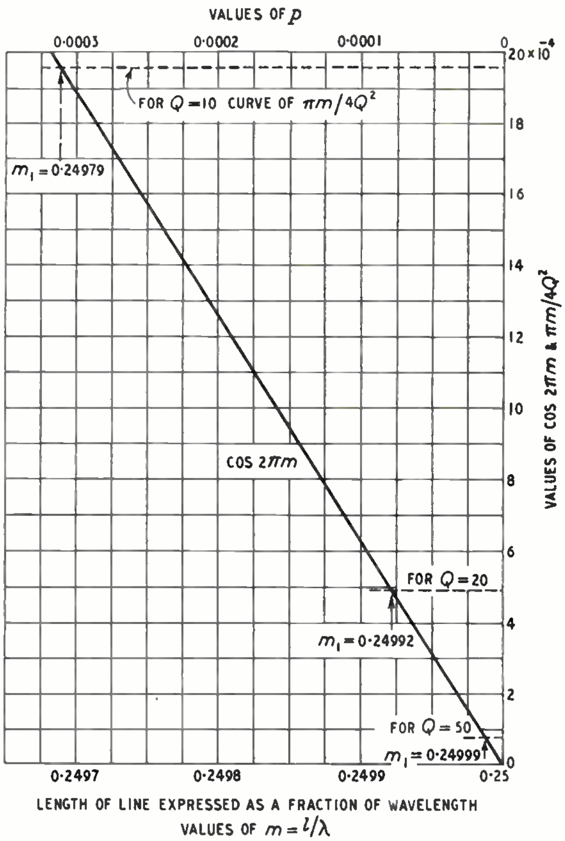
or so that even for Q values as small as (10)

$$R_{s,max}/Z_0 \approx \frac{1}{\alpha l} = \frac{Q}{\pi m_1} \quad (10a)$$

It is of interest to observe that the maximum value of the resistive component of the input impedance occurs when the length of line expressed as a fraction of the wavelength (i.e., m_1) is just less than 0·25, and becomes closer to 0·25 as the Q of the line is increased.

Fig. 1 (left). *Input impedance of an anti-resonant line with short-circuit termination and $Q = 10$.*

Fig. 2 (below). *Curves for solution of equation $\cos 2\pi m = \pi m/4Q^2$.*



For a given Q value the quantity $\pi m/4Q^2$ is very nearly represented over the range of interest by a horizontal straight line (see Fig. 2) of ordinate $\pi/16Q^2$ so that we have approximately $(0{\cdot}25 - m_1) = p_1 = 1/32Q^2$ defining the point at which $R_{s,max}$ occurs.

Maximum Value of Reactive Component of Input Impedance

For maximum reactance we differentiate (4) with respect to l and equate to zero giving:—

$$\beta l(1 + \alpha^2 l^2)\cos^2\beta l = \beta l(\alpha^2 l^2)\sin^2\beta l + 2(\alpha^2 l^2)\sin\beta l.\cos\beta l \qquad (11)$$

For values of βl approaching $\pi/2$ we have $\sin\beta l \to 1$ and $\cos\beta l \to 0$ so that with $\alpha^2 l^2 \ll 1$ the second term on the right-hand side of (11) is of smaller order than the other two terms and we can write:—

$$\cos^2\beta l \approx (\alpha^2 l^2)\sin^2\beta l$$

or

$$\pm \tan\beta l = 1/\alpha l \qquad (12)$$

Using (6) and (8) we find

$$\pm \tan 2\pi m = Q/m\pi \qquad (13)$$

In Fig. 3 both sides of this equation have been plotted for different values of m in the neighbourhood of 0·25 and the point of intersection at $m = m_1$, for a particular Q has been found. This curve relates to inductive values of $X_{s,max}$ for which $m_1 < 0{\cdot}25$ but it is repeated symmetrically about the vertical axis for capacitive values when $m_1 > 0{\cdot}25$. The points at which X_s reaches positive and negative maxima correspond to m_1 values significantly different from 0·25 especially at low Q.

Moreover the positive maximum representing an inductive reactance, which occurs with the shorter line, is larger than the negative maximum for the capacitive reactance. This is illustrated in Fig. 1, and will be apparent from equation (15).

From (12) we have

$$\cos^2\beta l = \frac{\alpha^2 l^2}{1 + \alpha^2 l^2} \approx \alpha^2 l^2 \qquad (14)$$

and the left-hand side of (13) has a slope

$$\frac{d}{dm}(\tan 2\pi m) = \frac{2\pi}{\cos^2 2\pi m} \approx \frac{1}{2\pi p^2}$$

for small values of p where $m = (0{\cdot}25 \pm p)$.

Alternatively $d/dm(\tan 2\pi m) \approx \dfrac{32Q^2}{\pi}$ at a point of intersection with the corresponding Q curve.

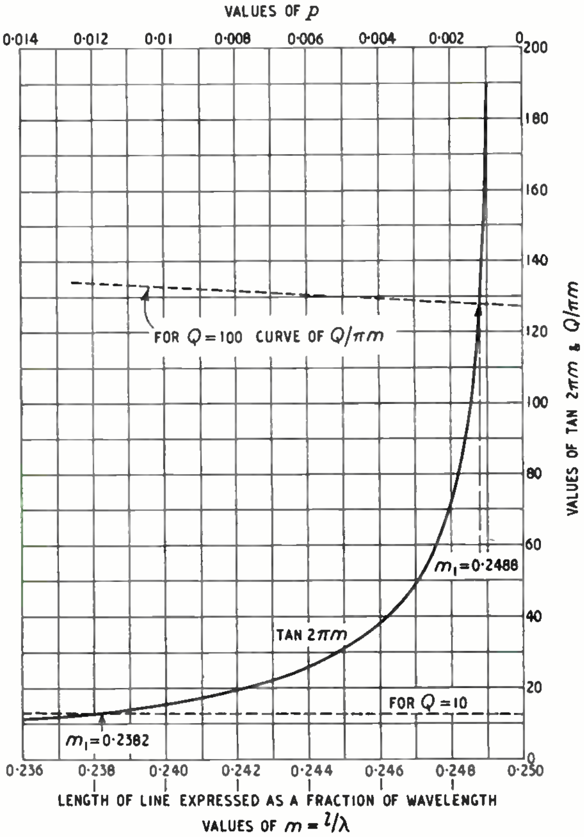


Fig. 3. *Curves for solution of equation tan $2\pi m = Q/\pi m$.*

The maximum value of X_s is found when (14) is inserted in (4) giving:—

$$X_{s,max}/Z_0 = \frac{1}{2\alpha l} \qquad (15)$$

Comparing (15) with (10a) it will be seen that the maximum value of the resistive component of the input impedance is approximately double the maximum value of the reactive component (see Fig. 1). The length of line at which these maxima occur is slightly different.

BEAT-FREQUENCY TONE SOURCE

Mathematical Theory of Mixing

By C. G. Mayo, M.A., B.Sc., M.I.E.E.

(Research Department, B.B.C. Engineering Division)

SUMMARY.—In a beat-frequency tone source it is required to obtain an output signal whose frequency is the difference of the frequencies of two signals, one of which is usually fixed and the other variable. Three methods are discussed:—*Method* 1: Multiplication in a square-law device. *Method* 2: Addition and linear rectification of the sum, it being understood that one signal has much greater amplitude than the other. *Method* 3: Addition of one component to a square wave synchronous with the other, and linear rectification of the sum.

1. Notation and Methods of Calculation

To facilitate calculation use will be made of Taylor's Theorem or the equivalent principle of the superposition of small variations. The distortion in which we are interested will, in a practical case, be of the order of 0·25% or 0·0025. If this is a first-order quantity, second-order quantities will be of the order 0·00000625 and will be neglected. Now by Taylor's Theorem if an output q is a function of several variables $x_1\, x_2 \ldots x_r$ the total change δq due to changes $\delta x_1\, \delta x_2 \ldots \delta x_r$ in $x_1\, x_2 \ldots x_r$ is to the first order

$$\delta q = \frac{\partial q}{\partial x_1}\delta x_1 + \frac{\partial q}{\partial x_2}\delta x_2 + \ldots . + \frac{\partial q}{\partial x_r}\delta x_r$$

2. Method 1: Multiplication in a Square-Law Device

Let the fixed oscillator signal be $E_1 \sin st$ and the variable $E_2 \sin (s + \omega)t$ and let the law of the device be $E = A + Bx + Cx^2$ where A, B and C are constant, and x is the applied signal. In the present case $x = E_1 \sin st + E_2 \sin (s + \omega)t$.

We then have

$$E = A + B[E_1 \sin st + E_2 \sin (s + \omega)t] + C[E_1 \sin st + E_2 \sin (s + \omega)t]^2$$

Now the only term in ωt is obtained from

$$E_3 = 2C\, E_1 \sin st . E_2 \sin (s + \omega)t = C\, E_1 E_2 [\cos \omega t - \cos (2s + \omega)t]$$

i.e., the beat-frequency output is proportional to the amplitude of both the fixed and variable oscillator voltages. If now the device has a law containing higher powers than the square, as will be usual in practice, output signals of frequency $n\omega/2\pi$ will appear, where n is an integer depending on the power referred to. Hence the freedom from harmonics will depend upon how exactly the multiplying device obeys a square law.

3. Method 2: Addition and Rectification

This case is treated by W. R. Bennett—"New Results in the Calculation of Modulation Products," *Bell System Technical Journal*, Vol. 12' 1933, pp. 228-243. His analysis is by means of a double Fourier series and the results are given in terms of elliptic integrals. His equation (9)

$$A_{mn} = A_{\pm mn} = \frac{2P}{\pi^2}\int_0^{\pi} \cos ny\, dy \int_0^{\arccos(-K \cos y)} (\cos x + K \cos y) \cos mx\, dx \quad (1)$$

gives the amplitude of the typical components of angular frequency $mp \pm nq$, for inputs $P \cos x$ and $PK \cos y$, where $x = pt$ and $y = qt$.

Imagine a three-dimensional diagram of

$$z = P \cos x + PK \cos y$$

against x and y. This diagram repeats along both x and y axes with a period 2π. All relevant information is contained in a rectangular section of the diagram between $-\pi$ and π on both x and y axes. If in this diagram all negative values of z are replaced by zero the effect is then like a rectangular plane with a rounded hill rising from it. The contour corresponding to the foot of the hill is given by the equation

$$z = P \cos x + PK \cos y = 0 \quad \ldots \quad (2)$$

Now $Z = [P \cos x + PK \cos y]$, where the square brackets indicate that all negative values are replaced by zero, corresponding to the effect of an ideal rectifier, can be represented by a Fourier series. The series is double, since there are two independent variables, and by symmetry contains only cosine terms. The series is

$$Z = \Sigma A_{mn} \cos mx \cos ny = A_{00} + A_{10} \cos x + A_{01} \cos y + A_{11} \cos x \cos y + \ldots\ldots..$$

and the coefficients are to be determined in a manner very similar to that used in a single Fourier series by the integral

$$A_{mn} = \frac{1}{\pi^2}\int_{\pi}^{\pi} Z \cos mx\, dx \int_{-\pi}^{\pi} \cos ny\, dy$$

In practice the fact that Z is zero for values of x and y which make $P \cos x + PK \cos y$ negative

MS accepted by the Editor, May 1951

is registered by proper choice of limits in the integrals as is usual also in single Fourier series.

Since $B_{mn} \cos mx \cos ny$

$$= \frac{B_{mn}}{2} \{\cos (mx - ny) + \cos (mx + ny)\}$$

the coefficient of the term containing

$\cos (mx \pm ny)$ is $B_{mn}/2$.

Now replace x by pt and y by qt so that the terms to be added and linearly rectified are

$$P \cos pt + PK \cos qt$$

(i.e., of angular frequencies p and q) then the components of angular frequency $mp \pm nq$ are given by $B_{mn}/2$ or by A_{mn} in the cited formula.

The above is not intended to be other than an outline of the paper cited which should be consulted for a full and rigorous discussion. This solution is valid for all values of K, and in general involves elliptic integrals.

The terms in which we are here interested are A_{11}, A_{22} and A_{33} giving the amplitude of the beat-frequency component and its second and third harmonics. A_{22} and A_{33} are not computed in the cited paper. When K is not small the harmonics are large and the practical case in which we are interested is when one of the signals is small compared with the other and K is of the order 0·01. In that case it is unnecessary to use elliptic integrals, and for uniformity A_{11}, A_{22} and A_{33} will be calculated approximately here.

From (1)

$$A_{mn} = \frac{2P}{\pi^2}\int_0^{\pi} \cos my\, dy \int_0^{\pi-\theta} (\cos x + \cos\theta) \cos mx\, dx$$

where $\cos\theta = K \cos y$

$$= \frac{2P}{\pi^2}\int_0^{\pi} \cos my\, dy \int_0^{\pi-\theta} \left\{ \frac{\cos (m+1)x}{2} + \frac{\cos (m-1)x}{2} + \cos\theta \cos mx \right\} dx$$

$$= \frac{P}{\pi^2}\int_0^{\pi} \cos my\, dy \left[\frac{\sin (m+1)x}{m+1} + \frac{\sin (m+1)x}{m-1} + 2\cos\theta \frac{\sin mx}{m} \right]_0^{\pi-\theta}$$

$$= (-1)^m \frac{P}{\pi^2}\int_0^{\pi} \cos my\, dy \left\{ \frac{\sin (m+1)\theta}{m+1} + \frac{\sin (m-1)\theta}{m-1} - 2\cos\theta \frac{\sin m\theta}{m} \right\}$$

Integrating by parts we get

$$(-1)^{m+1} \frac{P}{\pi^2}\int_0^{\pi} \frac{\sin my}{m} \left\{ \cos (m+1)\theta + \cos (m-1)\theta - 2\cos\theta\cos m\theta + \frac{2\sin\theta \sin m\theta}{m} \right\} \frac{d\theta}{dy}\, dy$$

$$= (-1)^{m+1} \frac{2P}{\pi^2}\int_0^{\pi} \left\{ \frac{\sin my}{m^2} (\sin m\theta)\, (K \sin y) \right\} dy \qquad \ldots \quad (3)$$

If in equation (3) $m = 1$ then

$$A_{11} = +\frac{2P}{\pi^2}\int_0^{\pi} (\sin y)(K \sin y) \sqrt{1 - K^2\cos^2 y}\, dy$$

$$= \frac{PK}{\pi} \text{ to first order of } K$$

Similarly,

$$A_{22} = -\frac{2P}{\pi^2}\int_0^{\pi} \frac{\sin 2y}{4} \left\{ 2\sqrt{1 - K^2\cos^2 y}(K^2 \cos y)(\sin y) \right\} dy$$

$$= -\frac{PK^2}{4\pi} \text{ to second order of } K$$

and

$$A_{33} = \frac{2P}{\pi^2}\int_0^{\pi} \frac{\sin 3y}{9} \left\{ (\sin 3\theta)(K \sin y) \right\} dy$$

$$= \frac{PK^3}{8\pi} \text{ to third order of } K$$

Thus the fundamental beat-frequency output has amplitude PK/π, the second-harmonic amplitude is $PK^2/4\pi$ and the ratio of second-harmonic amplitude to fundamental amplitude is $K/4$. The third harmonic is of the third order in K. The amplitude of the beat-frequency term is proportional to the amplitude of the smaller input component.

4. Method 3: Addition of a Square Wave and a Sine Wave

4.1. *General*

In this method a square wave synchronous with the variable frequency is added to the fixed-frequency sine-wave output and the sum is linearly rectified. The sine-wave peak value is less than that of the square wave and thus the sum is positive only when the square wave is positive. Thus the positively rectified signal consists of the positive square wave plus the segments of the sine wave which coincide in time with the positive square wave. Thus the square wave acts as a switch; when it is positive the sine wave appears in the output but not when it is negative. This is shown in Fig. 1. The operation of a switch is equivalent to multiplication by unity when the switch is closed and by zero when the switch is open and so the rectified result is equal to the square-wave positive half cycles plus the product of a 1, 0, 1, 0, square wave by a sine wave. Now the Fourier series of a square wave consists of fundamental and odd harmonics only. If such a square wave containing angular frequencies s, $3s$, $5s$, etc., is multiplied by a sine wave of angular frequency $(s - \omega)$ the resultant frequencies are

$$s - \omega \pm s, \quad s - \omega \pm 3s, \quad s - \omega \pm 5s, \text{ etc.}$$

Of these only one namely ω is an audio frequency —the rest are supersonic and in particular 2ω, 3ω, etc., are absent. Thus the output is free from harmonics of the fundamental beat frequency.

The effect is very similar to the scanning of a wave by a slit, as in film recording. As discussed above, a positive pulse added to a signal and the sum rectified acts as a switch and only those parts of the wave coincident in time with the scanning pulse reach the output. It is true that the pulses themselves are also present in the output but their time integral per cycle is constant and it is not worth while taking steps to cancel them out although this would not be difficult. If the positive pulses are made very narrow the effect is to select equidistant ordinates of the scanned wave like a repeated Dirac δ-function.

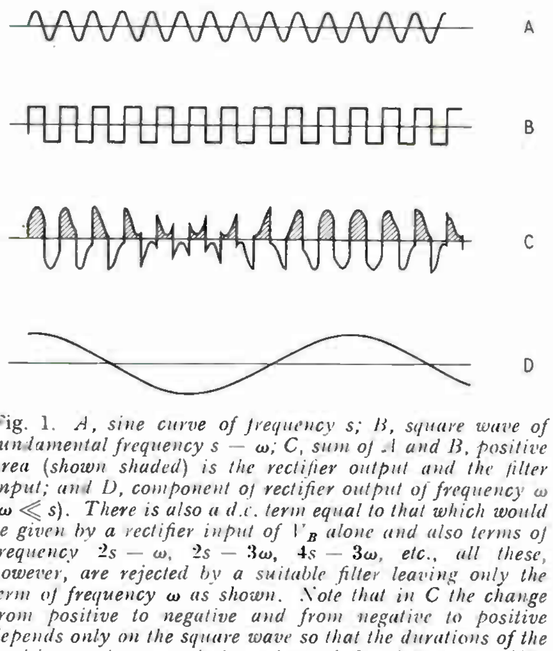


Fig. 1. A, sine curve of frequency s; B, square wave of fundamental frequency $s - \omega$; C, sum of A and B, positive area (shown shaded) is the rectifier output and the filter input; and D, component of rectifier output of frequency ω ($\omega \ll s$). There is also a d.c. term equal to that which would be given by a rectifier input of V_B alone and also terms of frequency $2s - \omega$, $2s - 3\omega$, $4s - 3\omega$, etc., all these, however, are rejected by a suitable filter leaving only the term of frequency ω as shown. Note that in C the change from positive to negative and from negative to positive depends only on the square wave so that the durations of the positive sections are independent of the sine wave. If it were not so there would be harmonic distortion of the output D as shown mathematically in the text.

If the results were displayed on an oscilloscope arranged to make the sine wave stationary, the pulse would be seen to scan the sine wave. The integrated results would be an exact copy of the scanned wave but at beat frequency. This is shown diagrammatically in Fig. 2.

With a narrow rectangular pulse as scanning wave the output would be proportional to the area under the scanned wave of a breadth equal to the width of the rectangular pulse. With a square wave the width of the scanning pulse would be approximately a half-cycle integral of the scanned wave whether this is a sine wave or not. For example, if the scanned wave were square the output would be a triangular wave. In harmonic terms if the scanned wave is arranged to have a specific harmonic content the output would contain the same harmonics, but with amplitude reduced in proportion to harmonic number. Even harmonics, however, would be absent. (Absence of even harmonics is a specific feature of half-cycle integration.)

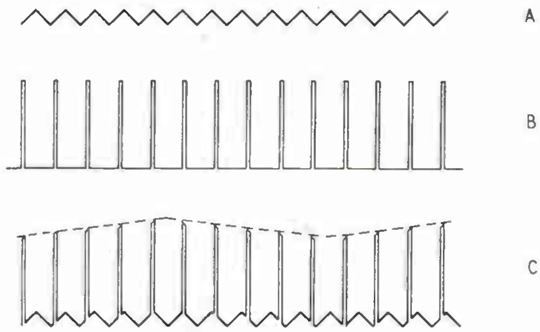


Fig. 2. A, scanned wave of repetition frequency s; B, scanning pulse of repetition frequency $s - \omega$; and C, sum of scanned wave and scanning pulse. Scanned wave is copied at repetition frequency ω, by peak envelope or by rectified outputs. (Note, $\omega \ll s$).

4.2. *Case of a Nearly Square Wave*

Proceeding as in Method 2 above, let us calculate the beat-frequency output and harmonic distortion terms arising when one component is a pure sine wave and the other is a nearly square wave. Suppose the sine wave input is $\gamma \cos y$ and the nearly square wave is $f(x)$ where

(1) $f(x)$ has period 2π, and
$$f(x + \pi) = -f(x)$$

(2) $f\left([2r + 1]\dfrac{\pi}{2}\right) = 0$ when r is an integer

(3) when $\left|x - \dfrac{\pi}{2}\right| \leqslant \dfrac{\gamma}{\lambda}$, $f(x) = \lambda\,[(\pi/2) - x]$

(4) when $\dfrac{\gamma}{\lambda} \leqslant \left|x - \dfrac{\pi}{2}\right| \leqslant \dfrac{\pi}{2}$, $\left|f(x)\right| \geqslant \gamma$

In Fig. 3(a) is shown a trapezoidal waveform which just meets these requirements. Other waveforms which do so are shown in Fig. 3(b) and (c); the waveform in Fig. 3(c) is $\lambda \cos x$ in which the unshaded part differs from Fig. 3(a) only in the third order of γ/λ.

Now consider

$$B_{mn} = \frac{1}{2\pi^2}\int_0^{\pi} \cos ny\, dy \int_0^{\theta} \{f(x) + \gamma \cos y\} \cos mx\, dx$$

B_{mn} is the amplitude of the Fourier component of frequency $(mx + ny)/2\pi$ of the positive rectified part of $f(x) + \gamma \cos y$ provided that the limit θ is chosen so that $f(x) + \gamma \cos y = 0$, in order that negative values of this expression may be excluded.

But from condition (3) governing $f(x)$, we have

$$\lambda\,[(\pi/2) - \theta] + \gamma \cos y = 0$$

therefore $\theta = (\pi/2) + (\gamma \cos y)/\lambda$

therefore $\cos \theta = -(\gamma \cos y)/\lambda$

We now seek to identify B_{mn} with A_{mn} in (1).

We first notice that if the limit θ is replaced by any quantity independent of y, and $n \neq 0$, B_{mn} is independent of $f(x)$. Also, if x is between $(\pi/2) - (\gamma/\lambda)$ and $(\pi/2) + (\gamma/\lambda)$

$$f(x) = \lambda[(\pi/2) - x] = \lambda \cos x$$

to the first order in x, that is to say, neglecting a quantity of order $(\gamma/\lambda)^2$. Hence the range 0 to θ can be divided up into two ranges, namely (i) 0 to $\pi/2$, in which $f(x)$ may be replaced by anything, including $\lambda \cos x$, (ii) $\pi/2$ to θ, in which $f(x)$ can be replaced by $\lambda \cos x$. Hence

$$B_{mn} = \frac{\lambda}{2\pi^2}\int_0^{\pi} \cos ny\, dy \int_0^{\theta} [\cos x + (\gamma/\lambda) \cos y] \cos mx\, dx$$

where $\cos \theta = -(\gamma \cos y)/\lambda$, which is of the same form as A_{mn} with (γ/λ) for K.

Hence the fundamental beat-frequency amplitude is γ/π, the second-harmonic amplitude is $\gamma^2/4\pi\lambda$, and the ratio of the second harmonic to the fundamental is $\gamma/4\lambda$, higher harmonics being of higher order in γ.

It is thus seen that the only important aspect of the nearly square wave for the determination of harmonics is the gradient near zero, and that the ratio of the second harmonic to the fundamental is one-quarter of the ratio of the slope near zero of the sine wave to the slope near zero of the scanning wave. It is also seen that the signal output is proportional only to the amplitude of the sine-wave input.

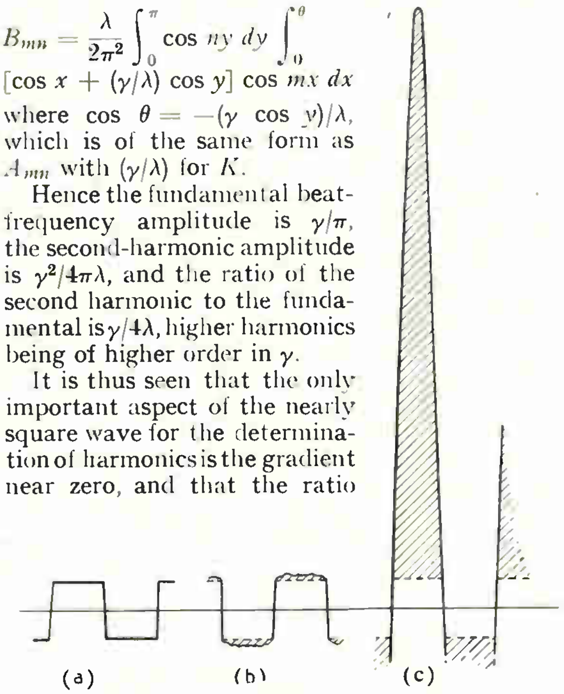
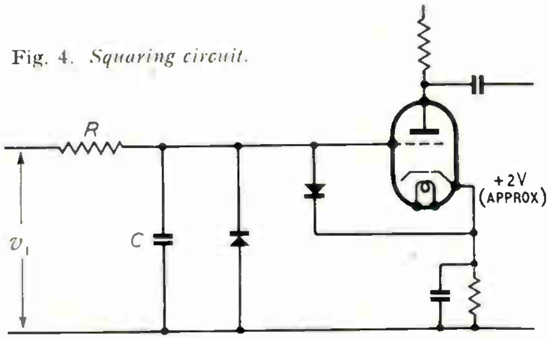


Fig. 3. *Various forms of $f(x)$. In each case the shaded area makes no contribution to the integral*

$$B_{mn} = \int_0^{\pi} \cos ny\, dy \int_0^{\cos^{-1}(-K\cos y)} \{f(x) + K \cos y\} \cos mx\, dx,\ (n \neq 0).$$

Fig. 4. *Squaring circuit.*



4.3. *Distortion due to Finite Size of Limited Wave*

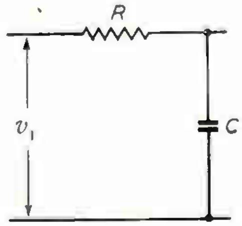
If the square wave is obtained by limiting by diodes a sine wave of amplitude λ, then, assuming no deterioration due to capacitance in the limiting circuit, the gradient near zero is λ, so that Section 4.2 applies and the second-harmonic percentage is $25\gamma/\lambda$.

4.4. *Distortion due to Capacitance across the Squaring Circuit*

The squaring circuit will be essentially as in Fig. 4. The sine wave input voltage is $v_1 = \mu \sin \theta = \mu \sin \sigma t$ where μ is of the order 25 V. The series resistance R is of the order 10,000 ohms and the total capacitance across the output including diode and grid capacitance is C, approximately 10 pF.

In the absence of capacitance the grid voltage v_2 is $\mu \sin \theta$ limited by the diodes between $\pm\, \delta$ V where $\delta \approx 1$. The amplifying valve has a mutual conductance g mA/V (approximately 5 mA/V), so that the anode current in the absence of capacitance is $\mu g \sin \theta$ limited by the diodes to $\pm\, g\delta$ mA.

Fig. 5. *Simplified squaring circuit when both diodes are non-conducting*



Consider what happens when $v_1 < -\delta$ and increasing. At that time $v_2 = -\delta$ and the negative diode is conducting. When v_1 reaches $-\delta$ the current in R changes sign and the diode current stops. The voltage v_2 begins to increase. From this time until the positive diode limits at $+\,\delta$ V the circuit is as Fig. 5. Solving this circuit in the usual way and using the notation of van der Pol,

$$v_2 \doteq v_1 \frac{CR}{1 + pCR}$$

Now over the range of interest $|v_2| < \delta$ and $|v_1|$ is also of the same order.

Hence $v_1 = \mu \sin \theta$

$$= \mu \left\{ \theta - \frac{\theta^3}{6} + \frac{\theta^5}{120} - \dots\dots \right\} \approx \pm \delta$$

where $\delta \approx 1$ and $\mu \approx 25$ and $\sin \theta \approx 0{\cdot}04$.

Thus to the second order we can write

$$v_1 = \mu\theta = \mu\sigma t \doteq \mu\sigma \frac{1}{p}$$

Hence $v_2 \doteq \mu\sigma \dfrac{1}{p} \cdot \dfrac{CR}{1 + pCR}$

Therefore $v_2 = \mu\sigma \{t - CR\,(1 - e^{-(t/CR)})\}$

Now this voltage v_2 is the positive voltage relative to the initial quasi-steady state in which v_2 is $-\,\delta$ V, so that the grid voltage relative to earth is then

$$v_2 = \mu\sigma \{t - CR\,(1 - e^{-(t/CR)}) - \delta\}$$

It is desirable to express this in terms of time from the instant at which $v_2 = 0$, so that the slope near zero may be evident. To do this suppose $v_2 = 0$ at $t = t_0$ and replace t by $t_0 + \delta t$.

Then $v_2 = \mu\sigma\, \delta t (1 - e^{-(t_0/CR)})$

Thus the rate of change near zero is

$$\mu\sigma\,[1 - e^{-(t_0/CR)}]$$

compared with a slope $\mu\sigma$ in the absence of capacitance.

Thus the second-harmonic distortion will be

$$\frac{\gamma}{4\mu} \frac{1}{[1 - e^{-(t_0/CR)}]} \quad \text{instead of} \quad \frac{\gamma}{4\mu}$$

i.e., greater in the ratio $1/[1 - e^{-(t_0/CR)}]$

In a particular case

$$\mu = 25,\ C = 10 \times 10^{-12}\ \text{F},$$

$$R = 10^4\ \Omega,\ \frac{\gamma}{\mu} = 0{\cdot}01$$

$$\text{and } 1 - e^{-(t_0/CR)} = 0{\cdot}65$$

so that second-harmonic distortion is increased in the ratio $1/0{\cdot}65 = 1{\cdot}54$ by the effect of capacitance in this part of the circuit.

4.5. *Distortion due to Capacitance across Mixing Diodes*

As stated in Section 1 it is permissible to regard the various causes of distortion as acting independently, the mutual effect being of a higher order of small quantities. In computing the distortion due to the mixing diodes it is therefore permissible and convenient to assume an ideal square wave on the grid of the final squaring valve. Thus the anode current of this valve may be written

$$i_2 = f(\theta)$$

when (1) $f(0) = 0$

(2) $f(\theta + \pi) = -f(\theta)$

(3) $f'(\theta) \to \infty$ as $\theta \to 0$

(4) $|f(\theta)| > \gamma$

where the fixed-frequency current i_1 is

$$i_1 = \gamma \sin(\phi + \theta)$$

and the total current is

$$i_t = i_1 + i_2 = f(\theta) + \gamma \sin(\phi + \theta)$$

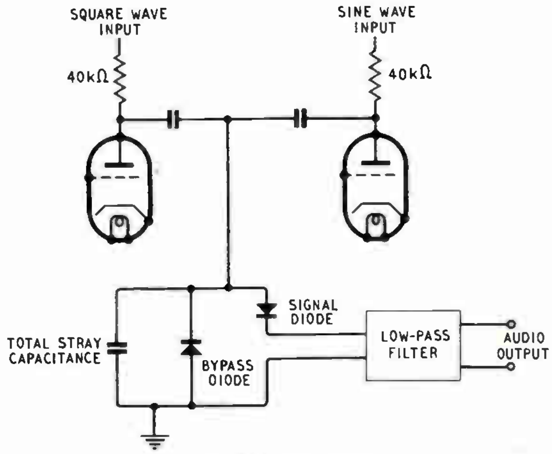


Fig. 6. *Mixing circuit.*

In the absence of shunt effects this current necessarily goes through one or other of the two diodes, namely the positive or signal diode and the negative or by-pass diode, as shown in Fig. 6. The current will change from one diode to the other exactly at its zero and the signal diode current will be the integrated positive part of i_t between limits set by the zeros of $f(\theta)$. Since $f(\theta)$ is a square wave the zeros are not deviated by i_1 and there is no harmonic distortion. Owing, however, to shunt capacitance across the diode circuit the rectifier current differs from the current i_t by the current necessary to charge the stray capacitance across the diodes to the diode back voltage.

Now when i_t is negative the diode voltage is very small. When i_t becomes positive the current in the negative diode stops and in the absence of capacitance the signal diode current immediately starts. There will be, however, generally a signal back voltage on the filter input, and the signal diode current cannot start until the forward diode voltage exceeds the back voltage. If there is a capacitance C across the diodes and the signal voltage is v volts, then the charge on the capacitance is Cv coulombs. At the end of the positive half wave the signal rectifier current stops. During the whole positive half cycle the current i_t has either gone into the signal diode or has charged the capacitor C so that the defect of the integrated current q due to the capacitance C is the charge on this capacitor at the end of the positive half cycle, that is

$$\delta q = C.v_3$$

Now v_3 differs from the signal instantaneous voltage only by the diode drop which will be considered later, and thus

$$\delta q \approx C.v_f$$

where v_f means the filter instantaneous voltage at the end of the positive half cycle.

Now by hypothesis the output is not distorted (i.e., contains no terms in $\sin 2\phi$, $\sin 3\phi$, etc.) and δq is a linear function of the output and thus produces no distortion. In effect the capacitance C acts as if it were a partial (linear) shunt across the output.

This may be checked by expressing the signal rectifier current as a Fourier series, and from this and the filter input impedance $Z(i\omega)$ computing the filter input voltage. It will be found to be a linear function of $\sin \phi$ and $\cos \phi$. As $\delta q = C.v_f$ is proportional to v_f, δq is also a linear function of $\sin \phi$ and $\cos \phi$ and this introduces no terms in $\sin^2 \phi$, etc., which would give rise to harmonic distortion terms.

To get a quantitative idea of this shunting effect we note that

$$q \approx \frac{1}{\sigma}\int_0^{\pi} [f(\theta) + \gamma \sin(\phi + \theta)]\, d\theta \text{ millicoulombs}$$

(where γ is in mA, $\theta = \sigma t$, $\sigma = 2\pi f \approx 10,^6$ and the integrand is replaced by zero if negative). Thus

$$q = A_0 + \frac{2\gamma \cos\phi}{\sigma} \text{ where } A_0 \text{ is independent of } \phi.$$

If the filter impedance is taken at 1,000 ohms, and C as 6 pF, the voltage

$$v_f \approx \frac{2\gamma}{\pi}\cos(\phi + \beta)$$

and $|\delta q| \approx C.v_f$

$$= 12(\gamma/\pi) \times 10^{-12} \cos(\phi + \beta) \text{ coulombs}$$

while $|q - A_0| = (2\gamma \cos\phi)/\sigma$

so that the required relative shunting effect is

$$\left|\frac{\delta q}{q - A_0}\right| = \frac{6\sigma}{\pi}\frac{\cos(\phi + \beta)}{\cos\phi} \cdot 10^{-9}$$

$$\approx 0{\cdot}002 \text{ or } 0{\cdot}2\%$$

if we neglect β, the phase angle of the impedance of the filter imput with $C = 6$ pF.

4.6. *Effect of Non-Linear Diodes*

It has already been seen that the only critical part of the grid-voltage waveform as regards distortion is near the zero of voltage. Under these conditions neither diode is conducting. Thus non-linearity of the diodes in the squaring circuit has no direct effect on distortion and makes the wave other than square only where exact squareness is of no importance.

To determine the effect of non-linear diodes on the mixing circuit, suppose the diode voltage/current relation is given by

$$v_d = F(i_d)$$

where v_d is the diode voltage and i_d is the diode current. Suppose the total current

$$i_t = i_1 + i_2 = \gamma \sin(\phi + \theta) + f(\theta)$$

where $f(\theta)$ represents a square wave of peak value $\pm g\delta$, $g \approx 5$ mA/V, $\delta \approx 1$ V and γ is less than 2·5 mA. Then the rectified integral current in the absence of shunt losses is

$$q_0 = \frac{1}{\sigma}\int_0^{\pi} \{g\delta + \gamma \sin(\phi + \theta)\}\, d\theta$$

$$= \frac{2\gamma}{\sigma}\cos\phi \text{ millicoulombs.}$$

Dropping constant terms and taking $\sigma = 10^6$,

$$q_0 = 2\gamma \cos\phi \times 10^{-9} \text{ coulombs.} \quad .. \quad .. \quad (4)$$

Now the diode back voltage $v_d = F(i_d)$ will give rise to two variations of q. The first, $\delta_1(q)$, is due to the stray capacitance C across the mixing circuit and the second, $\delta_2(q)$, is due to the shunting effect of the anode resistances of the output valves. The diode current at the end of the positive half cycle is $i_d = (g\delta - \gamma \sin\phi)$ and the charge on the stray capacitance is thus

$$\delta_1 q = C\,F\,(g\delta - \gamma \sin\phi) \text{ coulombs} \quad .. \quad (5)$$

The current lost through the anode resistances (each 40,000 ohms) is

$$\delta_2 q = \frac{1}{\sigma}\int_0^{\pi} \frac{Fg[\delta + \gamma \sin(\phi + \theta)]}{20{,}000}\, d\theta \quad .. \quad (6)$$

Now consider $v_d = F(i_d)$.

In Table 1 are given measured values of voltage and current for a specimen germanium rectifier. From these figures are derived $F(i_d)$ and various differential coefficients* at $i_d = 5$ mA.

$$F(5) = 0{\cdot}697 \qquad F_3(5) = -0{\cdot}0012$$
$$F_1(5) = 0{\cdot}0665 \qquad F_4(5) = 0{\cdot}0024$$
$$F_2(5) = -0{\cdot}0073$$

where $F_r(5)$ means $\dfrac{d^r F(i_d)}{d{i_d}^r}$ with $i_d = 5$

$$\text{Now } \delta_1(q) = C\,F\,(g\delta - \gamma \sin\phi) \quad .. \quad .. \quad (7)$$

* *Note.* It should be remarked that these coefficients are the necessary and useful data to know about any empirical curve such as those of a valve. Generally the best way to determine them is not by differentiating the static curve, but by direct measurement. It would be most useful if valve data included curves of:

(1) i_a as function of V_g. (The usual steady-state curve.)
(2) $\partial i_a/\partial v_g$; i.e., g or slope. (As measured on the mutual-conductance bridge.)
(3) $\partial g/\partial v_g$: (g); i.e., curvature κ giving 2nd harmonic
(4) $\partial\kappa/\partial v_g$ giving 3rd harmonic.
(3) and (4) could be measured by a harmonic analyser.)

and putting $g\delta = 5$

$$\delta_1(q) = C\,F\,(5 - \gamma \sin \phi)$$

$$= C\left\{F(5) - \gamma \sin \phi\, F_1(5) + \frac{\gamma^2}{2} \sin 2\phi\, F_2(5) - \frac{\gamma^3 \sin 3\phi}{6} F_3(5)\right\}$$

and putting $C = 10$ pF and selecting only terms in 2ϕ and 3ϕ

$$\delta_1(q) = 10^{-11}\,[(0{\cdot}001825\,\gamma^2 - 0{\cdot}0001\,\gamma^4) \cos 2\phi + 0{\cdot}00005\,\gamma^3 \cos 3\phi] \text{ coulombs}$$

with $q_0 = 2\gamma \cos \phi \times 10^{-9}$ coulombs.

The ratio of second harmonic is $0{\cdot}000009\,\gamma - 0{\cdot}0000005\,\gamma^2$ and the ratio of third harmonic is $0{\cdot}00000025\,\gamma^2$. These are quite negligible.

As regards

$$\delta_2(q) = \frac{1}{20000}\int_0^{\pi} F\,[5 + \gamma \sin(\phi + \theta)]\, d\theta$$

the lowest harmonic is the third, since even harmonics vanish in the integration. The third harmonic is approximately

$$\delta_2(q) = \frac{1}{200000}\int_0^{\pi} \gamma^3 \sin^3(\phi + \theta)\,\frac{F_3(5)}{6}\, d\theta$$

and the ratio to the fundamental q_0 is $8{\cdot}3\,\gamma^2 \times 10^{-7}$ which is also negligible.

TABLE 1

mA	Volts	mA	Volts	mA	Volts
0·01	0·058	2·0	0·462	5·0	0·698
0·05	0·11	2·5	0·51	5·5	0·730
0·1	0·145	3·0	0·555	6·0	0·762
0·5	0·265	3·5	0·594	6·5	0·785
1·0	0·350	4·0	0·630	7·0	0·817

5. Best Performance Obtainable

It has been seen that with a square wave of approximately $\pm$ 5 mA and a sine wave of fixed frequency of peak value of the order 2·5 mA or less, the harmonic distortion may be extremely small provided the rate of change of grid volts of the square-wave output valve is great enough. With a fixed-frequency current of 2·5 mA the filter input signal is of the order 1·25 V peak on a 2000-ohm filter. Thus this low harmonic content is not obtained at the cost of low output. The practical limiting feature so far is that of securing a sufficiently high rate of change of grid voltage across zero in the square-wave output valve.

It has been seen (4.4) that with a 10,000-ohm resistance feed to the squaring diodes and a 25-V supply of signal, the rate of change of grid voltage λ is $\lambda = 0{\cdot}65 \times 25\,\sigma$

where the input signal is $25 \sin \sigma t$ and $\sigma \approx 10^6$. Thus $\lambda \approx 16{\cdot}25$ V/μsec.

Now if this figure is to be substantially increased it will be by charging the grid and stray capacitance more quickly; i.e., with greater current starting more quickly. Positive feedback would only make available the total current-handling capacity of the valve used, and could not therefore cause any spectacular improvement. It will be useful to compute the improvement to be gained by using an additional squaring valve. Suppose this additional valve is similar to the first, operating at 5 mA anode current and with a mutual conductance of 5 mA/V. The circuit will be as in Fig. 7.

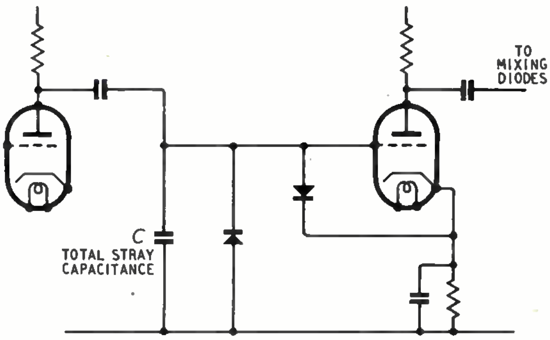


Fig. 7. *Second squaring circuit.*

As already computed (4.4) with an initial sine wave input of 25 V the grid voltage over the critical range is

$$v_g = 0{\cdot}65 \times 25\,\sigma t = 16{\cdot}25$$

and the anode current

$$i_{a_1} = 81{\cdot}25\,\sigma t \text{ mA}$$

On the second diode circuit the total capacitance is approximately

Anode of V_1	2 pF
Grid of V_2	6
Diodes	2
Total	10 pF

Hence the second grid voltage v_{g_2} is

$$v_{g_2} \doteqdot 81{\cdot}25\,\sigma \,.\, \frac{1}{p} \,.\, \frac{10^{12}}{10p}\,\frac{1}{1000}$$

$$= 81{\cdot}25\;10^{14}\,\frac{t^2}{2} \text{ volts}$$

This is, as before, based on the initial state of -1 V so that the grid voltage to earth is

$$x = 40{\cdot}625 \times 10^{14}\,t^2 - 1$$

and is zero when $t \approx 1{\cdot}6 \times 10^{-8}$ and then

$$\frac{dx}{dt} = 81{\cdot}25 \times 10^{14} t = 130 \times 10^6$$

or 130 V/μsec

compared with 16·25 V/μsec on the first grid.

Thus, other things being unchanged, an additional valve could give a reduction of distortion in the ratio $\frac{16{\cdot}25}{130}$ or 0·125 or with fixed-frequency current of 1 mA the second-harmonic ratio to fundamental would be

$$\frac{1}{4 \times 130 \times 5} \text{ or } < 0{\cdot}05\%$$

6. Conclusions

Of the three methods discussed, namely

(1) Multiplication in a square-law device,

(2) Addition and linear rectification of the sum,

(3) Addition of one component to a square wave synchronous with the other component and linear rectification of the sum,

the last is by far the best. The output and freedom from distortion obtained by this method with valve anode currents of only 5 mA would require signals of the order 100 mA if method 2 were used. Method 3 also results in reduced mixer noise and requires less filtration to remove h.f. components from the output.

Acknowledgments

The foregoing analysis was undertaken with reference to a portable tone source[1] developed for use in room acoustics by the B.B.C. Engineering Research Department.

The author is indebted to the Chief Engineer of the B.B.C. for permission to publish this article.

REFERENCE

[1] C. G. Mayo and D. G. Beadle. "Equipment for Acoustic Measurements," *Electronic Engineering*, October 1951, p. 368.

CATHODE-FOLLOWER OPERATION

Transient and Steady-State Performance with a Capacitive Load

By A. J. Shimmins, B.E.E., B.Com.

(E.M.I. Engineering Development-Advanced Development.)

1. Introduction

IT is proposed to give consideration to the operation of cathode-follower circuits with capacitive loads. In a previous paper[1] consideration was given to the operation of these circuits under pulse and sawtooth conditions. It was shown that with a large capacitive load, grid current can flow for a short time, when a steep-fronted waveform is applied to the grid. This prevents complete isolation between the input circuit and the load, and makes the output voltage of exponential form of time constant equal to the output impedance multiplied by the capacitance of the load.

In this paper it is proposed to give consideration to methods of improving the performance of cathode-follower circuits both under steady-state and transient conditions and three approaches have been considered, viz:

(a) The use of an inductance in series with the capacitive load.

(b) The use of a filter circuit as the cathode load, based on the fact that a low-pass π-filter with a capacitance in parallel with the input has a constant input impedance over a wide frequency band.

(c) The use of increased capacitance between grid and cathode effectively to transfer part of the cathode-load capacitance into the input circuit.

The main interest has been in attempting to improve the transient response of the cathode-follower, particularly by the use of inductive coupling circuits similar to those which have been successfully used to improve the performance of wideband amplifiers, but steady-state performances, particularly the cases of a series inductance and an ideal filter load are also considered.

It is shown that with a simple series inductance, the 10-90% rise time of the output voltage is improved by a factor up to 1·5, with an overshoot less than 5%. This is accompanied by a corresponding increase in the steady-state bandwidth. The presence of an inductance between cathode and anode load also modifies the cathode-voltage waveform, resulting in a reduced flow of grid current if the grid is driven positive with a step-function input.

With an ideal filter load in the cathode circuit, the bandwidth under steady-state conditions is

MS. accepted by the Editor, July 1951

increased, but the transient performance is difficult to determine and experiments show that the overshoot is objectionable which, together with its relative complexity, makes the circuit of no practical value.

The effect of the grid-cathode capacitance on the transient performance is then considered, and it is shown that with a step function input, the grid-cathode capacitance results in an initial step in the output voltage, followed by a subsequent exponential rise. The initial step can be increased by increasing the grid-cathode capacitance but this is accompanied by an exponentially decreasing input current to charge the grid-cathode capacitance, thus losing the high input impedance of the cathode-follower.

2. Equivalent Circuits and Superposition Integrals

2.1. *Equivalent Circuit for Large Grid–Cathode Impedance*

Before discussing the performance of cathode followers the equivalent circuits will be considered. If the valve characteristics are assumed to be straight and parallel and equidistantly spaced and, further, if the input signals are sufficiently small to remain on the linear part of the characteristics, the equivalent circuit is as shown in Fig. 1(b). This can be simplified to the circuit shown in Fig. 1(c), in which the input and output circuits are isolated, the input impedance being a resistance $R_g/(1 - A')$, in parallel with a capacitance $C_{ga} + (1 - A)C_{gk}$. Where R_g is the grid-leak resistor and A' is the gain from the grid to the bias point to which the grid is returned, and A is the gain from grid to cathode (both less than unity), and C_{ga} and C_{gk} are as shown.

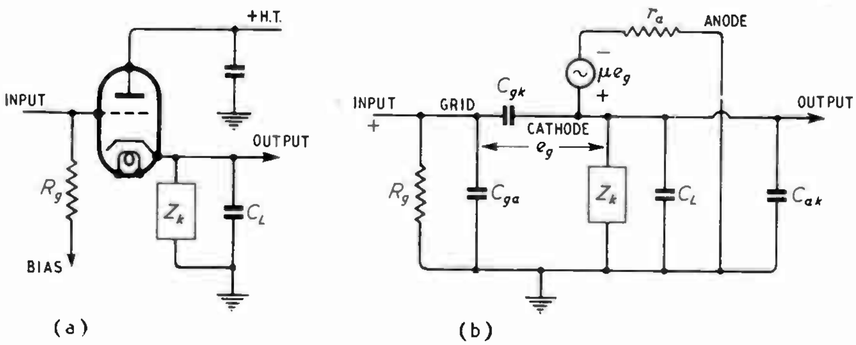
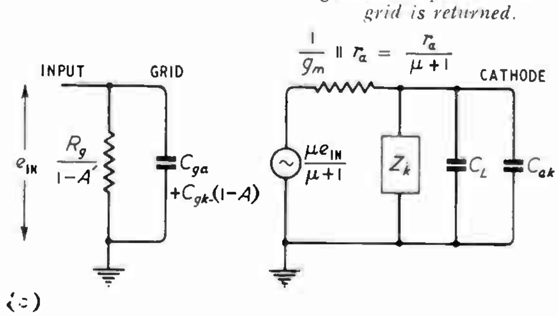


Fig. 1. *Simple cathode-follower circuit (a) and equivalent circuits (b) and (c); μ = amplification factor, g_m = mutual conductance grid to anode, r_a = anode resistance, C_{ga} = grid-anode capacitance, C_{ak} = anode-cathode capacitance, C_L = load capacitance, A = gain from grid to cathode, A' = gain from grid to bias point to which grid is returned.*

If the effects of C_{gk} and R_g are small, the output circuit can be represented by a generator giving a voltage $\mu e_{in}/(\mu + 1)$, and of internal impedance $r_a/(\mu + 1)$, which is the resistance of $1/g_m$ in parallel with the anode resistance r_a. This generator feeds the cathode load comprising the cathode impedance Z_k, the load capacitance C_L and the anode-cathode capacitance C_{ak}, which can be considered as being part of the load capacitance.



The term $\mu/(\mu + 1)$ can be called the intrinsic gain of the follower, and represents the maximum gain which can be obtained and is obviously slightly less than unity, and the quantity $r_a/(\mu + 1)$ is the intrinsic output impedance and represents the maximum value of the output impedance, the output impedance being this quantity in parallel with the cathode impedance Z_k.

2.2. *Equivalent Circuit with Grid–Cathode Impedance*

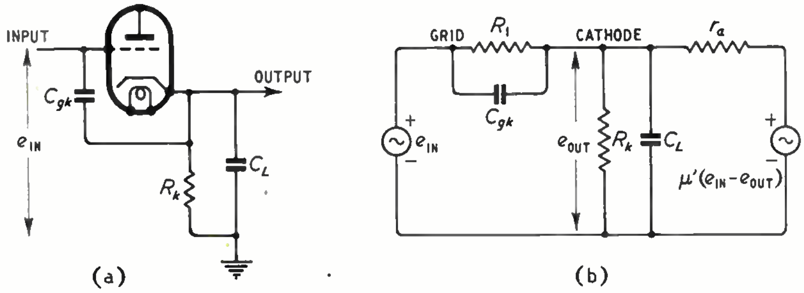
When the effect of the grid-cathode impedance cannot be neglected, the equivalent circuit is shown in Fig. 2(b). In this case the total capacitance between grid and cathode includes any capacitance added externally to the valve capacitance. Furthermore when the grid is driven positive with respect to the cathode the grid current which flows can be represented by a resistance R_1 between grid and cathode which is of the order of one to five thousand ohms.

The effect of the flow of grid current on the valve characteristics can, to a first approximation, be represented as shown in Fig. 2(c), where it is shown that on entering the positive grid region, the valve characteristics show a change in slope, the spacing between the curves of constant anode voltage remaining the same (i.e., the anode resistance is unchanged), but the mutual conductance g_m is reduced to $g_m - 1/R_1 = g_m'$, i.e., reduced by the conductance of the grid circuit, and the amplification factor is reduced accordingly to $\mu - r_a/R_1 = \mu'$.

Thus in the equivalent circuit of Fig. 2(b) the modified value of amplification factor has to be used when in the positive grid region. This is treated in Section 5. These circuits can be used in circuit analysis under both steady-state and transient conditions.

$$R_0 = \frac{R_K\, r_a}{(\mu + 1) R_K + r_a}.$$

(This was developed as Equation (3) in Reference 1.)



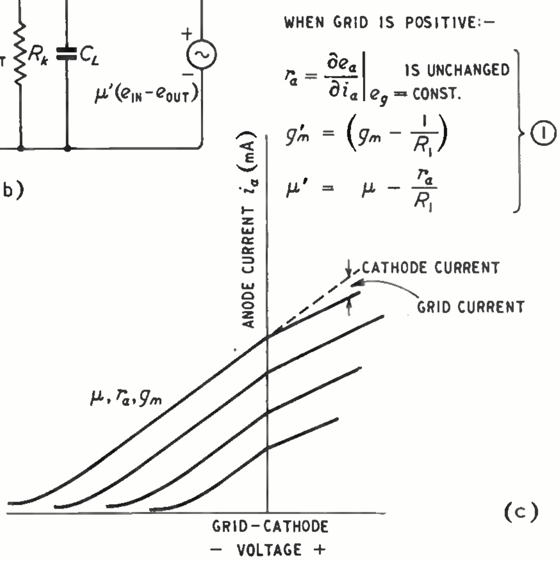


Fig. 2. *Circuit (a) and equivalent circuit (b) considering effects of grid-cathode capacitance and flow of grid current. Valve characteristics are shown in (c).*

2.3. *Use of Superposition Integral*[2,3]

In this paper when considering transient response the response to an input step function only will be considered but attention is drawn to the use of the Superposition Integral, with which the performance under any other input function can be determined from the step-function performance.

If we assume that the response $F(t)$ to a unit-step function is known, then the response $e_o(t)$ to an input function of voltage $e_{in}(t)$ is given by:

$$e_o(t) = e_{in}(0) . F(t) + \int_0^t \frac{d}{d\lambda} \cdot e_{in}(\lambda) . F(t-\lambda) . d\lambda \quad .. \quad (2)$$

or an alternative form is:

$$e_o(t) = e_{in}(t) . F(0) + \int_0^t e_{in}(\lambda) . \frac{d}{d\lambda} . F(t-\lambda) . d\lambda \quad .. \quad .. \quad .. \quad (3)$$

These are known as the superposition integral theorems. Here $e_{in}(0)$ is the value of $e_{in}(t)$ when $t = 0$; $F(0)$ is the value of $F(t)$ at $t = 0$ and the independent variable 't' has been designated by λ, and is used as the variable of integration to determine the response at any time 't'. Thus

$$t = \int_0^t d\lambda.$$

As an illustrative example, it is known that the response of a cathode-follower to a unit-step function is given by:

$$e_{out} = A[1 - \exp(-t/T_0)] = F(t) \quad .. \quad (4)$$

where $A = \text{Gain} = \dfrac{\mu R_K}{(\mu+1)R_K + r_a}$

$T_0 = \text{Time constant} = R_0 C_L.$

Making use of formula (2) above, the response to a constant-change function $E = kt = e_{in}(t)$ can be determined as follows:

$e_{in}(0) = 0$

$e_{in}(\lambda) = k\lambda$

$d/d\lambda \,.\, e_{in}(\lambda) = k$

$$F(t-\lambda) = A\left[1 - \exp\left(-\frac{t-\lambda}{T_0}\right)\right]$$

$$\text{Thus } e_0(t) = 0 + \int_0^t k.A.\left[1 - \exp\left(-\frac{t-\lambda}{T_0}\right)\right] d\lambda.$$

$$= k.A\left[t - T_0\left\{1 - \exp(-t/T_0)\right\}\right] \quad .. \quad .. \quad .. \quad (5)$$

gives the output voltage.

[This formula was developed as formula (8) in Reference 1.]

Thus if the response to a step function is known, the superposition integrals are a powerful tool for calculating the response to any input function.

3. Simple Series Inductance Compensation

3.1. *Actual and Equivalent Circuits*

It has been shown previously that the performance of cathode-followers with capacitive loads under transient conditions is limited by the fact that the load capacitance has to be charged through the effective output impedance of the cathode-follower. Consequently, the cathode voltage cannot rise rapidly enough compared with the grid voltage with the result that grid current can flow during part of the rise time of the cathode voltage.

One obvious method of limiting this effect is to place an inductance between the cathode and the load capacitance, thus permitting the cathode

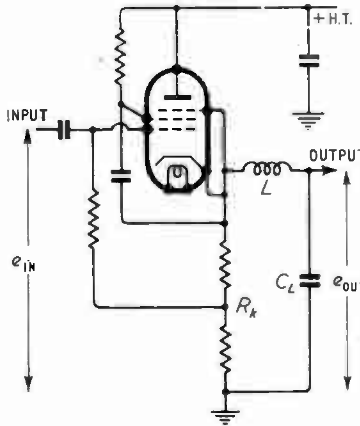
3.2. *Steady-State Performance*

Consider the circuit shown in Fig. 3(b) with a sinusoidal signal applied to the input. The analysis of the steady state relationship between the input and output voltages is given in Appendix 1, in which the output voltage is shown to be:

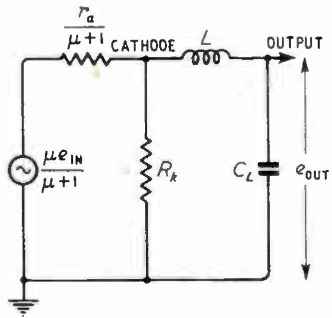
$$|e_{out}| = \frac{Ae_{in}}{\sqrt{1 + k^2(\omega C_L R_0)^4 + (1 - 2k)(\omega C_L R_0)^2}} \quad \dots \quad \dots \quad \dots \quad (6)$$

and the phase shift ϕ between input and output voltages is given by:

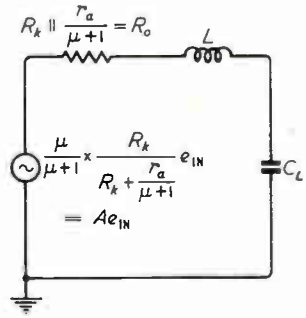
$$\tan\phi = \frac{\omega C_L R_0}{1 - (\omega C_L R_0)^2 k} \quad \dots \quad \dots \quad \dots \quad (7)$$



(a)



(b)



(c)

Fig. 3. *Cathode follower with series inductance (a) and equivalent circuits (b) and (c).*

voltage to differ from the output voltage. The circuit is shown in Fig. 3(a) and the simple equivalent circuit in Fig. 3(b). Both the transient and steady-state performances of this circuit will be considered. It should be noted that this form of compensation is analogous to the shunt-peaking circuit which is used to improve the gain-bandwidth product in wideband amplifiers or the gain rise time ratio of pulse amplifiers and it will be shown that similar advantages can be obtained with a cathode-follower with this form of compensation.

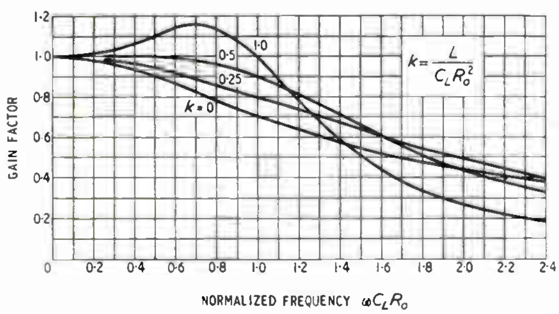


Fig. 4. *Relative gain of a cathode follower with a series inductance between cathode and load capacitance (circuit of Fig. 3).*

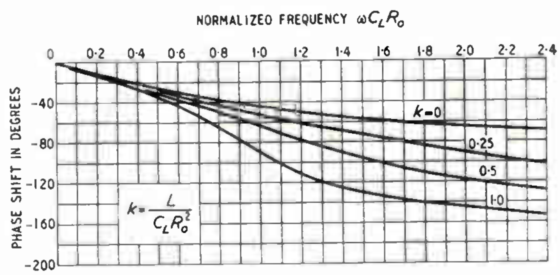


Fig. 5. *Phase shift between input and output voltages for cathode follower with a simple series inductance between cathode and load capacitance (circuit of Fig. 3).*

where:

A = Gain of cathode-follower

$$= \frac{\mu R_K}{(\mu + 1) R_K + r_a} = g_m \cdot R_0$$

k = $L/C_L R_0^2$ is the inductance parameter.

ω = angular frequency

C_L = Load capacitance

R_0 = Output impedance* = $R_K \| r_a \| 1/g_m$.

The relationship between the modulus of output voltage and the normalized frequency $\omega C_L R_0$ is graphed in Fig. 4, for increasing inductance values

* The symbol $\|$ is used to indicate 'in parallel with.'

corresponding to $k = 0{\cdot}25$; $k = 0{\cdot}5$; $k = 1$ and also for the case of no inductance ($k = 0$).

It can be seen that for $L = C_L R_0^2 (k = 1)$, there is an increase in gain up to 1·25 db, followed

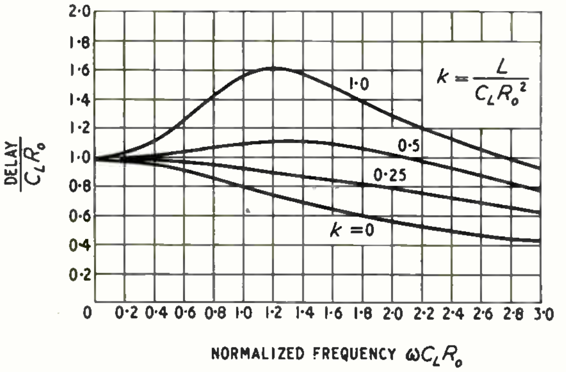


Fig. 6. *Delay characteristics of cathode follower with a simple series inductance between cathode and load capacitance (circuit of Fig. 3).*

by a fairly rapid fall off after $\omega C_L R_0 = 1$ and with $L = 0{\cdot}5 C_L R_0^2$ ($k = 0{\cdot}5$) the gain is fairly constant, the bandwidth to the 3-db down point being increased 1·4 times. The phase characteristics for the above values of inductance are shown in Fig. 5, in which the phase in degrees has been calculated from equation (7) and graphed against normalized frequency $\omega C_L R_0$.

The corresponding delay characteristics are shown in Fig. 6. With $L = 0{\cdot}5\, C_L R_0^2$ or slightly smaller the delay is approximately constant over the bandwidth, thus this value of inductance should produce minimum distortion with complex waveforms. (As expected this is also the optimum value of inductance for best performance under pulse conditions.)

It is obvious from these results that a considerable improvement in performance can be obtained from this simple addition to the circuit.

3.2. *Transient Performance*

The performance of the circuit shown in Fig. 3, when a step-function voltage of amplitude E is

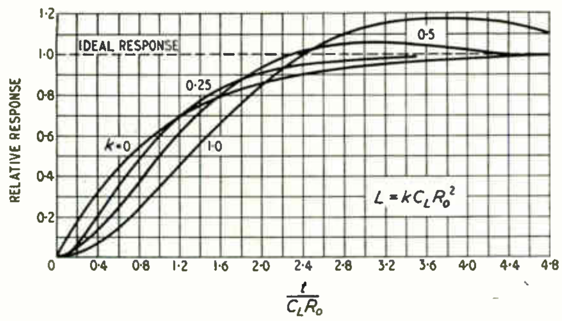


Fig. 7. *Response to a step function of a cathode follower with a series inductance between cathode and capacitive load.*

applied to the input, is determined in Appendix 2 by the use of Laplace transforms. Again the inductance is expressed by $L = k C_L R_0^2$, and it is shown that if k is equal to or greater than 0·25, the output voltage across the capacitive load C_L is given by:

$$e_{out} = A \,.\, E \left\{ \frac{1 - \exp\left(-t/2kC_L R_0\right)}{\sqrt{1 - 1/4k}} \,.\right.$$

$$\left. \sin\left(\frac{\sqrt{1 - 1/4k}}{C_L R_0 \sqrt{k}} \,.\, t + \phi \right) \right\} \quad .. \quad (8)$$

where

$$\tan \phi = \sqrt{4k - 1}$$

and A, C_L and R_0 are as previously defined.

t is time in seconds.

If k is less than 0·25, only exponential terms are present but these cases are not important.

Three particular cases are of interest; viz: those corresponding to $k = 0{\cdot}25$, $k = 0{\cdot}5$ and $k = 1{\cdot}0$.

$$\text{For } k = 0{\cdot}25;\ e_{out} = AE \left[1 - \left\{ \exp\left(- \frac{2t}{C_L R_0} \right) \right\} \left(1 + \frac{2t}{C_L R_0} \right) \right] \quad (9)$$

$$\text{For } k = 0{\cdot}5;\ e_{out} = AE \left[1 - \sqrt{2} \left\{ \exp\left(\frac{-t}{C_L R_0} \right) \right\} \sin\left(t/C_L R_0 + \pi/4 \right) \right] \quad (10)$$

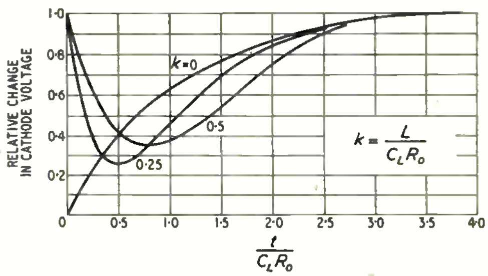


Fig. 8. *Change in cathode voltage for inductively-coupled cathode follower with step-function input.*

$$\text{For } k = 1{\cdot}0;\ e_{out} = AE \left[1 - 1{\cdot}15 \{\exp - (t/2C_L R_0)\} \sin\left(\frac{0{\cdot}866\, t}{C_L R_0} + \frac{\pi}{3} \right) \right] \quad (11)$$

These functions are shown in Fig. 7 in which the voltage across the load capacitance is graphed against the normalized time $t/C_L R_0$. For comparison, the case of $L = 0$ (i.e., the uncompensated case) is shown in addition.

It can be seen that as the inductance is increased, the front delay is increased but the 10-90% rise time is reduced although the time to reach 90%

of the final value is little affected. This is similar to the characteristics of a 4-terminal coupling network.

For large values of inductance there is considerable overshoot which limits the value of inductance which can be used in practice. The optimum value of L is just under $L = 0{\cdot}5\, C_L R_0{}^2$. The various rise times are shown in Table 1.

TABLE 1

Rise time and percentage overshoot for various values of inductance in inductively-compensated cathode-followers.

k	10% Rise time	90% Rise time	10-90% Rise time	% Overshoot
0	$0{\cdot}105\ C_L R_0$	$2{\cdot}30\ C_L R_0$	$2{\cdot}20\ C_L R_0$	0
0·25	$0{\cdot}267\ C_L R_0$	$1{\cdot}95\ C_L R_0$	$1{\cdot}68\ C_L R_0$	0
0·50	$0{\cdot}34\ C_L R_0$	$1{\cdot}87\ C_L R_0$	$1{\cdot}53\ C_L R_0$	5
1·0	$0{\cdot}48\ C_L R_0$	$2{\cdot}12\ C_L R_0$	$1{\cdot}64\ C_L R_0$	16

Again there is obviously considerable advantage in using inductive compensation and a value of $L = 0{\cdot}5\, C_L R_0{}^2$ recommends itself.

3.3. *Cathode Voltage*

So far the output voltage only has been considered but the voltage at the cathode is also of interest, since it determines whether or no grid current will flow during the rise time.

The cathode voltage is given by:

$$E_c = AE - C_L R_0 \cdot \frac{d}{dt}\, e_{out} \quad \ldots \quad \ldots \quad (12)$$

For the general expression of the output voltage, this leads to a complicated expression, but for the particular cases of $k = 0{\cdot}25$ and $k = 0{\cdot}5$ we have:

$$k = 0{\cdot}25;\ E_c = AE \left\{1 - \frac{4t}{C_L R_0} \cdot \exp(-2t/C_L R_0)\right\} \quad \ldots \quad (13)$$

$$k = 0{\cdot}5;\ E_c = AE \left[1 - 2\{\exp(-t/C_L R_0)\}\ \sin t/C_L R_0\right] \quad \ldots \quad (14)$$

These are shown in Fig. 8 in which the cathode voltage is graphed against $t/C_L R_0$. Without inductance this is an exponential rise of time constant $C_L R_0$, but with inductance the voltage rises immediately to its final value, then shows a dip and recovery to the final value. This shows that the possibility of grid current flowing is less, and the grid current is smaller in magnitude and shorter in duration than without inductance.

4. Use of an Ideal Filter Load

4.1. *Circuit*

It is a well-known fact that the input impedance of a low-pass π filter is resistive and has a rising characteristic as the cut-off frequency is approached, and that if shunted with a capacitance, a constant impedance can be maintained over the maximum possible bandwidth associated with the shunting capacitance. Such a circuit is used in wideband amplifiers, where the valve is considered as a constant-current source. In this case the effective supply impedance is of the same order as the load impedance.

Consider the circuit shown in Fig. 9(a) in which the cathode load is a low-pass π filter. The quiescent conditions are not considered here since the correct bias can always be obtained from a separate bias supply circuit if necessary.

The equivalent circuit is shown in Fig. 9(b), and the characteristic impedance of the terminated low-pass filter shown in Fig. 9(c). The characteristic impedance is given by:

$$Z_{0\pi} = \frac{\sqrt{L/C_1}}{\sqrt{1-(\omega/\omega_c)^2}} \quad \text{and} \quad \omega_c = \frac{2}{\sqrt{LC_1}}$$

is the cut off frequency where L and C_1 are as shown in Fig. 9(a).

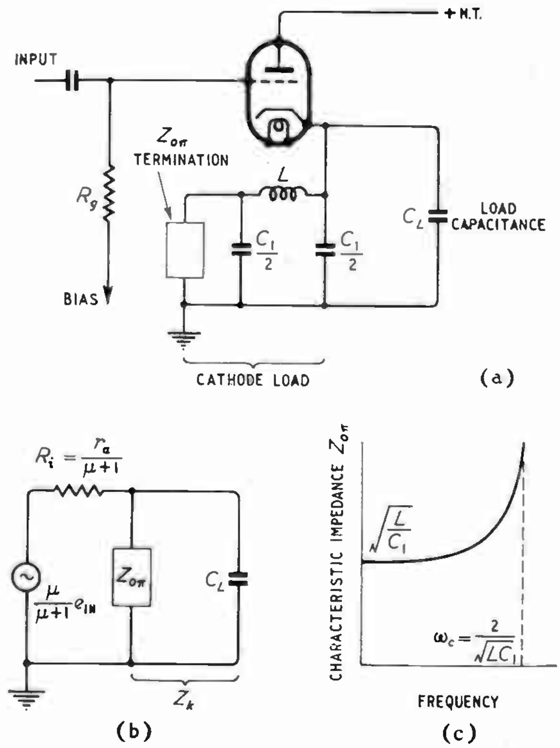


Fig. 9. *Cathode follower with low-pass filter as cathode load (a), its equivalent circuit (b) and characteristic impedance (c).*

When this is shunted by the load capacitance the resultant cathode impedance is given by:

$$Z_k = \frac{\sqrt{L/C_1}}{\sqrt{1-(\omega/\omega_c)^2} + j\omega C_L \sqrt{L/C_1}} \quad \ldots \quad (15)$$

4.2. *Steady-State Response*

Suppose $\sqrt{L/C_1}$, which is the input impedance of the filter at low frequencies and therefore its nominal impedance, is related to the intrinsic output impedance of the cathode-follower $R_i = r_a/(\mu + 1)$ by the factor K_1 thus $\sqrt{L/C_1} = K_1 R_i$ and that the load and filter capacitances are related by the constant K_2 given by: $C_L = K_2 C_1$.

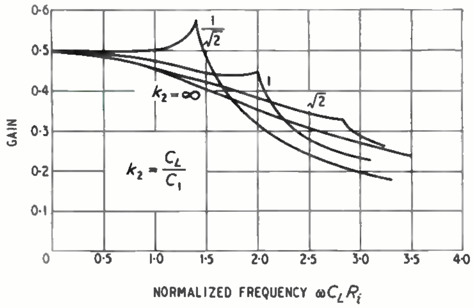


Fig. 10. *Gain of cathode follower with ideal low-pass filter as cathode load.*

Then the cut-off frequency of the filter is given by:

$$\omega c = \frac{2K_2}{K_1 R_i C_L}$$

The cathode impedance given in Equation (15) above is given by:

$$Z_k = \frac{K_1 R_i}{\sqrt{1 - (\omega/\omega c)^2} + j\omega C_L K_1 R_i} \quad \ldots \quad (16)$$

The gain is given by:

$$\text{gain} = \frac{\mu/(\mu + 1)}{\left\{1 + \frac{1}{K_1}\sqrt{1 - \left(\frac{\omega}{\omega c}\right)^2}\right\} + j\left(\frac{\omega}{\omega c}\right) \cdot \frac{2K_2}{K_1}} \quad \ldots \quad \ldots \quad \ldots \quad (17)$$

Best results are obtained with a filter of nominal impedance equal to the impedance of the cathode-follower output (i.e., equal to R_i or $K_1 = 1$).

With a filter of this nominal impedance, the gain at low frequencies will be just below 0·5.

With $K_1 = 1$, we have:

$$\text{gain} = \frac{\mu/(\mu + 1)}{\left\{1 + \sqrt{1 - \left(\frac{\omega}{\omega c}\right)^2}\right\} + j\left(\frac{\omega}{\omega c}\right) \cdot 2K_2} \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

The constant of the filter are:

$$\begin{cases} C_1 = C_L/K_2 \\ L = C_L R_i/K_2 \end{cases}$$

$\omega c = \dfrac{2K_2}{R_i C_L}$. is the cut-off frequency.

Then for frequencies below cut-off the gain is:

$$|\text{Gain}| = \frac{1}{\sqrt{2 + (\omega C_L R_i)^2(1 - 1/4K_2^2) + 2\sqrt{1 - \frac{(\omega R_i C_L)^2}{4K_2^2}}}} \quad \ldots \quad \ldots \quad \ldots \quad (19)$$

and the phase shift is given by:

$$\tan\phi = \frac{\omega R_i C_L}{1 + \sqrt{1 - \frac{(\omega R_i C_L)^2}{4K_2^2}}} \quad \ldots \quad \ldots \quad (20)$$

and for frequencies above cut off:

$$|\text{Gain}| = \frac{2K_2}{\omega R_i C_L \sqrt{(4K_2^2 + 1) + 4K_2\sqrt{1 - \frac{4K_2^2}{(\omega R_i C_L)^2}}}} \quad (21)$$

with a phase shift:

$$\tan\phi = \omega C_L R_i + \sqrt{\frac{(\omega R_i C_L)^2}{4K_2^2} - 1} \quad \ldots \quad (22)$$

Taking $\omega C_L R_i$ as a parameter these quantities are shown in Figs. 10 and 11. Fig. 10 gives the gain as a function of $\omega C_L R_i$ and Fig. 11 gives the phase shift as a function of $\omega C_L R_i$.

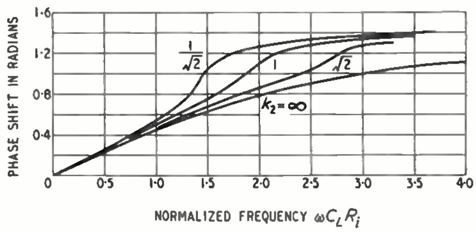


Fig. 11. *Phase shift in cathode follower with ideal low-pass filter load.*

In Fig. 10 the case of a filter with infinite cut-off frequency (i.e., a resistance R_i) is included.

As the cut-off frequency of the filter is increased (corresponding to K_2 decreasing) it can be seen that the gain is held up to the particular cut-off frequency, after which the gain decreases rapidly. With $K_2 = 1/\sqrt{2}$, the gain shows a peak followed by a rapid fall off. It can be seen from Fig. 10 that the curve $K_2 = 1$ shows the most promising improvement.

Looking at the phase-shift curves, it can be seen that the linearity is improved with increasing K_2 from $1/\sqrt{2}$ to $\sqrt{2}$. The phase-shift for a resistive cathode load of value R_i is also included in Fig. 11 for comparison.

4.3. *Transient Response*

The transient response of this circuit is extremely difficult to determine since the inverse Laplace transforms are not standard forms and the solution will not be given here. In general no marked improvement in transient response occurs and, considering the complexity of the circuit and the fact that this load gives a gain of less than 0·5, it is extremely doubtful if it has any practical value.

5. Capacitance Coupling Between Grid and Cathode

So far the effect of the grid-cathode capacitance has been assumed to be small.

Consider the circuit shown in Fig. 2. If a step function is applied to the input, it can be shown that the voltage across the load is given by: (see Appendix 3)

$$e_{out} = A - \left(A - \frac{C_{gk}}{C_{gk} + C_L}\right) \exp\left(-t/R_0(C_L + C_{gk})\right) \quad .. \quad (23)$$

where $A = g_m R_0$ is the gain of the cathode-follower. C_{gk} and C_L are as shown in Fig. 2(a). R_0 = output impedance of follower $= r_a || R_k || 1/g_m$.

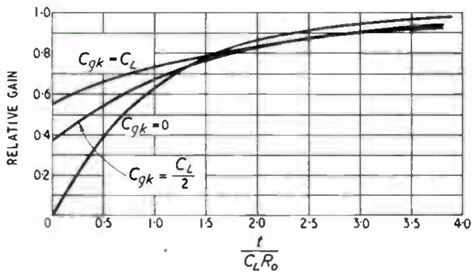


Fig. 12. *Effect of grid-cathode capacitance on response of cathode follower to a step-function input; gain A = 0·9.*

This is graphed in Fig. 12, for several values of C_{gk}. It can be seen that the effect of the grid-to-cathode capacitance is to cause an initial step in the output voltage followed by a subsequent exponential rise in the output voltage, of time constant equal to the output impedance multiplied by the capacitance of the load and grid-cathode capacitance in parallel. This might be useful if an initial step is required in the output.

The 10-90% rise time is obviously not improved. With a step function input voltage there is, of course, an input current given by:

$$I_{in} = e_{in} \cdot \frac{C_{gk}}{C_{gk} + C_L} \left\{ g_m - \frac{C_{gk}}{R_0(C_{gk} + C_L)} \right\} \exp\left(-t/R_0(C_{gk} + C_L)\right) \quad .. \quad (24)$$

This decays exponentially as the cathode voltage rises; thus the high input impedance is lost.

If the grid voltage is ever positive with respect to the cathode an additional current resulting from grid conduction occurs. It should be noted that in this analysis the effect of the flow of grid current has been considered as a resistance R_1 between grid and cathode but the grid current does not enter into the expression for the output voltage, because the flow of grid current is balanced by an equal reduction in the anode current, leaving the total cathode current unaffected.

It can be seen from equation (23) above that if the grid-cathode capacitance is small compared with the load capacitance, which is usually the case if it is not increased externally to the valve, then its effect on the rise time of the output voltage is obtained by considering the grid-cathode capacitance as being in parallel with the load capacitance. It has to increase to a value of the same order as the load capacitance before its effect is noticeable on the output waveform.

Unless the initial step in the output voltage waveform is of particular use increasing the grid–cathode capacitance externally to the valve does not improve the transient response and the flow of input current would make it necessary for the cathode-follower to be driven from a low-impedance source such as another cathode-follower.

6. Conclusion

It can be concluded that the steady-state and transient responses of a cathode-follower with a capacitive load can be improved by the inclusion of an inductance between the cathode and load.

The use of a filter circuit as the cathode load improves both the frequency and the phase-shift characteristics but the gain is less than 0·5 and the circuit is too complex for the improvement obtained.

The capacitance between grid and cathode has little effect on the performance of the cathode-follower if it is small compared with the load capacitance, but if the grid–cathode capacitance is increased externally to the valve an initial step can be obtained in the output-voltage waveform but this is accompanied by a current in the input circuit.

APPENDIX 1

Steady-State Performance of Circuit with Simple Series Inductance

Consider the circuit shown in Fig. 3, where the various symbols are defined.

We have:

$$e_{out} = \frac{Ae_{in}}{1 - \omega^2 LC_L + j\omega C_L R_o}$$

Expressing the inductance L by the parameter k, defined by

$$L = kC_L R_o^2$$

then

$$\frac{e_{out}}{Ae_{in}} = \frac{1}{1 - (\omega C_L R_o)^2 k + j(\omega C_L R_o)}$$

Thus giving:

$$\frac{e_{out}}{Ae_{in}} = \frac{1}{\sqrt{1 + k^2(\omega C_L R_o)^4 + (1 - 2k)(\omega C_L R_o)^2}}$$

as the relative gain, and

$$\tan \phi = \frac{\omega C_L R_o}{1 - (\omega C_L R_o)^2 k.}$$

where ϕ is the phase shift in degrees

$$\text{The delay} = \frac{\text{Phase shift in radians at frequency } \omega}{\omega}$$

$$= \frac{\phi}{180} \cdot \frac{\pi}{\omega}$$

APPENDIX 2

Transient Response of a Cathode-Follower with a Series Inductance Between Cathode and Capacitive Load.

Consider the circuit and equivalent circuit shown in Fig. 3. We have:

$$e_{out} + C_L R_o . \frac{d}{dt} . e_{out} + LC \frac{d^2}{dt^2} e_{out} = A.E$$

Where E is the value of the step-function input voltage. Taking Laplace Transforms of both sides this becomes:

$$\mathcal{L} \frac{e_{out}}{AE} = \frac{1}{LC_L(p^2 + pR_o/L + 1/LC_L)}$$

Taking inverse transforms. (Transform No. 51, Pipes[4], p. 134) when

$$L = C_L R_o^2/4$$

$$\frac{e_{out}}{AE} = 1 - (1 + tR_o/2L) . \exp(-tR_o/2L)$$

or if $L > C_L R_o^2/4$

$$\frac{e_{out}}{AE} = 1 - \frac{1}{\omega\sqrt{L.C_L}} \exp(-tR_o/2L.) \sin(\omega t + \phi)$$

$$\text{where} \quad \omega = \sqrt{\frac{1}{LC_L} - \left(\frac{R_o}{2L}\right)^2}$$

$$\text{and} \quad \tan \phi = \frac{\sqrt{1/LC_L - (R_o/2L)^2}}{(R_o/2L)}$$

$$\frac{e_{out}}{AE} = 1 - \frac{\exp(-R_o t)/2L}{\sqrt{1 - C_L R_o^2/4L}} . \sin\left(t\sqrt{\frac{1}{LC_L} - \left(\frac{R_o}{2L}\right)^2} + \phi\right)$$

$$\tan \phi = \sqrt{\frac{4L}{C_L R_o^2} - 1}$$

Putting $L = kC_L R_o^2$

$$\frac{e_{out}}{AE} = 1 - \frac{\exp(-t/2kC_L R_o)}{\sqrt{1 - 1/4k}} \sin\left(\frac{t\sqrt{1 - 1/4k}}{\sqrt{k}C_L R_o} + \phi\right)$$

and the phase shift is given by:

$$\tan \phi = \sqrt{4k - 1}$$

APPENDIX 3

Determination of the effect of Capacitance between Grid and Cathode.

Consider the typical circuit shown in Fig. 2(a) and the equivalent circuit shown in Fig. 2(b).

Suppose a step function of amplitude E is applied to the input, then the current in the input circuit is given by:

$$I_1 = (E - e_{out})(1/R_1 + pC_{gk})$$

and the current in the valve circuit is:

$$I_2 = \frac{(\mu - r_a/R_1)(E - e_{out}) - e_{out}}{r_a}$$

and $I_1 + I_2 = e_{out}(1/R_k + pC_L)$

Adding the first two equations and equating to the third equation gives the transform equation:

$$e_{out} = E\left\{\frac{\mu/r_a + pC_{gk}}{\mu/r_a + 1/r_a + 1/R_k + p(C_L + C_{gk})}\right\}$$

Taking inverse transforms we have:

$$\frac{e_{out}}{E} = A\left[1 - \exp\{-t/R_o(C_L + C_{gk})\} + \frac{R_o C_{gk}}{R_o(C_{gk} + C_L)} . \exp\{-t/R_o(C_L + C_{gk})\}\right]$$

(Transforms Nos. 7 and 8 p. 130, Reference 4)

$$\therefore \frac{e_{out}}{E} = A - \left(A - \frac{C_{gk}}{C_{gk} + C_L}\right) . \exp\{-t/R_o(C_{gk} + C_L)\}$$

Where $R_o = R_k \| r_a \| 1/g_m$.

$A = g_m R_o$ = gain of cathode follower.

The input current I_1 is given by:

$$I_1 = (E - e_{out})(1/R_1 + pC_{gk}).$$

If the grid is never positive, so that R_1 is infinitely large, the input current is:

$$I_1 = \frac{C_{gk}}{C_{gk} + C_L} \{g_m - C_{gk}/R_o(C_{gk} + C_L)\} \exp\{-t/R_o(C_{gk} + C_L)\}$$

REFERENCES

[1] Shimmins, A. J. "Cathode-Follower Performance, Pulse and Saw-tooth Conditions," *Wireless Engineer*, Dec. 1950.

[2] Vannevar Bush. "Operational Circuit Analysis," John Wiley & Sons, New York, 1929, Ch. 6.

[3] Arguimbau, L. B. "Vacuum Tube Circuits," John Wiley & Sons, New York, 1948, pp. 153 and 590.

[4] Pipes, L. A. "Applied Mathematics for Engineers and Physicists," McGraw-Hill Book Company, 1946, pp. 128–137.

INTERFERENCE IN TELEVISION PICTURES

Effect of Line-Deflection Circuits

By G. Diemer, Z. van Gelder and J. J. P. Valeton

(*Philips Research Laboratories, N.V. Philips' Gloeilampenfabrieken, Eindhoven, Netherlands*)

SUMMARY.—Interference, which takes the form of vertical lines on the left-hand side of a television picture, has been studied. It can be ascribed to intrinsic properties of the I_a-V_a characteristics of the power valve used for generating the line-deflection current. These properties cause irregularities in the anode current and give rise to signals of very-high frequency, which may penetrate into the r.f. or i.f. amplifier of the receiver.

1. Introduction

A VERY inconvenient form of interference, which sometimes occurs in very sensitive television receivers, is the appearance of distinct vertical lines at the left-hand side of the picture. It may be useful to stress that only high-frequency interference is meant here and not the well-known vertical bars, which also occur on the left-hand side of a television picture and are due to damped oscillations in the secondary windings of the line-scanning transformer. These oscillations, modulating the spot velocity, cause white bars at moments when the velocity is low. The form of interference that is dealt with in this article becomes visible by modulating the beam current of the cathode-ray tube.

In the American literature[1] this phenomenon is known as "Barkhausen effect." Generally it is supposed that, at the moment the anode of the horizontal-deflection valve goes negative with respect to the screen grid, Barkhausen oscillations will be set up by groups of electrons which travel to and fro between the anode and the control grid. Such oscillations would induce a voltage into sensitive parts of the receiver, giving rise to interference in the picture. As a remedy the use of a permanent magnet is suggested, affixed to the glass bulb of the valve.

Indeed, at low-anode voltages the electrons will circle round the wires of the screen grid. A closer examination of the distances between the electrodes and of the electrode potentials of normal tetrodes shows at once that the period of electron movement is so short that the lowest resonance-frequency will be found in the region of decimeter waves. Therefore, it is difficult to explain the interference on meter waves by this phenomenon. Hence we were inclined to reject the Barkhausen hypothesis.

In order to study this phenomenon more quantitatively we have made various experiments with the PL81, a beam power valve for horizontal-deflection circuits. Using this valve in a sensitive receiver with a particular line-sweep circuit, vertical black lines appeared in the picture, the receiver being designed for signals with negative modulation. With positive modulation white lines would appear in the picture. In this paper it will be shown that some circuit-elements, as well as the properties of the beam power valve, were responsible for the occurrence of this type of interference in the television picture.

MS accepted by the Editor, October 1951

The important factors are the leakage inductance of the transformer and the properties of the beam power valve, which differ considerably from those of an ideal switch. The secondary emission of the anode of this valve plays an important part, too.

2. Barkhausen Oscillations

Barkhausen[2] showed that, in a triode with positive grid and a low or negative anode-voltage, oscillations of very high frequency can arise if, between grid and anode, a parallel tuned circuit is present, having a resonance frequency equal to that of the electrons.

If we ignore the tangential velocities of the electrons, the frequency of the oscillations is given by

$$f = \frac{1}{\tau} \approx 10^7 \frac{\sqrt{V_g}}{d_1 + d_2} \text{ cycles per second,}$$

where d_1 is the cathode-grid distance, d_2 the anode-grid distance (in centimetres) and V_g the average potential in the grid plane.

If for the distance d_1 is taken the distance between the control and screen grids of the PL81, and for d_2 the distance from the screen grid to the reversing point of the electrons, d_1 and d_2 are 0·55 mm and about 1·0 mm respectively. With $V_g = V_{g2} = 180$ volts, the frequency of the oscillations is of the order of 1000 Mc/s. The presence of space-charges tends to decrease this frequency[3], but in any case the Barkhausen frequency is several times higher than the interference frequency observed, which lies in the order of metre waves ($f \leqslant 200$ Mc/s). These Barkhausen frequencies vary, however, also with changing potentials of the control-grid because at higher cathode currents the reversing point of the electrons near the anode varies, so d_2 changes.

If the anode voltage could be negative the transit time of the electrons would decrease and the frequency would become still greater.

It is obvious that these Barkhausen oscillations can only occur if the potential of the anode has such a low value that a part of the electrons travelling towards the anode must return to the screen-grid, because of energy loss in the direction of the anode due to deflections of the electron paths in the preceding grids.

This is only the case when the anode has a potential which is below the knee of the I_a–V_a characteristic.

Taking all this into account, it is difficult to explain how these Barkhausen oscillations can give rise to interference in the picture. There are, however, other effects, which will be discussed later, that can give rise to interference causing vertical lines in the picture.

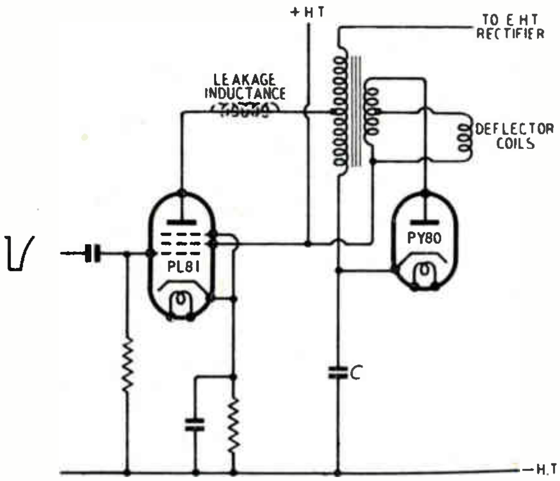


Fig. 1. *Circuit diagram of a line time-base with booster diode.*

3. Experiments

3.1. *Circuit*

In Fig. 1 a diagram of the horizontal-deflection circuit is given. At the end of the sweep the anode current of the power valve has a high value. At this moment t_1 (Fig. 2) the control grid of this valve suddenly goes negative, the anode current is cut off and an oscillation starts in the anode circuit; i.e., in the transformer. With an ideal transformer (no leakage inductance) the oscillations would cease entirely when the booster diode becomes conducting; that is, at the moment t_2 where the anode voltage of the diode tends to exceed the voltage at the booster capacitance C (Fig. 1). The leakage inductance of the actual transformer, however, forms an oscillatory circuit on its own (with a very high Q factor), which is not damped by the booster diode. Therefore the energy in this leakage inductance will cause damped oscillations [period T_1 Fig. 2(a)] with a long decay time.

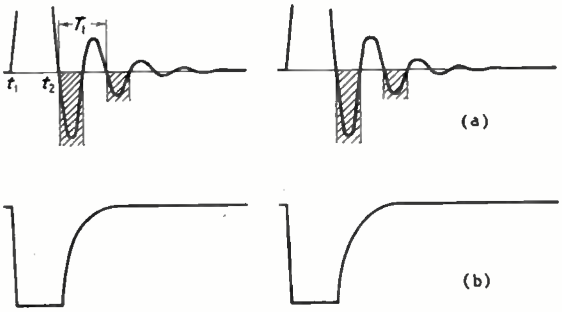


Fig. 2. (*a*) *Anode voltage of the power valve as a function of time, showing damped oscillations;* (*b*) *Shape of the pulse on the control grid of the power valve.*

In the meantime the line-deflection valve is gradually made conducting again. At this moment the anode voltage consists of a low positive voltage on which an alternating voltage component is superimposed due to the oscillations mentioned above. The input voltage of the power valve is given in Fig. 2(b).

The interference (i.e., the black lines) were visible at certain fixed positions in the picture. We found these positions to be *independent* of the moment at which the cathode current of the power valve was started. The lines, however, only showed up if the valve was conducting.

The time the spot needs to travel from one black line to the next proved to correspond to one period [T_1 in Fig. 2(a)] of the resonance frequency of the leakage inductance, which frequency was of the order of 0·5 Mc/s. From the positions of the black lines on the screen it appeared that the interference occurred near the minima in V_a [hatched parts in Fig. 2(a)]. Sometimes, with strong interference, pairs of lines appeared instead of single lines.

Having observed these phenomena we concluded that an explanation could perhaps be given in the following way.

We found the interference to occur at moments when the anode voltage had a certain low value while the anode current was also low. Therefore, we assumed that the interference was caused by irregularities in the flat part of the I_a–V_a characteristic of the power valve at low values both of I_a and V_a, just above the knee in this characteristic.

This was confirmed by a test with the booster diode short-circuited and the control grid and anode of the power valve at various direct voltages. It appeared that, with those voltages of anode and control grid at which the internal resistance r_a of the valve is negative, the circuit started to oscillate continuously. These oscillations showed considerable distortion due to the non-linearity of the I_a–V_a characteristic and the frequency spectrum extended into the range of metre waves. The oscillations ceased when the valve was shunted with a resistance of such a value that the total resistance became positive. Moreover, in the region of negative r_a, the tetrode characteristic at low anode currents shows maxima and minima caused by reflected electrons[4] from the anode (Fig. 3), which also contribute to the distortion of the oscillations.

These high-order harmonics can be picked up in the early stages, also, of other nearby receivers, not necessarily television receivers, and cause interference in the pictures.

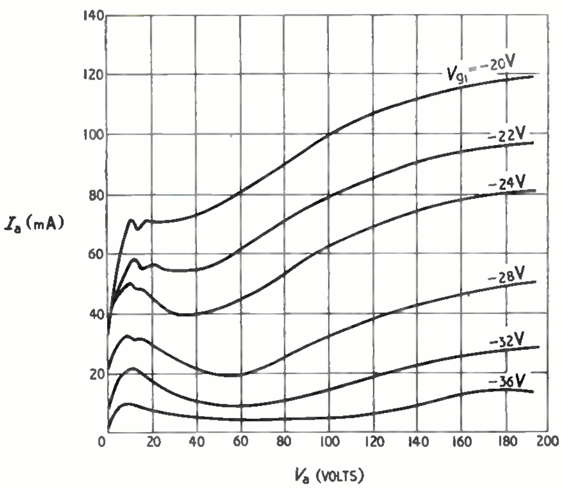


Fig. 3. *I_a–V_a characteristics of a normal beam power valve. With low cathode currents the secondary electrons are able to reach the screen grid.*

3.2. *Means of Suppressing Secondary Emission*

Our first attempt to eliminate the black lines consisted in trying to suppress secondary emission. An improvement was certainly obtained when using a valve with an anode of a special shape, which considerably reduced the secondary emission (Fig. 4). The intensity of the black lines was thereby reduced but they were still visible.

The same test was made with a pentode having a characteristic similar to that of the PL81, but showing of course less secondary emission. With the suppressor grid at cathode potential black lines were, however, still visible. After this the suppressor grid was given a negative voltage, resulting in the interference being so reduced in strength that the black lines became visible only in the case of extreme sensitivity of the receiver and with strong coupling between aerial and line time-base. It should be remarked, however, that now the anode current at low-anode voltages was reduced so much that full deflection could not be obtained.

The next step was to try out a special valve showing nowhere a negative slope in its I_a–V_a characteristic. This valve consisted of a pentode and a tetrode, the two systems in one bulb. These two electrode systems together gave an I_a–V_a characteristic with positive r_a over the whole region (Fig. 5). The construction of the valve was as follows:—

A small part of the PL81 system was changed into the desired pentode by introducing a fine suppressor grid consisting of gauze in the plane of the beam plates, and by giving the grids locally a larger pitch in order to shift the knee of the pentode characteristic by space charge. In this way a sufficient positive internal resistance of the pentode could be obtained in the region where the tetrode has a negative internal resistance.

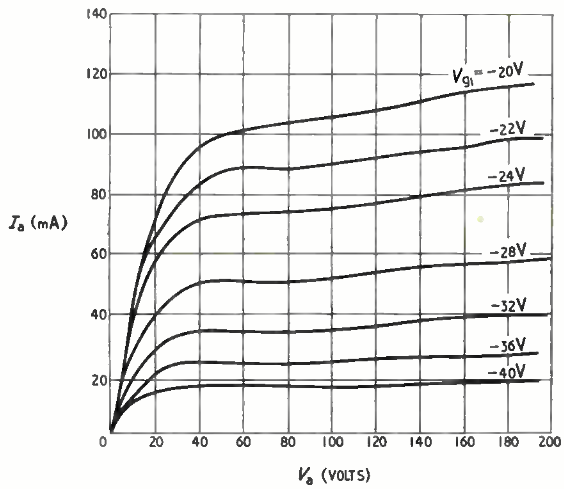


Fig. 4. *I_a–V_a characteristics of a special beam power valve at low anode currents. The anode of the valve has a special shape to suppress secondary emission.*

Notwithstanding the fact that this valve did not show a negative slope anywhere in its I_a–V_a characteristic (Fig. 6), it was still possible to produce the interference in the picture with a circuit like Fig. 1 if the transformer had an appreciable leakage inductance.

3.3. *Characteristics of Beam Power Valves*

The remaining interference was, therefore, presumably to be ascribed to the typical characteristic of the beam power valve. If the anode voltage is modulated in such a way that it passes the knee in the I_a–V_a characteristic, the anode current will contain harmonics of the modulation frequency. Especially when the anode voltage reaches negative values, causing rather abrupt cut-off, the current will contain very high-order harmonics. Such a modulation of the anode voltage can arise from inadequately damped oscillations of the transformer leakage-inductance.

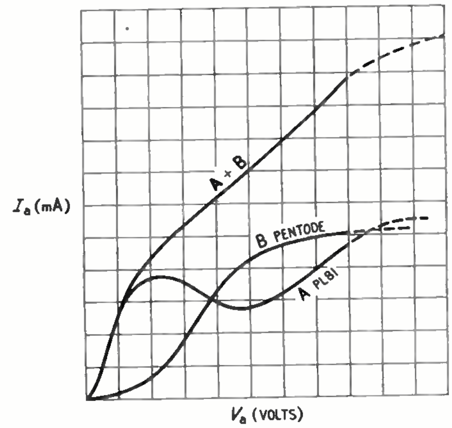


Fig. 5. *I_a–V_a characteristic of a valve having nowhere a negative slope (curve A + B) obtained by a combination of the characteristic of a tetrode (curve A) and a pentode with a poor current-division characteristic (curve B).*

The interruptions of the anode current occur with very steep flanks [see Fig. 7(b), time interval τ_1] giving rise to harmonics of such a high order that they are able to penetrate the receiver via its i.f. part. This follows immediately from an estimation of the steepness of the flanks. With large values of the alternating anode voltage, τ_1 can easily become 1% of the period T_1 of the leakage inductance, T_1 being of the order of 1 μsec. Therefore, the frequency spectrum of the pulses will contain frequencies of the order of 50 Mc/s.

From the above it will be clear, that *this sudden "switching" of the anode-current can give rise to pulse-like interference, penetrating the receiver via the h.f. or i.f. part.* Being synchronized with the line frequency it appears in the picture in orderly fashion in the form of vertical lines.

A similar effect of current switching occurs at the moment when the booster diode becomes conducting. This switching on of the booster-diode current also occurs within a very short time, resulting in high-frequency interference that is visible as a vertical line at the left-hand edge of the raster. This line, however, is normally suppressed by the blanking signal.

4. Quantities affecting the Intensity of the Interference

From the experiments described above it can be concluded that the most important factor influencing the intensity of the interference is the leakage inductance of the transformer. The oscillatory circuit formed by this inductance and the stray capacitance is not damped by the booster diode nor by the power valve until the latter becomes conducting. Thus, with a considerable leakage inductance, the oscillations start with a large initial voltage-amplitude. The possibility that the anode of the power valve attains periodically a voltage below zero thus depends greatly on the properties of the output transformer.

When employing a good transformer with a very small leakage inductance and a normal PL81 the interferences were indeed no longer perceptible.

5. Further Considerations of Previous Results

The results of the first test with modified valves described in Section 3.2. can be accounted for by our hypothesis. Owing to the typical properties of these valves, which have a relatively high internal resistance at very low values of anode voltage (i.e., below the knee), the current-division characteristic is more or less impaired, but only at low anode currents (see Figs. 3-5). These modifications result in the interruptions of the anode current being less abrupt. The sharp variations in the anode current as a function of time (Fig. 7) are smoothed out, so that the higher frequencies in the spectrum are attenuated. This explanation holds, too, for the case where the characteristic is altered by the presence of a magnetic field.

Another effect which can be explained by our hypothesis is the occurrence of *pairs* of lines. This points to a second burst of interference during the negative phase of the anode alternating voltage.

Experimentally, it was shown that by

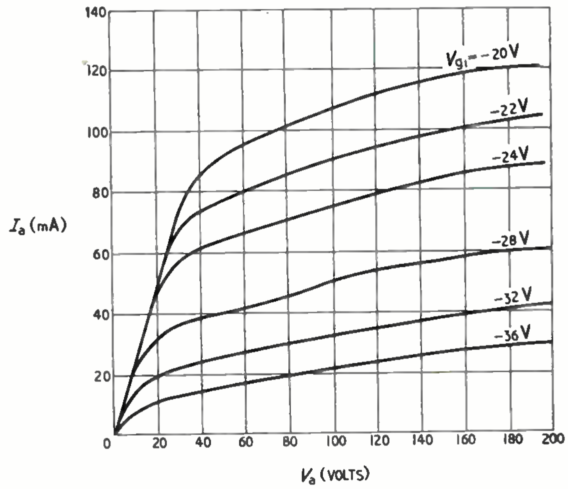


Fig. 6. *I_a–V_a characteristics of a tetrode-pentode combination. At low cathode currents the slope of the characteristic is positive everywhere.*

continually changing the input voltage on the control grid of the power valve it was possible to make these lines approach each other and eventually merge together. This effect may be explained by assuming that the amplitude of the oscillations was made smaller and smaller, thus causing the duration of the negative excursion of the anode voltage to be shortened. At a certain moment the two lines can no longer be seen separately and merge.

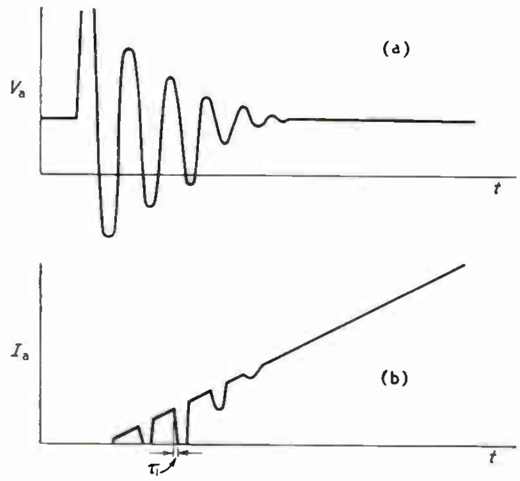


Fig. 7. (*a*) *Anode voltage of the power valve as a function of time;* (*b*) *Anode current of the power valve as a function of time. Every time the anode voltage becomes negative the anode current is interrupted.*

6. Instabilities in the I_a–V_a Characteristics of Beam Power Valves

Another property of the I_a–V_a characteristic of beam-power valves is the instability of this characteristic at low anode potentials and high anode currents. Due to the virtual cathode, which may be formed under these conditions, the anode current has two stable positions at low anode potentials. If the anode of a power valve, as used in a line-sweep circuit was modulated with an alternating voltage it was found that the anode current jumped from one stable position to the other. This occurred within a time that is of the same order as the transit time of the electrons and with a repetition frequency equal to the modulating frequency. These jumps could be made visible on a wideband cathode-ray oscilloscope. With a sufficiently tight coupling between the aerial of the receiver and the beam-power valve this type of interference indeed appeared in the television picture in complete correlation with the presence of jumps in the anode current. They were, however, weaker than the vertical lines mentioned first. The frequency spectrum was the same as the switching phenomenon mentioned above. Owing to the conditions given above the interference caused by this effect is visible as lines situated more to the centre and to the right side of the picture.

7. Conclusion

The interference in the form of vertical lines in the picture of television receivers was investigated. Beside Barkhausen oscillations there are other effects which cause vertical bars in the television pictures.

In this article it has been shown that the typical properties of the power-valve characteristics and the transformer used in the line-deflection circuit cause vertical bars in the picture.

It has been possible to avoid this interference, partly by adopting a special valve-construction whereby the secondary emission of the anode is suppressed. An important factor, however, proved to be the leakage inductance of the transformer used. If this inductance is large the anode voltage becomes periodically negative during the sweep and even with an ideal I_a–V_a characteristic the interference is still apt to occur.

Acknowledgments

Our thanks are due to Mr. P. J. H. Janssen of the Philips television laboratory and Mr. A. Ciuciura of the Mullard valve measurements and application laboratories for valuable criticism.

REFERENCES

[1] See, e.g., D. Lerner and J. Howell, *Radio and Television News*, Vol. 44. No. 3, pp. 50-51. 1950.

[2] H. Barkhausen, *Phys. Zeitschrift*, Vol. 21, pp. 1-6, 1920.

[3] K. S. Knol and G. Diemer. "High-frequency Diode Admittance with Retarding d.c. Field," *Philips Res. Rep.* (to appear in the near future).

[4] J. L. H. Jonker, *Philips Res. Rep.*, Vol. 2, pp. 331-339, 1947.

BRITISH STANDARDS

Cotton-Covered Round Copper Wires

B.S.1791:1951. Covers single and double cotton-covered wire of 0·006 to 0·192 in, diameter, the covering being helically wound, not braided. Price 2s.

Enamel (oleo-resinous) and Cotton-Covered Round Wires

B.S.1815:1952. Covers combined enamel and cotton-covered wire of 0·006 to 0·16 in. diameter; single and double cotton coverings are dealt with, each in both ordinary and fine types. Price 2s.

Nickel Silver Strip and Foil for the Telecommunication Industries

B.S.1824:1952. Covers chemical compositions, temper grades, mechanical properties and tolerances on dimensions. Also includes information on test certificates and independent tests. Price 2s.

Transformers for Cinematograph Equipment

B.S.1793:1952. Price 2s.

Enamel Round Copper Wire [Synthetic Resin (Vinyl Acetate Base) Enamel]

B.S.1844:1952. Price 4s.

Copies of British Standards are obtainable from the British Standards Institution, Sales Branch, 24 Victoria Street, London, S.W.1.

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.

Cathode-Coupled Amplifier

SIR,—In your March 1952 issue, Mr. J. A. Lyddiard gives an interesting method of small-signal analysis of the very versatile cathode-coupled amplifier circuit. However, for some applications his equivalent circuits are not sufficiently general, although it may be possible to extend his method to obtain more general results. An alternative and more general equivalent circuit which I have found useful is shown in Fig. 1. This enables the magnitudes of the currents in both anodes and in the cathode resistor to be calculated when either one or two signals are applied. The output impedance at either anode or the cathode can be found by calculating the impedance seen from terminals X, Y, or Z; the impedance obtained in this way is multiplied by the same factor as the resistor joined to the terminals; e.g., the impedance seen from terminals Y, Y, is multiplied by $(\mu_1 + 1)$.

I am afraid that Mr. Lyddiard's treatment of harmonic distortion may be misleading. In particular, his statement that the harmonic distortion of the cathode-follower stage is usually negligible in practice, while probably quite justified in the case which he considers in detail, is not true in all practical cases. The cathode-follower, whose voltage ratio lies in the region of 0·5 to 0·7, can in fact produce distortion in amounts similar to, or even greater than, the earthed-grid stage. The even harmonics from the two stages are in opposite phase and some cancellation always takes place. The use of a high value of R_K, made possible by application of a positive bias to both grids, is advantageous in obtaining cancellation. A slight adjustment of the bias on one of the grids will usually permit almost complete cancellation of the even harmonics. In addition to improving linearity, the application of positive bias has other advantages: the anode currents of V_1 and V_2 are made more nearly equal, thus avoiding the danger of exceeding the anode dissipation of V_1 mentioned by Mr. Lyddiard, the increased current in V_2 reduces r_{a2} which, together with the increased value of R_K, leads to a higher stage gain (see Fig. 1), and the greater d.c. feedback from the increased value of R_K stabilizes the operating conditions of the valves, giving greater stability of gain.

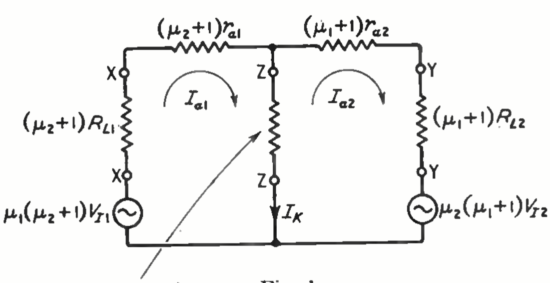


Fig. 1.

An alternative method of calculating harmonic distortion, which is convenient and accurate, makes use of the input/output transfer characteristic of the stage. This can easily be calculated from the published valve curves as follows: the dynamic characteristic of V_2 with its anode load is first found in the normal way; from this the values of V_{g2} are found for a number of values of I_{a2} and the following quantities calculated in turn for each value of I_{a2}; V_k, $(I_{a1} + I_{a2})$, I_{a1}, V_{g1} (from the appropriate valve curve), and then V_{I1}. This permits the transfer characteristic (I_{a2}/V_{I1}) to be plotted. The bias for V_1, if not already fixed, can now be chosen for minimum distortion and the harmonic distortion can then be calculated by any of the normal methods.

I. F. MACDIARMID.

Dollis Hill,
London, N.W.2.
9th April, 1952.

Image Impedances of Active Linear Four-Terminal Networks

SIR,—The image impedance at the input and output terminals of a passive four-terminal network is a well-known and widely-used concept.[1] It is conveniently calculated or measured by using the formula:

$$Z_{I1}^2 = z_{11}/y_{11} \quad \ldots \quad (1)$$

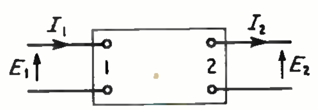


Fig. 1.

where: Z_{I1} is the image impedance at terminals 1 (see Fig. 1)
z_{11} is the input impedance at terminals 1 with terminals 2 open circuited.
y_{11} is the input admittance at terminals 1 with terminals 2 short circuited.

Equation (1) still applies when the network is active, provided, of course, that the active elements are linear. This can be shown as follows, a knowledge of the application of simple matrix techniques[2] to four-terminal networks being assumed;

$$\text{In Fig. 1,} \quad \begin{bmatrix} E_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} E_2 \\ I_2 \end{bmatrix} \quad \ldots \quad (2)$$

If the network is entirely passive, the restriction $AD - BC = 1$ is imposed on the matrix elements. No such restriction will be applied here.[3]

Let $AD - BC = \eta$

$$\text{Then} \quad \begin{bmatrix} E_2 \\ I_2 \end{bmatrix} = \begin{bmatrix} \dfrac{D}{\eta} & -\dfrac{B}{\eta} \\ -\dfrac{C}{\eta} & \dfrac{A}{\eta} \end{bmatrix} \begin{bmatrix} E_1 \\ I_1 \end{bmatrix} \quad \ldots \quad (3)$$

Now let the network be terminated at terminals 1 and 2 by external impedance Z_1 and Z_2 and let the impedances measured looking in to the corresponding terminals of the network under these conditions be Z' and Z''. Then from (2)

$$Z' = \frac{AZ_2 + B}{CZ_2 + D} \quad \ldots \quad (4)$$

from (3)

$$Z'' = \frac{DZ_1 + B}{CZ_1 + A} \quad \ldots \quad (5)$$

from (4)

$$z_{11} = \frac{A}{C},\ \frac{1}{y_{11}} = \frac{B}{D},\ \frac{z_{11}}{y_{11}} = \frac{AB}{CD} \quad \ldots \quad (6)$$

Now let the external impedances be the image impedances; i.e., $Z_1 = Z' = Z_{I1}$, $Z_2 = Z'' = Z_{I2}$

Then substituting in (4) and (5) and eliminating Z_{I2} gives: $Z_{I1}{}^2 = \frac{AB}{CD}$ .. .. (7)

From (6) and (7) $Z_{I1}{}^2 = z_{11}/y_{11}$

REFERENCES

[1] Guillemin. "Communication Networks," John Wiley & Sons.

[2] Pipes. "Applied Mathematics for Engineers and Physicists," McGraw-Hill.

[3] Brown and Bennett. "Application of Matrices to Vacuum Tube Circuits," *Proc. Inst. Radio Engrs.* July 1948, Vol. 36, No. 7.

H. SUTCLIFFE.

University of St. Andrews,
University College,
Dundee.

16th April, 1952.

NEW BOOKS

Radio Astronomy

By BERNARD LOVELL, O.B.E., Ph.D., F.Inst.P., F.R.A.S., and J. A. CLEGG, Ph.D., F.Inst.P., F.R.A.S. Pp. 238 with 120 illustrations. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 16s.

Most radio engineers are aware that the application of radio methods has extended in recent years beyond the terrestrial region. They are aware of solar and galactic noise as a source of interference with communications and they remember the radar experiments of 1946 when echoes were received from the moon. It may come as a mild surprise, however, to find that radio methods are now so much used in the extra-terrestrial region that they form a new branch of science—radio astronomy.

Astronomy is defined as "the science of the heavenly bodies." If we except meteors, of which we obtain some direct knowledge, since fragments land on earth, all our information about heavenly bodies comes by means of electromagnetic radiation. Radio astronomy thus differs from ordinary astronomy only in obtaining information by means of radiation in a different part of the spectrum. This is not, however, to say that the difference is merely one of a carrier frequency and that the information conveyed by it is the same. The difference is far more profound. Some of the information gleaned by visual and by radio methods may be the same, it is true, but the mere fact of the existence of radiation at radio-frequency tells us something. There are bodies which emit radiation at radio-frequencies without giving any visible sign of their presence and whose existence was, therefore, unsuspected until radio methods were adopted. These bodies are by no means rare and "radio stars" appear to be as common as ordinary visible ones. The importance of radio to astronomy is thus great.

This book on the subject opens with a chapter on Fundamental Astronomy in which the meanings of common astronomical terms, such as, right ascension, the celestial sphere, etc., are explained. The second chapter deals with The Solar System, Stars and Galaxies and in it the laws of planetary motion are explained. These two chapters give an essential basis for the understanding of the later part of the book and must be read carefully by anyone unfamiliar with astronomical matters.

Chapters 3 and 4 cover The Transmission and Reception of Electromagnetic Waves and Radio Methods of Investigation and are written primarily for those with little or no acquaintance with radio matters. They give the essential background of radio much as Chapters 1 and 2 provide a technical basis for the astronomical part.

The rest of the book deals, in the main, with the results obtained by radio methods and there are seven chapters covering various aspects of comets and meteors, four chapters dealing with solar matters and three with the galaxy. Chapter 19 covers the twinkling of radio stars, which apparently twinkle in much the same way as visible ones! Chapter 20 deals with radio and the aurora borealis and Chapter 21 with the moon. The last chapter has the incomprehensible (to a non-astronomer) title of The Planets and the Gegenschein. It seems that the Gegenschein is a bright patch which appears opposite the sun in a faint band of light which can sometimes be seen in the west after sunset or in the east before sunrise.

Some of the methods described illustrate clever radio technique. From the radio engineer's point of view it is a fault of the book that the equipment and the detail of the methods employed are treated too briefly. None the less, the book is an interesting one and can be recommended to all who want to know something of radio astronomy.

W. T. C.

Industrial Magnetic Testing

By Professor N. F. ASTBURY, M.A., M.I.E.E., F.Inst.P. Pp. 132 with 41 diagrams. The Institute of Physics, 47 Belgrave Square, London, S.W.1. Price 25s.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for April, 1952

Date 1952 April	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1029-1130 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1*	+ 0·2	+ 1	— 19·2
2*	+ 0·1	+ 1	— 20·0
3*	+ 0·1	— 2	— 19·3
4*	0·0	0	— 17·2
5	0·0	0	— 18·5
6	+ 0·1	N.M.	— 19·2
7*	+ 0·2	+ 1	— 16·9
8*	+ 0·2	0	— 18·4
9**	—	+ 1	—
10*	+ 0·2	+ 1	— 17·6
11	+ 0·2	+ 1	— 17·0
12	+ 0·3	+ 1	— 18·2
13	+ 0·2	+ 2	— 18·2
14	N.M.	N.M.	N.M.
15	+ 0·4	+ 1	— 16·2
16*	+ 0·2	+ 1	— 16·6
17	+ 0·3	+ 2	— 16·4
18	+ 0·2	+ 2	— 16·0
19	+ 0·3	+ 1	— 15·7
20	+ 0·3	+ 2	— 15·3
21*	+ 0·3	+ 3	— 15·8
22*	+ 0·3	+ 3	— 14·7
23*	+ 0·4	+ 3	— 15·0
24	+ 0·3	+ 3	— 15·5
25	+ 0·3	+ 3	— 16·2
26**	—	+ 2	—
27	+ 0·2	+ 2	— 19·3
28	+ 0·3	+ 3	— 17·8
29*	+ 0·3	+ 3	— 18·3
30**	—	+ 4	—

The transmitter employed for the MSF 60-kc/s signal is sometimes required for another service.

N.M. = Not measured.

* = No MSF transmission at 1029 G.M.T. Results for 1429-1530 G.M.T.

** = No MSF transmission at 1029 G.M.T. or at 1429 G.M.T.

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 **1487**

References to Contemporary Papers on Acoustics.—R. T. Beyer. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 724–730.) Continuation of 293 of February.

534.121.2.001.362 **1488**

Transformer Analogs of Diaphragms.—B. B. Bauer. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 680–683.) The action of a diaphragm is shown to be analogous to that of a system of ideal transformers, each corresponding to a particular area of the diaphragm. Equivalents for various types of diaphragm are described.

534.152 **1489**

Demonstration of Standing Waves in the Free Acoustic Field, and a Simple Receiver for Short Acoustic Waves.—R. W. Pohl. (*Naturwissenschaften*, Nov. 1951, Vol. 38, No. 21, pp. 486–490.) Two methods of recording a standing-wave pattern are described. A shadow picture is obtained of (*a*) the disturbance of a liquid surface (water or petrol) over which the sound wave is projected, or (*b*) the turbidity in a soap film arranged obliquely in the sound field. Application of the first method in acoustic measurements is illustrated.

534.21-14 **1490**

Normal-Mode Propagation in Three-Layered Liquid Half-Space by Ray Theory.—C. B. Officer, Jr. (*Geophys.*, April 1951, Vol. 16, No. 2, pp. 207–212.) The integral expression for the field due to a point source is derived by summation of multiple reflections of waves from a plane-wave source. Its physical significance is discussed.

534.21-14 + 621.3].001.362 **1491**

The Formal Connection between the Corpuscular Theory of Sound Propagation in Liquids and Problems of Electrical Engineering, particularly the Theory of Quadripoles.—K. Altenburg. (*Frequenz*, Oct. 1951, Vol. 5, No. 10, pp. 285–289.)

534.213.4 **1492**

Propagation of Sound in a Duct with Constrictions.—U. Ingård & D. Pridmore-Brown. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 689–694.) Iris partitions are uniformly distributed along the duct. Attenuation is determined as a function of frequency for ducts with hard and with absorptive walls, charts being given for hard-wall ducts. By proper choice of wall absorption and iris dimensions and spacing, a wide attenuation band can be obtained. Measured values of attenuation are in good agreement with theory.

534.231 : 534.26 **1493**

On the Relation between the Sound Fields Radiated and Diffracted by Plane Obstacles.—F. M. Wiener. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 697–700.) The analogy between acoustic diffraction and radiation problems is discussed; a very simple relation exists between a plane rigid scatterer exposed to a perpendicularly incident plane wave and a plane piston radiator. The radiated field is proportional to the scattered field. A similar relation exists between the radiation impedance and the force per unit pressure exerted on the scatterer.

534.232 : 538.652 **1494**

Magnetostriction Transducer Measurements.—H. J. Round. (*Wireless Engr*, April 1952, Vol. 29, No. 343, pp. 101–105.) Description of a simplified method, based on equivalent circuits, for determining the constants of a magnetostriction oscillator.

534.24 : 534.213.4 **1495**

An Analysis of the Effect of the Discontinuity in a Bifurcated Circular Guide upon Plane Longitudinal Waves.—L. L. Bailin. (*Bur. Stand. J. Res.*, Oct. 1951, Vol. 47, No. 4, pp. 315–335.) A theoretical treatment of the scattering of sound by a semi-infinite tube of small diameter inserted axially into a larger tube of infinite length. The theory is rigorous and explicit provided the waves incident at the discontinuity are restricted to the lowest mode of propagation. The problem is a particular case of the general problem of the effects of obstacles on the propagation of acoustic or e.m. waves in guides.

534.321.9.047 : 621.431.75 **1496**

Ultrasonic Spectra emitted by Aircraft Propulsion Mechanisms, and their Physiological Effects.—P. Grognot. (*Ann. Télécommun.*, Nov. 1951, Vol. 6, No. 11, pp. 341–344.)

534.373-13 **1497**

The Absorption of Sonic and Ultrasonic Waves in Gases.—M. Dubois. (*J. Phys. Radium*, Nov. 1951, Vol. 12, No. 9, pp. 876–884.) Theoretical and experimental investigation of the general problem of the attenuation of a plane wave due to the physical properties of the medium and to variations of pressure, temperature and molecular energy caused by the passage of the wave. 80 references.

534.422 : 534.321.9.043/.047 **1498**

A New Improved Type of Ultrasonic Siren.—L. Pimonow. (*Ann. Télécommun.*, Nov. 1951, Vol. 6, No. 11, pp. 337–341.) An earlier design (1822 of 1951) has been improved by truncating the rotor, increasing its speed, and introducing a reflecting surface.

534.522 **1499**

Theory of Optical Method of Sound Analysis.—J. Picht. (*Ann. Phys., Lpz.*, 1949, Vol. 5, Nos. 3/5, pp. 117–132 & 1951, Vol. 9, No. 8, pp. 381–400.) Development of the theory and derivation of formulae applicable to the method described by Schouten (1549 of 1939, 1062 and 1452 of 1940), and modification of the method for combinations of acoustic frequencies with arbitrary phase relations.

534.75 **1500**

Compression Properties of the Ear.—H. Mol. (*Tijdschr. ned. Radiogenoot.*, Nov. 1951, Vol. 16, No. 6, pp. 277–291.) Analysis of the compressive action of the ossicle chain. The large dynamic range of the human ear is ascribed to this.

534.79 : 534.839 **1501**

On the Measurement of the Loudness of White Noise.—I. Pollack. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 654–657.) A scale of loudness for white noise was obtained by independent subjective methods which show consistency among themselves. It is suggested that this scale forms a better measure for complex sounds than a pure-tone scale.

534.79 : 534.839 **1502**

On the Threshold and Loudness of Repeated Bursts of Noise.—I. Pollack. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 646–650.) Full paper. Summary noted in 543 of 1951.

534.79 : 534.839 **1503**

Sensitivity to Differences in Intensity between Repeated Bursts of Noise.—I. Pollack. (*J. acoust. Soc. Amer.*, Nov. 1951, Vol. 23, No. 6, pp. 650–653.)

534.833.4-13/-14 **1504**

Absorption of Sound in Fluids.—J. J. Markham, R. T. Beyer & R. B. Lindsay. (*Rev. mod. Phys.*, Oct. 1951, Vol. 23, No. 4, pp. 353–411.) A critical review and unified presentation of the fundamental theories of absorption of sound in gases and liquids, together with an outline of experimental techniques. A summary of the reliable experimental results (mainly at frequencies between 1 Mc/s and 100 Mc/s) is included and these are interpreted in terms of the theoretical concepts. 225 references.

534.844.1 : 621.396.615.11 **1505**

Equipment for Acoustic Measurements: Part 3—Acoustic Pulse Measurements.—Mayo, Beadle & Wharton. (See 1681.)

534.846.4 **1506**

Speech Reinforcement in St. Paul's Cathedral.—P. H. Parkin & J. H. Taylor. (*Wireless World*, Feb. & March 1952, Vol. 58, Nos. 2 & 3, pp. 54–57 & 109–111.) Discussion and illustrated description of the system recommended, showing results of acoustic tests. The dome area is served by two 11-ft vertical arrays of loudspeakers mounted beside the pulpit and the lectern; they comprise eleven 10-in. and nine 3½-in. loudspeakers covering the restricted range of 250–4 000 c/s with crossover at 1 000 c/s. Six 6-ft loudspeaker arrays mounted on piers in the nave are fed through a time-delay mechanism consisting of a turntable carrying an 11-in. disk of plastic magnetic material with recording, playback and erasing heads, an ultrasonic erasing head being provided to guard against failure of the magnetic erasing equipment.

534.85 : 681.84 **1507**

New Developments in the Gramophone World.—L. Alons. (*Philips tech. Rev.*, Nov. 1951, Vol. 13, No. 5, pp. 134–144.) Properties required of record disks and reproducing apparatus for long playing are discussed: a description is given of Philips apparatus which gives a playing time of 22½ min for a microgroove 12-in. record and which also takes ordinary records.

534.861.1 **1508**

Orchestral Studio Design. Recent Modifications to the B.B.C. Maida Vale Studio.—T. Somerville & H. R. Humphreys. (*Wireless World*, April 1952, Vol. 58, No. 4, pp. 128–131.) The acoustic properties of this studio in its original state were characterized by excessive reverberation time at low frequencies and extreme deadness at high frequencies. Architectural modifications which have led to very large improvements are described; these include the application to the walls of box-type 'membrane absorbers' resonant at various low frequencies, and the rebuilding of the orchestra platform.

621.395.61/.62 : 621.395.625.3 **1509**

Investigation of Transducers by Repeated and Retrograde Re-recording.—W. Meyer-Eppler. (*Fernmeldetech. Z.*, Nov. 1951, Vol. 4, No. 11, pp. 507–512.) Analysis showing how slight linear distortion occurring in an electroacoustic transducing system due to transit-time effects may be determined by repeated re-recording. By reversing the direction of motion of the magnetic tape, phase distortion can be totally compensated. Application of the technique in acoustic tests of rooms is outlined.

621.395.623.7 **1510**

Direct Radiator Loudspeaker Enclosures.—H. F. Olson. (*Audio Engng*, Nov. 1951, Vol. 35, No. 11, pp. 34–38 . . 64.) A comprehensive analysis of the effects of various shapes of cabinet shows that for optimum performance the cabinet front must have no sharp edges.

621.395.625.3 **1511**

Magnetic Sound-Recording.—F. Duchâteau. (*HF, Brussels*, 1951, No. 11, pp. 303–312.) Discussion of the main technical difficulties involved and of the means devised to overcome them.

621.396.645.029.3 **1512**

An Ultra-Linear Amplifier.—D. Hafler & H. I. Keroes. (*Audio Engng*, Nov. 1951, Vol. 35, No. 11, pp. 15–17.) The screen-grid of a tetrode is energized with d.c. from a low-impedance source through a special winding on an Acrosound TO-300 output transformer, in which the effects of the anode and screen-grid currents are combined. The screen-grid load impedance must be about 18·5% of the anode load impedance. A circuit incorporating this output stage is shown diagrammatically, with component values. The power output of over 20 W is undistorted within 1 db from 20 to 20 000 c/s, intermodulation being less than 2%.

AERIALS AND TRANSMISSION LINES

621.392.26 **1513**

The Susceptance of a Thin Iris in Circular Wave Guide with the TM_{01} Mode Incident.—K. L. Dunning & R. G. Fellers. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1316–1320.) "Certain wave-guide boundary value problems can be formulated in terms of lumped-constant circuits and distributed-constant transmission lines. Equivalent circuit voltages and currents can be introduced as measures of the transverse electric and magnetic fields. Making use of these concepts and Schwinger's integral equation method, the susceptance of a thin circular iris in circular cylindrical wave guide with the TM_{01} mode incident is discussed and calculated. Results are compared with experimental data."

621.392.26.012.3 **1514**

$TM_{1,1}$ Waves in Rectangular Waveguides.—(*Radio & Televis. News, Radio-Electronic Engng Section*, Sept. 1951, Vol. 46, No. 3, p. 32.) Nomogram for determination of cut-off frequency from dimensions of guide.

621.392.43.012.3 **1515**

Matching-Stub Calculations.—S. Yamasita. (*Radio & Televis. News, Radio-Electronic Engng Section*, Aug. 1951, Vol. 46, No. 2, pp. 32, 31.) A nomogram for determining the position and length of matching stubs on Lecher-wire lines.

621.396.67 **1516**

General Theory of Symmetric Biconical Antennas.—S. A. Schelkunoff. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1330–1332.) The input admittance of a biconical aerial of arbitrary angle is expressed as the limit of a certain sequence of functions. The first term of this sequence approaches the exact expressions for input admittance as the cone angle approaches either zero or 90°; hence it probably constitutes a good first approximation for all angles.

621.396.67 : 621.397.6 **1517**

Antennas for U.H.F.—E. C. Johnson & J. D. Callaghan. (*FM-TV*, Nov. 1951, Vol. 11, No. 11, pp. 16–18 . . 56.) Illustrated descriptions of different types of aerial for television reception, showing their field patterns and gain characteristics.

621.396.677 **1518**

The Necessary Number of Elements in a Directional Ring Aerial.—H. L. Knudsen. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1299–1306.) Discussion based on the theories of Page (1862 of 1948 and 308 of 1949) and Stenzel (1929 Abstracts, p. 450). An expression is derived for the characteristic of an array with a finite number of elements. This expression shows that no uniform improvement of the approximation of the array characteristic to the ideal is obtained by increasing the number of elements, but in this respect odd numbers of elements are better than even numbers.

621.396.677.029.6 **1519**

Measurement of Radiation of U.S.W. Aerials.—J. Delcambe. (*HF, Brussels*, 1951, Nos. 11 & 12, pp. 293–302 & 327–335.) If the aerial aperture field distribution is very nearly plane and equiphase, application of the Kottler formulae previously noted [2109 of 1951 (Divoire & Delcambe)] gives numerical values for the directive properties of waveguide, horn and microwave-lens aerials with an accuracy sufficient to meet practical requirements. Gain can be estimated to within 7%. The measurement methods applied and precautions taken are detailed. An appendix describes a microwave bolometer-type wattmeter developed for the purpose.

621.396.677.5† : 621.318.424 **1520**

Ferromagnetic Loop Antennas.—W. J. Polydoroff. (*Radio & Televis. News, Radio-Electronic Engng Section*, Nov. 1951, Vol. 46, No. 5, pp. 11–13 . . 24.) From results on loops with ferrite cores it is concluded that, in their design, (*a*) a balance must be struck between an acceptable value of Q and maximum effective permeability, a Q of 125–150 being considered most suitable, (*b*) cylindrical cores of length/diameter ratio > 10 give greater effective height; (*c*) the winding should cover 80% of the core length; (*d*) the wire, if insulated by vinylite or double cotton covering, may be wound directly on the core; (*e*) the core should be in the shape of hollow tubing. Various applications where reduction in aerial size is important are suggested. See also 2485 of 1946 (Burgess).

621.396.677.6.029.6 : 621.396.931/.933].2 **1521**

Rotating H-Type Adcock Direction-Finders for Metre and Decimetre Wavelengths.—H. G. Hopkins & F. Horner. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 96–97.) Long summary of 598 of March.

CIRCUITS AND CIRCUIT ELEMENTS

621-52 : 621.396.611.3 **1522**

Applications of Electrical Methods of Differentiation to Control Problems.—Rateau. (See 1688.)

621.3.015.7 **1523**

A Pulse Mixing Unit.—R. R. Rathbone & R. L. Best. (*Radio & Televis. News, Radio-Electronic Engng Section*, Sept. 1951, Vol. 46, No. 3, pp. 10–11, 27.) The unit described accepts pulses from up to eight external lines, mixes them at the input, and delivers them with a delay of 0·08 μs as a single output train. Positive pulses of random form with amplitudes from 6 to 60 V are converted to pulses of half-sine-wave form, of duration 0·1 μs, with amplitudes varying by not more than 5%.

621.3.015.7 **1524**

The Pulse Standardizer.—R. R. Rathbone. (*Radio & Televis. News, Radio-Electronic Engng Section*, Nov. 1951, Vol. 46, No. 5, pp. 6–7, 31.) Description of equipment which accepts pulses of random amplitudes (>6 V), with repetition rate up to 3×10^6/sec, and converts them to a set of pulses of uniform amplitude (up to 37 V) and with the same recurrence frequency. Pulses with repetition rates up to 5×10^6/sec can be accepted if a 10% reduction of the maximum output-pulse voltage can be tolerated.

621.3.015.7 : 621.396.6 **1525**

Pulse Circuits for the Millimicrosecond Range.—F. H. Wells. (*J. Brit. Instn Radio Engrs*, Nov. 1951, Vol. 11, No. 11, pp. 491–503.) The circuits described include pulse-shaping circuits, pulse generators, amplifiers, scalers, and recording oscilloscopes. Applications to high-speed coincidence measurements, millimicrosecond time-interval measurements and fast counting are described.

621.3.015.7 : 621.396.619.16 **1526**

Conversion of Rectangular Pulses of Given Width and Variable Height into Rectangular Pulses of Given Height and Variable Width.—W. Vogt. (*Funk u. Ton*, Nov. 1951, Vol. 5, No. 11, pp. 578–584.) In the circuit described, a.m. pulses charge a capacitor in the positively biased grid circuit of a switching valve which is conductive during the capacitor discharge. The discharge time is a function of the charging voltage. The nonlinearity of this relation is studied. Methods of improving the linearity are indicated.

621.314.222.012.3 **1527**

Charts for the Calculation of Mains Transformers.—G. Pavel. (*Funk u. Ton*, Nov. 1951, Vol. 5, No. 11, pp. 561–577.) Design parameters are discussed. The charts relate core size, number of turns, wire diameter, and resistance of primary and secondary windings of mains transformers for powers up to 150 W. Special charts are provided for E- and M-shaped laminated cores of standard materials.

621.314.3† : 621.314.5 **1528**

Magnetic Modulators.—E. P. Felch, V. E. Legg & F. G. Merrill. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 113–117.)

621.316.8 **1529**

The Problem of a Non-ohmic Resistor in Series with an Impedance.—E. B. Moullin. (*Proc. Instn elect. Engrs*, Part I, Nov. 1951, Vol. 98, No. 114, pp. 344–346.) Discussion on 2657 of 1951.

621.316.8.029.5 **1530**

Resistors at Radio Frequency.—T. J. F. Pavlasek & F. S. Howes. (*Wireless Engr*, Feb. 1952, Vol. 29, No. 341, pp. 31–36.) The r.f. resistance and distributed capacitance of resistors of the metallized-filament type enclosed in insulating sleeves were investigated at frequencies from 0·5 to 40 Mc/s. Curves of R_{rf}, R_{dc}, Z/R_{dc} and ϕ are plotted against $f.R_{dc}$, R_{rf} and R_{dc} being respectively the r.f. and d.c. resistances, f the frequency and ϕ the phase angle. A comparison is made with the theoretical values obtained by considering the resistor as a transmission line having distributed resistance and capacitance.

621.318.572 : 621.385.2 **1531**

R.F. Bursts actuate Gas-Tube Switch.—Geisler. (See 1786.)

621.318.572+681.142] : 621.385.5.032.212 **1532**

The Single-Pulse Dekatron.—Acton. (See 1787.)

621.385.2 : 546.289] + 621.314.7 **1533**

Germanium Crystal Valves.—Bettridge. (See 1785.)

621.392.012 : 517.63 **1534**

Block-Diagram Network Transformation.—T. D. Graybeal. (*Elect. Engng, N.Y.*, Nov. 1951, Vol. 70, No. 11, pp. 985–990.) Essentially full text of 1951 A.I.E.E. Pacific General Meeting paper. A convenient method particularly applicable to the analysis and synthesis of servo systems. The Laplace transform equations of the system are expressed in the form of a block diagram. By methods similar to the star-delta transformation, complicated systems may be reduced to one of a few simple forms.

621.392.5 **1535**

A Network Theorem.—E. E. Zepler. (*Wireless Engr*, Feb. 1952, Vol. 29, No. 341, pp. 44–45.) To find the effect of connecting an impedance Z across two points of a network, the impedance may be replaced by one in parallel with the generator. A simple formula is given from which the value of this parallel impedance can be found.

621.392.5 **1536**

Minimum Phase Networks.—J. A. Tanner. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, pp. 418–423.) A 'minimum phase network' is a feedback network having a minimum value of phase lag at every frequency while satisfying a specified gain/frequency characteristic. The properties of such networks are summarized in relation to their use in servo systems.

621.392.5 : 621.396.621.53 **1537**

The Parallel-T Network as a Linear Mixer.—J. S. Nisbet. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, pp. 432–433.) Two oscillations are mixed by applying them across opposite sides of a parallel-T network and taking the output from one of the mid-shunt arms. The circuit operation is analysed. Coupling between the two oscillators is avoided over a small frequency band.

621.392.5 : 681.142 **1538**

Linear Networks with Time-Varying Lumped Parameters.—J. Brodin. (*C. R. Acad. Sci., Paris*, 12th Nov. 1951, Vol. 233, No. 20, pp. 1168–1170.) The equations of a system are expressed by means of an integration operator. The method is applied to the calculation of the pass band of an electro-analogue multiplier with a time-varying network.

621.392.5.029.3 : 621.3.012.3 **1539**

The Prediction of Audio-Frequency Response: No. 1—Circuits with Single Reactance Element.—N. H. Crowhurst. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, pp. 440–443.) The first of a series of data sheets; applicability is not restricted to the a.f. range.

621.392.5.029.3 : 621.3.012.3 **1540**

The Prediction of Audio-Frequency Response: No. 2—Circuits with Two Reactance Elements.—N. H. Crowhurst. (*Electronic Engng*, Dec. 1951–Feb. 1952, Vols. 23–24, Nos. 286–288, pp. 483–489, 33–38 & 82–86.) Formulae and charts are presented applicable to circuits reducible to a combination of series inductance and shunt capacitance, or series capacitance and shunt inductance, together with resistances.

621.392.52 **1541**

Tchebyshev Filters and Amplifier Networks.—V. Belevitch. (*Wireless Engr*, April 1952, Vol. 29, No. 343, pp. 106–110.) Darlington's method of filter synthesis (1361 of 1940) is applied to the design of simple low-pass filters composed of alternate series coils and shunt capacitors. Resistive termination of both ends is considered and also the case where one end is on open circuit. When filters of this type are used as amplifier input or output networks, the prescribed value of the terminal shunt capacitance imposes a physical limitation on the gain-bandwidth product of the stage; this limitation has been studied by Bode (583 of 1946), but additional precision results from the present analysis. More general filters with one open-circuit termination are mentioned and a new method of design is outlined. The results can be extended to other than low-pass filters by means of frequency transformations.

621.392.52 **1542**

RC Networks as Filters.—H. Pieplow. (*Arch. tech. Messen*, Oct. & Dec. 1951, Nos. 189 & 191, pp. T117–T118 & T141–T142.) Common types are analysed and tabulated according to their filter properties. Means of improving the filtering action are considered, and it is found possible, by the addition of amplitude-stabilizing devices, to reduce distortion at the output of an RC oscillator to about 0·1–0·2%. Negative feedback in conjunction with specially designed networks enables amplifiers to be constructed with almost any desired filter properties and stability of operation.

621.392.52 **1543**

Ladder Filters without Attenuation Fluctuations in the Pass Band (Power-Law Filters).—G. Bosse. (*Frequenz*, Oct. 1951, Vol. 5, No. 10, pp. 279–284.) An analytical treatment deriving design formulae for such filters, the overall group delay of which is expressed by a power series. Attenuations, phase delays and transfer functions are shown for symmetrical low-pass filters with 1, 2 and 3 T-type units.

621.392.52 **1544**

Composite Ladder Filters, Second-Order Image Impedances.—R. O. Rowlands. (*Wireless Engr*, Feb. 1952, Vol. 29, No. 341, pp. 50–55.) Design formulae are derived for filters comprising up to three half-sections and having equal or inverse impedance functions of the second order. Their performance is equal to that of the conventional *m*-derived type, but fewer components are required. Other filters are described having a second-order impedance function at only one pair of terminals.

621.392.52 **1545**

The Numerical Calculation of Filter Circuits with Generalized Parameters, using Modern Theory with Special Attention to Cauer's Work.—V. Fetzer. (*Arch. elekt. Übertragung*, Nov. 1951, Vol. 5, No. 11, pp. 499–508.) Cauer applied the operating-parameter theory to the calculation of Tchebycheff-type filters; in the present paper this theory is applied to filters having infinite-attenuation points anywhere in the attenuation band while conforming to Tchebycheff type within the pass band. Explicit formulae are derived for the coefficients of the characteristic function, which is developed as a polynomial; the zero-attenuation points are found by solving the corresponding higher-order equation. The relations between Cauer's Q functions and the characteristic function are shown for both symmetrical and 'antimetrical' filters. Darlington's formulae (1361 of 1940) for the basic low-pass circuit are used to calculate the circuit elements from the no-load impedance.

621.392.52 : 518.4 **1546**

New Graphical Methods for Analysis and Design.—W. Saraga & L. Fosgate. (*Wireless Engr*, March 1952, Vol. 29, No. 342, pp. 68–79.) Methods are developed for transforming a given performance characteristic into a straight line or a set of straight lines. The method is demonstrated by applying it to the analysis and design of image-parameter and insertion-parameter filters.

621.392.6 **1547**

Synthesis of Passive Electrical Networks with *n* Pairs of Terminals and Prescribed Scattering Matrix [matrice de répartition].—V. Belevitch. (*Ann. Télécommun.*, Nov. 1951, Vol. 6, No. 11, pp. 302–312.) Full paper referred to in 2128 of 1951. The term 'matrice de répartition' is used to denote the matrix representing the distribution of power between given terminating impedances.

621.395.661.1 **1548**

Repeater Coil with Divided Secondary Winding and Single Stray-Resonance Peak.—O. Illner. (*Frequenz*, Oct. 1951, Vol. 5, No. 10, pp. 265–272.) From an equivalent circuit the conditions are determined under which only one resonance peak occurs in the response characteristic. In this case the pass band can be about 40% wider than that obtainable when the response characteristic has two peaks. See also 1088 of 1951 (Schmitt & Schrag).

621.396.6-181.4 **1549**

Miniaturization—Crux of Contemporary Product Design.—W. H. Hannahs & B. S. Ellefsen. (*Elect. Mfg, N.Y.*, June 1950, Vol. 45, No. 6, pp. 86–91 . . 200.) Review of techniques applied to circuit components and subassemblies.

621.396.611.4 **1550**

Some Results from the Theory of Coupled Electromagnetic Cavity Resonators.—E. Ledinegg & P. Urban. (*Acta phys. austriaca*, Dec. 1950, Vol. 4, Nos. 2/3, pp. 180–196.) Based on the theory developed previously (1115 of 1951), the coupling frequencies are calculated for some systems of particular interest, e.g. cylindrical cavities coupled at the flat ends by windows or coaxial-line sections. Comparison is made with experimental results.

621.396.615 **1551**

Blocking-Oscillator Amplitude Control.—(*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, p. 439.) Circuits are described in which the amplitude of the oscillator output is adjusted by varying the bias applied to the suppressor grid.

621.396.615 **1552**

Distortion in Beat-Frequency Sources.—C. G. Mayo. (*Wireless Engr*, April 1952, Vol. 29, No. 343, pp. 92–94.) At higher beat frequencies than that at which the two oscillators of a beat-frequency source lock, there is distortion of the beat-frequency waveform. Analysis shows that at any frequency $\omega/2\pi$ the second-harmonic distortion is $\omega_0/2\omega$, where $\omega_0/2\pi$ is the highest beat frequency at which locking occurs.

621.396.615.077.2/.3 **1553**

An Amplidyne Phase Shift Oscillator.—J. C. West. (*J. sci. Instrum.*, Nov. 1951, Vol. 28, No. 11, pp. 336–339.) An oscillator is described capable of developing a peak current of 3 A in a 17-Ω load over the frequency range 0·06–18 c/s. The amplidyne with its associated electronic amplifier has a local feedback loop to reduce the effects of saturation and hysteresis in the magnetic circuit. Application is to the determination of the frequency response of servomechanisms.

621.396.615.17 **1554**

Three-Valve Pulse Generator with Fixed Repetition Rate.—F. A. Benson & G. V. G. Lusher. (*Wireless Engr*, April 1952, Vol. 29, No. 343, pp. 90–91.) A development of the 2-valve generator previously described (1244 of May), producing short positive pulses of amplitude about 50 V, using a square-wave input from either a multivibrator or a clipping circuit.

621.396.615.17 : 621.314.7 **1555**

Transistors as Multivibrators.—I. Queen. (*Radio-Electronics*, Sept. 1951, Vol. 22, No. 12, pp. 92, 94.) A multivibrator using two transistors, and a flip-flop circuit using a single transistor, are described. Both are triggered by a differentiated square-wave voltage.

621.396.615.17.012.3 **1556**

A Nomogram for Multivibrator Design.—W. R. Luckett. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, p. 448.)

621.396.615.18 **1557**

Decade Multivibrator Design. Method of Stabilizing Frequency Division from a Crystal Drive.—J. E. Attew. (*Wireless World*, March 1952, Vol. 58, No. 3, pp. 114–116.) Spasmodic momentary jumping to the next lower ratio of division often occurs in a crystal-driven decade multivibrator, even though synchronized for even division. This is prevented by applying a pulse of opposite polarity just before the true synchronizing pulse. A complete circuit diagram of a stable frequency-division system embodying this principle is shown. Setting-up procedure is outlined.

621.396.645 **1558**

A New Push-Pull Amplifier Circuit.—A. P. G. Peterson. (*Gen. Radio Exp.*, Oct. 1951, Vol. 26, No. 5, pp. 1–7.)

Power-amplifier applications of the circuit described in 1250 of May are considered.

621.396.645 **1559**

Cathode-Coupled Amplifier.—J. A. Lyddiard. (*Wireless Engr*, March 1952, Vol. 29, No. 342, pp. 63–67.) Analysis is presented in which the cathode-coupled amplifier is treated as a cathode follower driving an earthed-grid voltage amplifier. The method is simple, and leads to a straightforward design procedure.

621.396.645 : 621.3.015.3 **1560**

Overvoltage Effect in R.F. Power Amplifiers.—E. Rizzoni. (*Alta Frequenza*, Oct. 1951, Vol. 20, No. 5, pp. 200–209.) The effect of overvoltage at the terminals of the anode oscillatory circuit of a r.f. power amplifier, due to detuning of the anode circuit, is discussed, and a method of calculating it as a function of load admittance and amount of detuning is described. Calculations and graphs are presented for the FIVRE beam tetrode Type 4-C500 and the R.C.A. triode Type 893 A-R.

621.396.645 : 621.315.612.4 **1561**

Dielectric Amplifier Fundamentals.—A. M. Vincent. (*Electronics*, Dec. 1951, Vol. 24, No. 12, pp. 84–88.) The input voltage is applied across a capacitor with ferroelectric dielectric, whose reactance is thereby varied; the capacitor is in circuit with an a.c. power source and load, and amplified power variations appear across the latter. The impedance of the circuit is relatively high. Practical forms of the amplifier are described, and its operation is compared with that of magnetic and valve amplifiers; at present the frequency range appears to have an upper limit at about 10 Mc/s. Numerous applications are indicated.

621.396.645 : 621.392.52 **1562**

Broad-Banding by Stagger Tuning.—R. C. Wittenberg. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 118–121.) Simplified methods are presented for design calculations of multi-circuit stagger-tuned i.f. amplifiers to have either Butterworth or Tchebycheff type of response. Tables and charts are given and their use illustrated. Circuits with bandwidths up to twice the centre frequency are designed using noncritical values of components.

621.396.645.37 **1563**

Highly-Selective Amplification at Low Frequencies.—F. J. Hyde. (*Wireless Engr*, April 1952, Vol. 29, No. 343, pp. 85–90.) Analysis is presented of the unbalanced twin-T RC filter with series arms R and C and shunt arms $aR/2$ and $2C$. The locus of the transmission vector, for $0{\cdot}5 < a < 1{\cdot}0$ is approximately circular and the phase angle varies continuously from 0 to 2π radians. A single-stage amplifier with series feedback through a balanced twin-T filter ($a = 1$) has a Q value of $A/4$ when A, the loaded stage gain without feedback, is large. Unbalancing of the filter by making $a < 1$, gives improved selectivity, and a working Q of 20 can be readily obtained. Experimental results are given for filters with resonance frequencies of about 0·25, 0·5, and 360 c/s.

621.396.645.37 **1564**

Dual Circuit of a Feedback Amplifier.—D. A. Bell. (*Wireless Engr*, Feb. 1952, Vol. 29, No. 341, pp. 40–43.) The dual circuit of a single-stage amplifier is first obtained. The duals of (*a*) a single-stage amplifier with a voltage-feedback branch, (*b*) a combination of two amplifiers with local feedback round each and additional feedback over both stages, are then derived.

621.396.645.37.012.8 **1565**

Equivalent Circuits to Simplify Feedback Design.—R. S. Burwen. (*Audio Engng*, Oct. 1951, Vol. 35, No. 10, pp. 11–12 . . 45.) Equivalent circuits are used in analysis of the general feedback amplifier. Voltage and current feedback are considered separately. The effect of a small amount of feedback on power amplifiers is studied; a 16-Ω loudspeaker winding in the cathode circuit is sufficient to provide 6 db feedback in a single-pentode output stage. Application of equivalent circuits in the design of a preamplifier circuit with a prescribed response curve is described.

621.396.645.371.011.1 **1566**

Expressions for the Reduction of Distortion and Output Impedance in Terms of db of Feedback.—W. J. Kessler & S. E. Smith. (*Audio Engng*, Oct. 1951, Vol. 35, No. 10, p. 13.) The feedback factor is eliminated from the formulae usually applied, in order to obtain formulae expressed in terms of parameters easily measured.

621.3.015.7 **1567**

Pulse Techniques. [Book Review]—S. Moskowitz & J. Racker. Publishers: Prentice-Hall, New York, 1951, 300 pp., $5.00. (*J. Franklin Inst.*, Aug. 1951, Vol. 252, No. 2, p. 203.) "The book is apparently intended as a text for a brief undergraduate course in the pulse aspects of electronics."

621.392 **1568**

Circuits in Electrical Engineering. [Book Review]—C. L. Vail. Publishers: Prentice-Hall, New York, 1950, 560 pp., $5.75. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, p. 1391.) "Primarily intended as a textbook [of circuit analysis] and not as a reference work."

621.396.6 + 621.385 **1569**

Introduction to Electronic Circuits. [Book Review]—Feinberg. (See 1792.)

GENERAL PHYSICS

519.27 : 517.433 **1570**

Two Classes of 'Observation Operators'.—R. Vallée. (*C. R. Acad. Sci., Paris*, 26th Nov. 1951, Vol. 233, No. 22, pp. 1350–1351.) Two types of linear operators are described, with a dual correspondence between the space-time variables on the one hand and the space-periodicity variables on the other. Special cases of these operators are frequently met with in all experimental fields, and they play an important part in information theory.

534.014.2 **1571**

Predominantly Subharmonic Oscillations.—C. A. Ludeke. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1321–1326.) Theory of the demultiplication of the frequency applied to a nonlinear system. Subharmonic resonances of amplitude greater than the applied fundamental are particularly considered. Experiments with electromechanical apparatus confirmed the theory. Transitions between different subharmonics are discussed.

534.21 + 538.56 **1572**

Wiener-Hopf Techniques and Mixed Boundary-Value Problems.—S. N. Karp. (*Commun. pure appl. Math.*, Dec. 1950, Vol. 3, No. 4, pp. 411–426.) The parallelism between the method of separation of variables and the Green's function integral-equation method is shown to hold as in problems of more classical type. This relation leads to a characterization (from the standpoint of co-ordinate systems) of those problems in which the Wiener-Hopf type of problem (in an extended sense) arises. Certain heuristic advantages of the separation-of-variables procedure are also pointed out. Special applications considered include the diffraction of a plane wave by a staggered array of semi-infinite planes, and the e.s. charge distribution on a cone, including the special case of a disk.

535.13 : 538.3 **1573**

Is there an Aether?—P. A. M. Dirac. (*Nature, Lond.*, 24th Nov. 1951, Vol. 168, No. 4282, pp. 906–907.) The difficulty of reconciling the concept of the aether with the principle of relativity is removed by applying quantum mechanics; the existence of an aether is implicit in the new theory of electrodynamics (1574 below).

537.122 **1574**

A New Classical Theory of Electrons.—P. A. M. Dirac. (*Proc. roy. Soc. A*, 7th Nov. 1951, Vol. 209, No. 1098, pp. 291–296.) In the theory of the electromagnetic field without charges, the potentials are not fixed by the field, but are subject to gauge transformations; thus more variables are involved than are physically needed. It is possible by destroying the gauge transformations to make the superfluous variables acquire a physical significance and describe electric charges. This leads to a simplified classical theory of electrons which appears to be more suitable than the usual one as a basis for a passage to the quantum theory.

537.226 : 539.11 **1575**

Note on the Interaction of an Electron and a Lattice Oscillator.—E. P. Gross. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 818–823.)

537.311.1 **1576**

Application of Collective Treatment of Electron and Ion Vibrations to Theories of Conductivity and Superconductivity.—D. Bohm & T. Staver. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 836–837.)

537.311.33 **1577**

The Effect of the Mean Free Path of Electrons on the Electrical Properties of Non-metals.—R. W. Wright. (*Proc. phys. Soc.*, 1st Nov. 1951, Vol. 64, No. 383A, pp. 984–999.) The conductivity, thermoelectric power, Hall coefficient, fractional change of conductivity in a magnetic field and the Nernst, Ettinghausen and Righi-Leduc coefficients are calculated on the Lorentz-Sommerfeld theory, using the most appropriate mean-free-path function for the non-metal concerned. The theoretical variations of the electrical properties with temperature so obtained agree well with experimental results.

537.531 + 539.18] : 535.43 **1578**

Multiple Scattering of Waves.—M. Lax. (*Rev. mod. Phys.*, Oct. 1951, Vol. 23, No. 4, pp. 287–310.) Coherent and incoherent scattering of light, X-rays, neutrons and photon waves are fully discussed. Using the self-consistent field method the scattered fields are derived, including the effects of anisotropic scattering, scattering of quantized waves, creation and absorption of particles, Doppler shifts, and randomly, partially or completely ordered systems of scatterers. The connection between collisions with a multiparticle system and multiple scattering is considered.

538.114 **1579**

Application of the Bethe-Weiss Method to the Theory of Antiferromagnetism.—Yin-Yuan Li. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 721–730.)

538.311 : 621.318.423 : 513.647.1 : 621.385.029.6 **1580**

Properties of the Electromagnetic Field of Helices.—É. Roubine. (*C. R. Acad. Sci., Paris*, 12th Nov. 1951, Vol. 233, No. 20, pp. 1174–1176.) Field distributions corresponding to the theory for the thin-wire helix given in 2978 of 1951 are compared with those corresponding to the theory of the continuous-cylinder guide; the two agree under certain stated conditions, which are satisfied for travelling-wave valves with narrow beams but not for wide-beam valves.

538.522 **1581**

The Proximity Effect and Coefficient of Mutual Induction at High Frequency for a Wire and Part of a Very Thick Plate, both being Conducting and Parallel.—A. Colombani. (*C. R. Acad. Sci., Paris*, 19th Nov. 1951, Vol. 233, No. 21, pp. 1267–1269.) Assuming the wire radius is smaller than the depth of penetration of the current into the wire, the mutual inductance is given by $M = \mu^{\frac{1}{2}}/d\,(\pi\omega\gamma')^{\frac{1}{2}}$, where d is the distance of the wire from the plate, and γ' the conductivity of the plate. This result is extended to the case in which the wire radius is not negligible compared with d.

538.56 : 535.42 **1582**

On Systems of Linear Equations in the Theory of Guided Waves.—W. Magnus & F. Oberhettinger. (*Commun. pure appl. Math.*, Dec. 1950, Vol. 3, No. 4, pp. 393–410.) An investigation of the diffraction of an e.m. wave by a plane strip between two parallel planes, or in a rectangular waveguide, assuming that essentially only one type of wave exists. For an incoming wave of the type $\exp(i\alpha x)\cos\beta y$, α and β being real, the diffracted wave components can be expanded in a Fourier series, whose coefficients are uniquely determined by the condition of the finiteness of the total energy in any finite part of the space; they are given by an infinite set of linear equations. The special case of a diffracting strip half the width of the waveguide is treated in detail: in this case the linear equations can be dealt with by successive approximation, and the first steps can be carried out explicitly.

538.56 : 535.42 **1583**

On the Theory of Electromagnetic-Wave Diffraction by an Aperture in an Infinite Plane Conducting Screen.—H. Levine & J. Schwinger. (*Commun. pure appl. Math.*, Dec. 1950, Vol. 3, No. 4, pp. 355–391.) A procedure for solving this problem exactly has been described by Meixner (94 of 1951), but approximations are required which are suitable for computation and accurate over a frequency range. This paper, a sequel to previous ones concerned with diffraction in a scalar field (83 and 1897 of 1950), describes variational principles for obtaining some of the desired information. A formal description is given of the fields and boundary conditions involved. Expressions for the field vectors in any region are derived in terms of the tangential components of the electric or magnetic field components on its boundary, with the aid of tensor Green's functions. These expressions are first found for the regions on each side of the screen by using integrals involving the tangential electric field over the aperture. From the equality of the tangential magnetic fields for each region in the aperture, an integral equation for the tangential electric aperture field is obtained. From this, a stationary property of the radiation field at large distances from the aperture is derived. Similarly, variational principles are obtained by consideration of the tangential magnetic field over screen and aperture, and of the current over the screen. The connection between the principles and the plane wave 'transmission cross-section', which is a measure of the ratio of energy passing through the aperture per second to that transported per unit area of the incident wave, is demonstrated. Numerical results are given for the cross-section of a circular aperture with normally incident plane wave, and are compared with those from the Kirchhoff and Rayleigh approximations.

538.566 **1584**

The Analytical Expression of Huyghens' Principle for Electromagnetic Waves.—A. da Silveira. (*C. R. Acad.*

Sci., Paris, 19th Nov. 1951, Vol. 233, No. 21, pp. 1269–1272.)

538.566 **1585**

The Magnetic Dipole over the Horizontally Stratified Earth.—J. R. Wait. (*Canad. J. Phys.*, Nov. 1951, Vol. 29, No. 6, pp. 577–592.) Analysis of the radiation characteristics of a vertical magnetic dipole above a two-layer or three-layer earth, with particular reference to the effects produced within the earth. Transient effects are considered in some cases.

538.566 **1586**

Transient Electromagnetic Propagation in a Conducting Medium.—J. R. Wait. (*Geophys.*, April 1951, Vol. 16, No. 2, pp. 213–221.) Expressions are developed by Laplace transformation for the electric fields due to different types of step-function current source in an infinite conducting medium. Sources considered are the electric dipole, the magnetic dipole, and step-function currents in insulated wires of finite and infinite length.

538.566 : 538.63 **1587**

The Influence of Magnetic Fields upon the Propagation of Electromagnetic Waves in Artificial Dielectrics.—E. R. Wicher. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1327–1329.) "An effect corresponding to the Faraday effect in natural dielectrics is predicted for a class of artificial dielectrics because of the existence of a Hall effect in the metallic components of the structure. A formula for Verdet's constant as a function of element polarizability and Hall coefficient is obtained. The resonance shift to be expected in a cavity resonator, filled with an artificial dielectric, and subjected to a strong magnetic field, is calculated."

538.691 **1588**

The Effect of a Magnetic Field on Electrons in a Periodic Potential.—J. M. Luttinger. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 814–817.) A theorem due to Wannier for treating the motion of electrons in a perturbed periodic field is generalized to include the effect of a slowly varying magnetic field. The problem reduces to that of solving a Schrödinger equation.

546.212 : 536.421.4 **1589**

Experimental Investigation of Icing Phenomena.—D. Melcher. (*Z. angew. Math. Phys.*, 15th Nov. 1951, Vol. 2, No. 6, pp. 421–443.) Detailed study of the formation of ice (*a*) artificially in a wind tunnel, (*b*) in the open air, taking account of the influence of an electric field.

501 : 530.12 **1590**

Mathematics of Relativity. [Book Review]—G. Y. Rainich. Publishers: Chapman & Hall, London, 1950, 174 pp., 28s. (*Beama J.*, Nov. 1951, Vol. 58, No. 173, pp. 375, 377.) "This book is an admirable survey of relativity theory and it can well be recommended on that account."

537.1 + 538.1 **1591**

A History of the Theories of Aether and Electricity. [Book Review]—E. Whittaker. Publishers: Nelson, London, 1951, 434 pp., 32s. 6d. (*J. Franklin Inst.*, Nov. 1951, Vol. 252, No. 5, p. 441.) To be completed in two volumes; this first volume deals with classical theories.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 : 551.510.535 **1592**

The Wave-Frequency Dependence of the Duration of Radar-Type Echoes from Meteor Trails.—V. C. Pineo & T. N. Gautier. (*Science*, 2nd Nov. 1951, Vol. 114, No. 2966, pp. 460–462.) Simultaneous measurements on 27·2 and 41·0 Mc/s recorded at the National Bureau of Standards between 1st November 1948 and 1st October 1949 support Lovell's conclusion (3402 of 1948) that the duration of radar echoes from meteor trails is approximately proportional to the square of the wavelength. An indication is given of the methods used.

523.72 : 621.396.822 **1593**

On Bailey's Theory of Amplified Circularly Polarized Waves in an Ionized Medium.—R. Q. Twiss. (*Phys. Rev.*, 1st Nov. 1951, Vol. 84, No. 3, pp. 448–457.) Detailed critical analysis of Bailey's theory (1909 of 1950). The growing waves, which Bailey interprets as amplified waves, can only be excited by reflection. It is contended that Bailey's theory can explain neither the excess r.f. radiation from sunspots nor that from discharge tubes. Power amplification is, however, possible in a drifting ionized medium under certain ideal conditions, which are discussed. See also 624 of 1951.

538.12 : 521.15 **1594**

A Fundamental Theory of the Magnetism of Massive Rotating Bodies.—G. Luchak. (*Canad. J. Phys.*, Nov. 1951, Vol. 29, No. 6, pp. 470–479.) A theory based on a relativistically covariant generalization of Maxwell's equations to include gravitational fields.

550.372 + 550.382 **1595**

An Electromagnetic Interpretation Problem in Geophysics.—L. B. Slichter. (*Geophys.*, July 1951, Vol. 16, No. 3, pp. 431–449.) A flat earth in which permeability μ, conductivity σ and permittivity ϵ vary only with depth, is considered subjected to an alternating field produced by a vertical magnetic dipole above the surface. Expressions for the variation of μ, σ and ϵ are obtained in the form of Taylor's series, the coefficients of which may be determined by measurement of the magnetic field intensity H at the surface. The horizontal and vertical components of H above the surface are shown to be mutually dependent, and formulae independent of the electrical characteristics of the ground are derived which connect the two.

550.381 **1596**

Measurements of the Variation with Depth of the Main Geomagnetic Field.—S. K. Runcorn, A. C. Benson, A. F. Moore & D. H. Griffiths. (*Philos. Trans. A*, 27th Nov. 1951, Vol. 244, No. 878, pp. 113–151.) The main geomagnetic field is attributable either to a source seated at the core or to a fundamental property of rotating matter, corresponding to a source distributed throughout the earth. Measurements made in five mines in northern England provide evidence in favour of the core theory.

550.384.4 **1597**

The Equatorial Electrojet as Detected from the Abnormal Electric Current Distribution above Huancayo, Peru, and elsewhere.—S. Chapman. (*Arch. Met. Geoph. Bioklimatol. A*, 1951, Vol. 4, pp. 368–390. In English.) Abnormally large daily variations of the horizontal component of magnetic force observed at Huancayo indicate the daily rise and decline of a concentrated eastward electric current, termed 'equatorial electrojet', above the station; similar effects have been observed at stations in Africa and India. The influence on the phenomenon of position with respect to geographic and magnetic equators is examined, and observations required to determine the height, intensity, width and return current flow are discussed.

551.5 + 550.37 + 550.38 **1598**

General Assembly of the International Union of Geodesy and Geophysics, Brussels, 1951.—H. W. L. Absalom.

(*Met. Mag.*, Nov. 1951, Vol. 80, No. 953, pp. 326–330.) Brief report of the proceedings; recent work in the fields of meteorology and terrestrial magnetism and electricity was reported and discussed.

551.510.5 **1599**

Abrupt Seasonal Changes in Tropopause Level and Stratosphere Temperature at Habbaniya.—D. Dewar. (*Met. Mag.*, Nov. 1951, Vol. 80, No. 953, pp. 323–326.)

551.510.53 : 551.557 **1600**

Evidence for a Stratospheric Circulation in Vertical Meridional Planes between Polar and Equatorial Regions in Winter.—L. S. Clarkson. (*Met. Mag.*, Nov. 1951, Vol. 80, No. 953, pp. 309–318.)

551.510.535 **1601**

Some Characteristics of the Ionosphere E Region.—K. Rawer & É. Argence. (*C. R. Acad. Sci., Paris*, 12th Nov. 1951, Vol. 233, No. 20, pp. 1208–1210.) Recent observational data indicate either that the dissociation of O_2 takes place at a height greater than that suggested by Penndorf (2224 of 1949) or that the ionization process does not involve the dissociation of O_2.

551.510.535 **1602**

The Half-Year Period in the Ionization of the F_2 Layer.—O. Burkard. (*Arch. Met. Geoph. Bioklimatol. A*, 1951, Vol. 4, pp. 391–402. In German.) The amplitude and phase of the observed half-yearly variations of ionization depend on the geographical location of the observation stations. A straightforward explanation of the effect is based on the assumption that the intensity of the solar ultraviolet radiation depends on latitude and is least at the solar equator. This theory also explains the half-yearly variation of apparent height of the layer.

551.510.535 : 550.386 **1603**

Geomagnetic Bays and their Relation to Ionospheric Currents.—H. Wiese. (*Z. Met.*, Nov. 1951, Vol. 5, No. 11, pp. 341–347.) Daily and yearly variations of the frequency of occurrence of bays as shown by the geomagnetic records obtained at Niemegk during the period 1937–1944 are investigated. A tendency to repetition at 27-day and 24-hour intervals is indicated. Correlation with ionospheric air currents is observed; air currents in the ionosphere are related to those in lower atmospheric layers, at least in winter.

551.510.535 : 551.510.4 **1604**

Ozone Measurements during Sudden Ionospheric Disturbances.—S. Fritz. (*Arch. Met. Geoph. Bioklimatol. A*, 1951, Vol. 4, pp. 343–350. In English.) The measurements were made in order to study the effect on total atmospheric ozone of the enhanced solar emission of ultraviolet radiation associated with the flares causing the ionospheric disturbances. On the assumption that the ratio of the extraterrestrial intensity of sunlight at 3 110 Å to that at 3 300 Å (the observation wavelengths) is unaffected, the observations indicate that the variation of ozone content due to the disturbances is small or nil, as would be expected from theoretical considerations.

551.510.535 : 621.396.11 **1605**

Ionosphere Review: 1951. Greatly Reduced Rate of Decrease in Sunspot Activity and M.U.Fs.—T. W. Bennington. (*Wireless World*, March 1952, Vol. 58, No. 3, pp. 121–122.) Shows the monthly mean sunspot numbers and F_2-layer noon and midnight critical frequencies from the last sunspot maximum to 1951, and 12-month running averages of the same three quantities since the last sunspot minimum. Solar activity is likely to decrease slowly during 1952. An estimate is made of s.w. propagation conditions during 1952.

551.510.535 : 621.396.72 : 621.3.087.47 **1606**

Ionospheric Sounding Stations.—(*U.R.S.I. Inform. Bull.*, Nov./Dec. 1951, No. 72, pp. 20–23.) Stations in Austria and those under the Bureau Ionosphérique Français and the Service de Prévision Ionosphérique Militaire are listed, with operating data.

551.515.4 **1607**

The Electrical and Meteorological Conditions Inside Thunderclouds.—J. Kuettner. (*J. Met.*, Oct. 1950, Vol. 7 No. 5, pp. 322–332.)

551.594.12 **1608**

Height Variations in the Concentration of Ions near the Ground during Quiet Summer Nights at Uppsala.—H. Norinder & R. Siksna. (*Tellus*, Nov. 1951, Vol. 3, No. 4, pp. 234–239.)

LOCATION AND AIDS TO NAVIGATION

621.396.9 **1609**

Origins of Radar. Background to the Awards of the Royal Commission.—(*Wireless World*, March 1952, Vol. 58, No. 3, pp. 95–99.) An account of the development of radar in Britain from 1935 to the war years, based on evidence given before the Royal Commission on Awards to Inventors.

621.396.9 : 526.9 **1610**

The Effect of Meteorological Conditions on the Measurement of Long Distances by Electronics.—C. I. Aslakson & O. O. Fickeissen. (*Trans. Amer. geophys. Union*, Dec. 1950, Vol. 31, No. 6, pp. 816–826.) Surveying using shoran (frequency range 220–350 Mc/s) requires data on meteorological conditions between the two receiving stations. From these the changes in atmospheric refractive index with height are computed and the necessary corrections to velocity of wave propagation and path length determined. Correction methods are described, with numerical examples.

621.396.9 : 526.9 **1611**

Accuracy in Electromagnetic Distance Measurement.—J. Moline. (*Radio franç.*, Nov. 1951, No. 11, pp. 1–5.) Principles of pulse, f.m. and phase-displacement methods of distance measurement by means of e.m. waves reflected from the distant point are outlined, including that combining pulse and phase measurement. In phase-displacement systems accuracy to within $\lambda/100$ is attainable for frequencies up to 30 Mc/s.

621.396.9 : 551.5 **1612**

Abnormal Displacement of some Echoes from Rain.—R. Lhermitte. (*C. R. Acad. Sci., Paris*, 12th Nov. 1951, Vol. 233, No. 20, pp. 1210–1212.)

621.396.9 : 551.578.4 **1613**

Some Quantitative Measurements of Three-Centimeter Radar Echoes from Falling Snow.—R. C. Langille & R. S. Thain. (*Canad. J. Phys.*, Nov. 1951, Vol. 29, No. 6, pp. 482–490.) The variation of back-scatter intensity with rate of snowfall was observed during four storms. Analysis of the size distribution of snowflakes, which appears to be important in calculating radar echo intensity, was only carried out for one storm. Observed echo intensities are in fair agreement with values calculated from Ryde's formula (2062 of 1948).

621.396.9 : 551.594.22 **1614**

Lightning Detection by Radar.—M. G. H. Ligda. (*Bull. Amer. met. Soc.*, Oct. 1950, Vol. 31, No. 8, pp. 279–283.) Description of the method used, showing records obtained.

621.396.93 : 621-526 **1615**

Analysis and Construction of a Position-Fixing Servomechanism.—Klein. (See 1739.)

621.396.932 **1616**

Radio Aids to Marine Navigation: The Seaman's Requirements.—F. J. Wylie. (*J. Brit. Instn Radio Engrs*, Nov. 1951, Vol. 11, No. 11, pp. 478–490.)

621.396.933 **1617**

Aircraft Navigational Aids.—(*Engineer, Lond.*, 30th Nov. 1951, Vol. 192, No. 5001, pp. 702–703.) An account of the facilities provided by (*a*) the Marconi v.h.f. d.f. system for instantaneous visual indication of bearings and position fixing by the ground station, (*b*) the Mullard 'telescribe' equipment in experimental use at London airport, by which written messages, maps, etc., are instantaneously reproduced on the c.r. screen of a distant receiving unit, and (*c*) the Decca 'flight log'.

621.396.933 **1618**

A Simplified Multiple-Track Range Airborne Equipment.—R. S. Styles. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 88–92.) Description of Australian M.T.R. light-weight equipment suitable for installation in small aircraft. In association with the appropriate ground-station installation it provides pilots with accurate azimuthal track guidance for a range of 100 miles, flying at 5 000 ft, and/or a localizer path, for use during final approach to a runway, accurate to within ±22 yd at the normal touch-down point.

621.396.933 : 629.13.053 **1619**

Automatic Track-Plotting Instrument for Aircraft.—(*Engineering, Lond.*, 9th Nov. 1951, Vol. 172, No. 4476, p. 589.) Description of the operation of the Decca 'flight log' in use with the Mark VII receiver, which displays aircraft position and ground track directly on a chart. See also 2192 of 1951.

621.396.933.2.001.4 **1620**

ILS Field Test Set.—Ellis. (See 1685.)

629.13.05 : 538.74 **1621**

Stroboscopic Earth-Inductor Compass.—S. A. Schwartz. (*Elect. Engng, N.Y.*, Nov. 1951, Vol. 70, No. 11, pp. 1001–1003.) A compass is described in which a coil carrying a compass card is rotated in the earth's magnetic field by a small air turbine. The sinusoidal voltage induced in the coil is amplified and squared, and the leading edge of the square wave is used to trigger a stroboscopic source of light. When observed by this light the compass card appears to be stationary and indicates a direction relative to magnetic north.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.787.7 **1622**

Sensitive Differential Manometer.—J. M. Los & J. A. Morrison. (*Rev. sci. Instrum.*, Nov. 1951, Vol. 22, No. 11, pp. 805–809.) The mercury surfaces in the manometer arms serve as the moving plates of two separate parallel-plate capacitors which form part of the oscillatory circuits of two similar 3·6-Mc/s oscillators. The capacitance change, due to a differential pressure change, varies the beat frequency of the oscillators, which is measured by reference to an a.f. signal generator. An accuracy to within 0·1–0·2 μ is obtained for pressure changes between 0 and 0·02 cm Hg. For larger pressure changes (up to 0·25 cm Hg) the accuracy is within 0·1%.

535.37 **1623**

Recent Developments in Luminescent Materials.—S. T. Henderson. (*Research, Lond.*, Nov. 1951, Vol. 4, No. 11, pp. 492–497.) The subject is considered under the headings (*a*) research on known materials to discover the mechanism of luminescence, (*b*) investigations to improve utility and extend applications of known materials and (*c*) discovery of new materials. 105 references.

537.228.1 : 546.431.824-31 **1624**

Electrical, particularly Piezoelectric Properties of Barium Titanate.—J. H. van Santen & G. H. Jonker. (*Tijdschr. ned. Radiogenoot.*, Nov. 1951, Vol. 16, No. 6, pp. 259–274. Discussion, pp. 275–276.) These properties are discussed in relation to crystal structure. The piezoelectricity of pre-polarized $BaTiO_3$ can be considered as a combination of a linear electrostriction and a piezoelectric effect. Possible applications of $BaTiO_3$ ceramics are noted.

537.311.33 + 535.37 **1625**

New Views on Oxidic Semi-Conductors and Zinc-Sulphide Phosphors.—E. J. W. Verwey & F. A. Kröger. (*Philips tech. Rev.*, Oct. 1951, Vol. 13, No. 4, pp. 90–95.) The mechanism of conduction in nonstoichiometric oxidic semiconductors is similar to that in other semiconductors in which desired changes of valency are produced by the admixture of suitable impurities. Members of the latter group are not characterized by lack of thermal stability (due to vacant lattice sites) as are nonstoichiometric compounds. Similar considerations are applicable to ZnS phosphors, in which the fluorescence centres consist of activator ions surrounded by sulphur ions. The accuracy of this hypothesis is confirmed by the fact that the role of the halogen ions in the formation of Zn phosphors may be taken over by trivalent cations.

537.311.33 : 517.944 **1626**

A Note on the Partial Differential Equations describing Steady Current Flow in Intrinsic Semiconductors.—R. C. Prim. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1388–1389.)

537.311.33 : 537.211 **1627**

An Effect of Light on Semiconductors: Variation of the Contact Potential Difference.—W. Veith & G. Wlérick. (*C. R. Acad. Sci., Paris*, 5th Nov. 1951, Vol. 233, No. 19, pp. 1097–1101.) Measurements were made of the contact potential with and without illumination for a CdS layer in vacuum, using a retarding-potential method. The results support a formula obtained by a classical calculation [see e.g. 4444 of 1940 (Mott & Gurney)]. Variation of the contact potential with temperature is also noted.

537.311.33 : 546.482.21 **1628**

Measurements of the Electrical Conductivity of CdS Crystals Irradiated by Medium-Energy Electron Beams.—H. Benda. (*Ann. Phys., Lpz.*, 15th Nov. 1951, Vol. 9, No. 8, pp. 413–422.)

537.312.8 : 546.87 **1629**

The Anomalous Magnetoresistance of Bismuth at Low Temperatures.—P. B. Alers & R. T. Webber. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 863–864.) Results are reported of measurements made on Bi crystal rods 2 mm in diameter and 2–3 cm long, using transverse fields of strength 60 kilogauss and over.

537.533.9 : 537.226 **1630**

Direct Demonstration of the Conductivity of a Thin Dielectric under Electron Bombardment.—C. Dufour. (*J. Phys. Radium*, Nov. 1951, Vol. 12, No. 9, pp. 887–888.) Measurements were made of the current through a composite target comprising an evaporated layer of ZnS between evaporated layers of Al across which a variable voltage was applied.

538.221 **1631**

Systematic Relations between Hysteresis, Creep of Nonlinearity Products and the Richter After-Effect.—R. Feldtkeller, H. Wilde & G. Hoffmann. (*Z. angew. Phys.*, Nov. 1951, Vol. 3, No. 11, pp. 401–409.) Measurements made on three similarly treated specimens of Si/Fe-alloy stampings are reported; systematic differences between them are discussed. The measurement equipment is described.

538.221 **1632**

Ferromagnetic Resonance and the Internal Field in Ferromagnetic Materials.—J. R. MacDonald. (*Proc. phys. Soc.*, 1st Nov. 1951, Vol. 64, No. 383A, pp. 968–983.) "A classical treatment of the domain energy terms of a homogeneous ferromagnetic solid leads to a formula for the internal field contributions from these terms. With this result, modifications in the resonance condition of ferromagnetic resonance arising from self energy, exchange energy, magnetocrystalline anisotropy and applied or intrinsic stress are obtained and are applied to various crystalline anisotropy and stress conditions of interest in ferromagnetic resonance experiments. Finally, the bearing of the results on the anomalous *g*-values obtained in resonance experiments is considered."

538.221 **1633**

The Magnetization Process in Ferrites.—H. P. J. Wijn & J. J. Went. (*Physica*, Nov./Dec. 1951, Vol. 17, Nos. 11/12, pp. 976–992.) "The initial magnetization curve of ferrites has been measured as a function of frequency up to 2 Mc/s. It has been found that the magnetization of sintered ceramic ferrites with a high permeability is brought about by at least two processes, one of which, in the frequency range covered, is independent of frequency and determines the initial permeability. The other process has a relaxation frequency of about 200 kc/s and is responsible for the irreversible processes during magnetization. From measurements on samples of sintered ferrites fired at different temperatures it has been concluded that the frequency-dependent magnetization is caused by irreversible Bloch-wall displacements, whilst the initial permeability is caused by a reversible rotation of the magnetization in Weiss domains in the direction of the external magnetic field (in contrast to what is believed to be the case in cast ferromagnetic metals). A discussion shows that neither eddy current effects nor any inertia effects so far known are responsible for the relaxation frequency of the Bloch wall at about 200 kc/s."

539.23 : 537.311.31 **1634**

Variation, as a Function of Temperature and Applied E.M.F., of the Electrical Resistance of Very Thin Metal Films Deposited on Diamond, Amber and Plexiglass.—N. Mostovetch & T. Duhautois. (*C. R. Acad. Sci., Paris*, 19th Nov. 1951, Vol. 233, No. 21, pp. 1265–1267.) The results do not differ appreciably from those previously obtained (1701 and 2535 of 1950); they suggest that the semiconductivity of very thin films is not an impurity-type semiconductivity in which the support plays an essential part.

546.289 + 546.815.221] : 537.311.33 **1635**

A Study of Rectification Effects at Surfaces of Germanium and Lead Sulphide.—C. A. Hogarth & J. W. Granville. (*Proc. phys. Soc.*, 1st Nov. 1951, Vol. 64, No. 383B, pp. 992–998.) Chemical etching or thermal treatment in vacuo can remove the amorphous surface layer produced by polishing. Heat treatments up to 900°C can improve the rectification characteristics considerably, and may be preferable to etching. Surface modifications resulting from the various treatments were examined by electron diffraction.

546.289 : 539.164.9 **1636**

Electron-Hole Production in Germanium by Alpha-Particles.—K. G. McKay. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 829–832.) The number of electron-hole pairs produced in Ge by alpha-particle bombardment was determined by collecting the internally produced carriers across a reverse-biased *n-p* junction. There was no evidence of trapping of carriers in the barrier region. The energy lost by a bombarding particle per electron-hole pair produced is $3 \cdot 0 \pm 0 \cdot 4$ eV. The difference between this and the energy gap is attributed to losses to the lattice from the internal carriers.

546.289 : 539.185.9 **1637**

Evidence for Production of Hole Traps in Germanium by Fast Neutron Bombardment.—J. W. Cleland, J. H. Crawford, Jr, K. Lark-Horovitz, J. C. Pigg & F. W. Young, Jr. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 861–862.)

546.289 : 621.314.7 : 535.215 **1638**

The *n-p-n* Junction as a Model for Secondary Photoconductivity.—K. G. McKay. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 833–835.) Experiments are discussed in which a Ge *n-p-n* junction is subjected to bombardment by alpha particles, producing excess-hole currents in the *p* region. Since the secondary currents observed in photoconductive insulators have similar characteristics, study of the *n-p-n* junction is expected to lead to better understanding of secondary photoconductivity.

547.476.3 : 537.228.1 **1639**

Rochelle-Salt Specimens moulded from Crystal Plates under High Pressure.—F. Blaha. (*Acta phys. austriaca*, Dec. 1950, Vol. 4, Nos. 2/3, pp. 272–277.) Disks cut from Rochelle-salt crystals were moulded into pastilles by application of pressure up to 17 000 kg/cm² in the direction of the α axis. Conductivity and permittivity measurements over a range of temperatures are given.

549.211.091.3 **1640**

The Effect of Inhomogeneities on the Electrical Properties of Diamond.—A. J. Ahearn. (*Phys. Rev.*, 15th Nov. 1951, Vol. 84, No. 4, pp. 798–802.)

549.514.51 **1641**

A New Crystal Cut for Quartz with Zero Temperature Coefficient.—E. J. Post. (*Appl. sci. Res.*, 1950, Vol. B1, No. 6, pp. 420–428.) See 115 of 1950.

621.314.634 **1642**

Reversible Changes in the Boundary Layer of Selenium Rectifiers.—A. Hoffmann, F. Rose, E. Waldkötter & E. Nitsche. (*Z. Naturf.*, Aug. 1950, Vol. 5a, No. 8, pp. 465–467.) Two observed phenomena are discussed: (*a*) an increase of working resistance on changing over from application of forward voltage only to application of alternating voltage; (*b*) a decrease of the capacitance of the boundary layer with increase of the operating bias voltage, for the same instantaneous total voltage. Both effects can be explained by assuming a migration of impurity centres due to increased mean field strength at the boundary.

621.314.7 **1643**

Transistors: Part 2—Physics and Construction of the Transistor.—J. Malsch. (*Arch. elekt. Übertragung*, Sept. & Oct. 1951, Vol. 5, Nos. 9 & 10, pp. 425–433 & 467–473. Addendum, ibid., Feb. 1952, Vol. 6, No. 2, pp. 73–79.) A survey paper. Part 1: 2726 of 1951.

621.315.616 : 547-128† **1644**

Silicone Rubber emerges as a Dielectric Material.—J. F. Dexter. (*Elect. Mfg, N.Y.*, June 1950, Vol. 45, No. 6, pp. 100–103 . . 204.) Results of tests on the effects of aging and temperature on the physical and electrical properties of silicone rubber are shown. Samples with brittle points at −90°C are serviceable at 250°C and withstand temperatures up to about 175°C indefinitely. They show exceptional resistance to heat, cold, moisture, oxidation, corona discharge and fatigue. Points of production technique are noted.

621.396.622.63 **1645**

The Temperature Dependence of the Static Characteristics of Crystal Rectifiers, and its Theoretical Significance.—K. Seiler. (*Z. Naturf.*, July 1950, Vol. 5a, No. 7, pp. 393–397.) An experimental investigation was made of rectifiers composed of a layer of Si with traces of Al, in combination with a Mo contact point; the temperature range covered was −80°C to +95°C. I/V characteristics are plotted; the significance of the results for determining the concentration of impurity centres is discussed.

669.715 : 537.311.31 **1646**

Effect of Alloying Elements on the Electrical Resistivity of Aluminum Alloys.—A. T. Robinson & J. E. Dorn. (*J. Metals*, June 1951, Vol. 3, No. 6, pp. 457–460.) "The electrical resistivities of aluminum alloys containing Cu, Ge, Zn, Ag, Cd, and Mg were found to increase linearly with the atomic percentage of the solute atoms. Application of Linde's rule to these data suggests that each aluminum atom contributes 2·5 electrons to the metallic bond."

778.3 : [621.317.755 + 621.397.621.2 **1647**

Photography of Oscillograms and Television Images.—H. Aberdam. (*Toute la Radio*, Nov. & Dec. 1951, Nos. 160 & 161, pp. 339–342 & 365–368.) An examination of the technique, with particular reference to the photographic emulsion required, the spectral brightness of the scanning spot, and the fluorescence of the screen.

621.315.59 **1648**

Semiconducting Materials. [Book Review]—H. K. Henisch (Ed.). Publishers: Academic Press, New York, 1951, 281 pp., $6.80. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 336–338.) Proceedings of conference held at the University of Reading in July 1950; contains the full text of the 28 papers presented.

MATHEMATICS

681.142 **1649**

A General Purpose Differential Analyser: Part 1—Description of Machine.—G. L. Ashdown & K. L. Selig. (*Elliott J.*, Sept. 1951, Vol. 1, No. 2, pp. 44–48.) This analyser is designed for robustness and for rapidity of problem setting rather than for high accuracy. The integrator is of ball-and-disk type. Mechanically independent units are connected by servo-links with electrical connections through a central cross-connection panel.

681.142 **1650**

The Use of the EDSAC for Mathematical Computation.—M. V. Wilkes. (*Appl. sci. Res.*, 1950, Vol. B1, No. 6, pp. 429–438.) A simple explanation of the constituent elements of any programme.

681.142 **1651**

On the Background of Pulse-Coded Computors.—T. J. Rey. (*Electronic Engng*, Jan. & Feb. 1952, Vol. 24, Nos. 287 & 288, pp. 28–32 & 66–69.)

681.142 **1652**

An Electronic Multiplier.—M. J. Somerville. (*Electronic Engng*, Feb. 1952, Vol. 24, No. 288, pp. 78–80.) Description of a multiplier circuit with negligible time lag for use in an analogue computer. The frequency of a carrier is modulated in proportion to one of the multiplicands, while its amplitude is modulated in proportion to the other. The resulting signal is fed to a phase discriminator, whose output is proportional to the required product. The use of the technique is illustrated in solving Airey's equation.

51 : 62 **1653**

Advanced Engineering Mathematics. [Book Review]—C. R. Wylie, Jr. Publishers: McGraw-Hill, New York, 1951, 640 pp., $7.50. (*Electronics*, Feb. 1952, Vol. 25, No. 2, p. 330.) ". . strongly recommended as text book or reference for advanced students in electrical engineering."

519.2 **1654**

Statistische Methoden für Naturwissenschafter, Mediziner und Ingenieure (Statistical Methods for Scientists, Physicians and Engineers). [Book Review]—A. Linder. Publishers: Birkhäuser, Basle, 2nd enlarged edn 1951, 238 pp., 31.20 Swiss francs. (*Z. angew. Math. Phys.*, 15th Nov. 1951, Vol. 2, No. 6, pp. 494–495.) ". . . will have a wide circulation and give excellent service in both theory and practice."

681.142 **1655**

The Preparation of Programs for an Electronic Digital Computer. [Book Review]—M. V. Wilkes, D. J. Wheeler & S. Gill. Publishers: Addison-Wesley Press, Cambridge, Mass., 1951, 167 pp., $5.00. (*J. Franklin Inst.*, Nov. 1951, Vol. 252, No. 5, pp. 445–446.) "Although the system of subroutines discussed and given in this book might not be applicable to all machines, it can serve as a pattern which will greatly facilitate the development of such a system for a particular machine."

681.142 **1656**

Synthesis of Electronic Computing and Control Circuits. [Book Review]—Staff of the Computation Laboratory, Harvard University. Publishers: Harvard University Press, Cambridge, Mass., 1951, 278 pp., $8.00. (*Electronics*, March 1952, Vol. 25, No. 3, pp. 400 . . 406.) "The book deals entirely with digital computing circuits, considering no analog devices. The control circuits mentioned in the title are of the type in which all of the relevant information is handled in digital form, rather than of the type associated with servomechanisms."

MEASUREMENTS AND TEST GEAR

538.71 **1657**

Some Developments in Electronic Magnetometers.—A. W. Brewer, J. Squires & H. McG. Ross. (*Elliott J.*, Sept. 1951, Vol. 1, No. 2, pp. 38–43.) Developments discussed include (*a*) airborne equipment suitable for geophysical surveys and capable of measuring variations as small as 1 gamma in the total geomagnetic field, (*b*) equipment for absolute measurements.

621.3.018.41 (083.74) **1658**

Standard-Frequency Transmissions.—(*Wireless Engr*, March 1952, Vol. 29, No. 342, p. 82.) Actual values for the frequencies of the standard-frequency transmissions from Rugby (1027 of April) and Droitwich (718 of March), as determined at the National Physical Laboratory, are to be reported regularly in *Wireless Engineer*. The first report, presented here, gives values for January 1952.

621.3.018.41(083.74) + 529.786] : 538.569.4 **1659**

'Atomic' Clocks and Frequency Stabilization on Microwave Spectral Lines.—C. H. Townes. (*J. appl.*

Phys., Nov. 1951, Vol. 22, No. 11, pp. 1365–1372.) "Application of the various types of radiofrequency spectral lines to accurate frequency stabilization and time standards is surveyed. Pertinent characteristics of microwave gas absorption lines and the various types of errors in frequency stabilization due to the nature of these absorption lines or to fundamental thermal noise are discussed in detail. It is shown that time standards synchronized with microwave absorption in ammonia or resonances in molecular or atomic beams have limits of accuracy of the order of 1 part in 10^{12} for a short time, and still smaller limiting fractional errors over longer periods of time."

621.317.18.083.4 **1660**

Balance Approach in A.C. Measurement Circuits: Part 1 — Theoretical Bases.—H. Poleck. (*Arch. tech. Messen*, Oct. 1951, No. 189, pp. T115–T116.) An ideal balancing operation is characterized by unambiguous indication of the direction in which the balancing elements must move, and by a single movement of the balancing elements to reach zero. Operation of null indicators with and without phase dependence is analysed.

621.317.3 : 621.385.032.216 **1661**

A Method of Measuring the Interface Resistance and Capacitance of Oxide Cathodes.—C. C. Eaglesfield & P. E. Douglas. (*Brit. J. appl. Phys.*, Nov. 1951, Vol. 2, No. 11, pp. 318–320.) The interface impedance, consisting of a resistance and a capacitance in parallel, causes frequency-dependent feedback. Another frequency-dependent network is added to make the gain independent of frequency, in which case the interface components are equal to the measurement components. The apparatus and its operation are described briefly.

621.317.328.029.62 **1662**

V.H.F. Microvoltmeter and Field-Strength Measurement Set.—P. Lygrisse. (*Électronique, Paris*, Nov. 1951, No. 60, pp. 29–31.) Circuit diagram and description of an instrument for field-strength measurements at levels between 5 μV/m and 0·1 V/m in the frequency range 75–195 Mc/s.

621.317.335.3.029.64 **1663**

Balance Methods for the Measurement of Permittivity in the Microwave Region.—T. J. Buchanan. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 61–66.) From measurements of propagation constant the permittivity of water and aqueous solutions were calculated for wavelengths of 3·2 cm and 1·26 cm.

621.317.335.3.029.64 **1664**

A New Method for Measurement of the Dielectric Constant of Low-Conductivity Fluids in the Centimetre Waveband.—E. Ledinegg, P. Urban & F. Reder. (*Acta phys. austriaca*, July 1950, Vol. 4, No. 1, pp. 9–17.) A method using a cylindrical cavity resonator is described; operation is at fixed frequency and at fixed cylinder length. The determination is made by measuring the volume of fluid introduced into the cavity to make it resonate at the same frequency as when empty. Accuracy to within a few parts per thousand is possible.

621.317.336 **1665**

Impedance Measurement at High Frequency using Bridged-T and Parallel-T Elements.—K. Lamberts. (*Arch. tech. Messen*, Oct. 1951, No. 189, pp. T108–T109.)

621.317.374 **1666**

Determination of Loss Angle of Materials with High Dielectric Constant.—E. Ledinegg & P. Urban. (*Acta phys. austriaca*, Dec. 1950, Vol. 4, Nos. 2/3, pp. 197–212.) The method is based on the introduction into a cylindrical cavity resonator of a layer of the dielectric material of thickness such that the resonance frequency is the same as for the empty cavity. See also 1664 above.

621.317.411.029.62/.63 **1667**

On the Determination of the Complex Permeability of Ferromagnetic Conductors at High Frequencies.—A. Wieberdink. (*Appl. sci. Res.*, 1950, Vol. B1, No. 6, pp. 439–452.) The complex propagation constant for e.m. waves in a concentric Lecher system, the outer conductor being copper, while the inner conductor is a wire of the material under investigation, is measured. From this constant the complex permeability of the wire can be calculated. Results of measurements on a Ni-Fe wire show a decrease of permeability with increasing frequency and a resonance phenomenon at a frequency of 320 Mc/s.

621.317.7 : 621.396.615.17 **1668**

The Multivibrator as Test Apparatus.—O. Limann. (*Funk u. Ton*, Nov. 1951, Vol. 5, No. 11, pp. 585–599.) Description of the circuit and its operation and of many applications in square-wave testing.

621.317.723 : 621.385.5 **1669**

A Stabilized Mains-Supplied Valve Electrometer Circuit.—G. Bonfiglioli & G. Montalenti. (*Alta Frequenza*, Oct. 1951, Vol. 20, No. 5, pp. 210–213.)

621.317.725 **1670**

Linear Diode Voltmeter.—R. E. Burgess. (*Wireless Engr*, March 1952, Vol. 29, No. 342, p. 80.) Correction to paper abstracted in 736 of March.

621.317.725 : 621.314.671 **1671**

Valve Voltmeter. The Rectifier Section.—M. G. Scroggie. (*Wireless World*, March 1952, Vol. 58, No. 3, pp. 89–94.) Detailed discussion of the design of a double-diode rectifier unit suitable for use in the measurement of alternating voltages with the d.c. instrument described in 1041 of April.

621.317.725 : 621.317.32 : 621.396.822 **1672**

Slideback and Infinite-Impedance Voltmeters.—R. E. Burgess. (*Wireless Engr*, March 1952, Vol. 29, No. 342, pp. 59–62.) Extension of analysis given previously for diode and anode-bend voltmeters (189 of January). In the slideback voltmeter the increase of mean current through a diode or triode on application of an input voltage is counterbalanced by additional negative bias; this type of voltmeter gives indications lower than the peak voltage of a c.w. signal and has a square-law response to noise. The infinite-impedance voltmeter has the same rectification characteristics as a simple diode voltmeter, but takes no power from the source. Curves are given for the rectification characteristics of the different types of voltmeter for c.w. signals and fluctuation noise applied separately; formulae are derived for the response to any arbitrary mixture of signal and noise.

621.317.73 **1673**

A Microwave Swept-Frequency Impedance Meter.—E. A. N. Whitehead. (*Elliott J.*, Sept. 1951, Vol. 1, No. 2, pp. 57–58.) Description of an instrument based on the use of directional couplers, for the rapid testing of waveguide components. A klystron oscillator has its frequency swept by a mechanical drive, the waveband of 3·0 to 3·4 cm being covered once in two seconds. The amplitude and phase of the voltage reflection coefficient of the component on test are displayed on a c.r.o.

621.317.733.011.21 : 621.392.26 **1674**

A Reflectionless Wave-Guide Termination.—R. E.

Grantham. (*Rev. sci. Instrum.*, Nov. 1951, Vol. 22, No. 11, pp. 828–834.) Developed as a reference standard for microwave impedance bridges, the termination, also applicable to coaxial transmission lines, consists of a section of waveguide with a movable dissipative load, and is preceded by a tuner which can be adjusted to cancel in magnitude and phase the small reflection coefficient of the load. Reflection coefficients of 0·001 were obtained with X-band waveguide terminations.

621.317.733.011.21 : 621.396.611.21 **1675**

The Design and Use of an Admittance Bridge for Piezoelectric Crystals.—J. F. W. Bell. (*Brit. J. appl. Phys.*, Nov. 1951, Vol. 2, No. 11, pp. 324–327.) A radio-frequency bridge for the rapid measurement of resistance and *Q*-factor of piezoelectric crystals is described. The limitation of the accuracy of measurement due to frequency fluctuations of the generator used and to variations in the stray capacitance of the variable resistance arm of the bridge is discussed. Examples of the use of the bridge at 250 kc/s are given.

621.317.737 **1676**

A *Q*-Meter based on Free Damped Oscillations.—K. Franz & S. F. Pinasco. (*Rev. telegr. Electronica, Buenos Aires*, Nov. 1951, Vol. 40, No. 470, pp. 731–733.) The meter is designed for determining accurately the *Q* value of the tank circuits of power oscillators, with *Q* values < 20. The circuit under test is introduced in the anode lead of a type-6SH7 pentode with pulsed input. The arrival of a pulse charges the tank-circuit capacitor, which discharges by free oscillation of the circuit; the voltage changes across the capacitor are applied to the grids of a double triode (the two sections connected in parallel) which is cathode coupled to a c.r.o. The *Q* of the tank circuit is given by $Q = n\pi/\log(B_0/B_n)$, where B_0 is the initial amplitude and B_n that of the *n*th wave in the damped wave train displayed on the c.r.o. Typical oscillograms corresponding to *Q* values ranging from 63 to $< 0{\cdot}5$ are shown.

621.317.755 : 621.385.012 **1677**

Electron-Tube Curve Generator.—M. L. Kuder. (*Electronics*, March 1952, Vol. 25, No. 3, pp. 118–124.) Description, with detailed circuit diagram, of equipment for c.r.o. display of families of anode characteristics, together with the locus of the load line and coordinates for direct measurement.

621.317.755 : 621.385.012 **1678**

Electron Tube Curve Tracer.—J. H. Kuykendall. (*Radio & Televis. News, Radio-Electronic Engng Section*, Aug. 1951, Vol. 46, No. 2, pp. 9–11, 29.) Description of c.r.o. equipment with direct-reading current voltage scales.

621.317.76.029.3 **1679**

Frequency Comparator.—P. Riéty. (*Ann. Télécommun.*, Nov. 1951, Vol. 6, No. 11, pp. 332–336.) Description of a circuit designed for the calibration of a.f. oscillators at frequencies between 20 c/s and 20 kc/s. A series of sub-harmonic frequencies is derived from a standard-frequency 1-kc/s oscillator by frequency division in a multivibrator circuit. The signal of frequency to be measured is made to beat with a suitable harmonic of one of the derived frequencies, a magic-eye or loudspeaker being used as indicator. Calibration points are available at the first 50 harmonics of nine frequencies ranging from 20 c/s to 1 kc/s. A circuit diagram and component values are shown.

621.317.784 **1680**

Power Meter and Mismatch Indicator.—R. G. Medhurst & J. A. Knudsen. (*Wireless Engr*, April 1952, Vol. 29, No. 343, p. 112.) A closed expression is given for an integral used by Boff (449 of February) in determining the sensitivity of a pickup loop.

621.396.615.11 : 534.844.1 **1681**

Equipment for Acoustic Measurements: Part 3—Acoustic Pulse Measurements.—C. G. Mayo, D. G. Beadle & W. Wharton. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, pp. 424–428.) A tone pulse is radiated into a studio under test, and the sound is picked up by a microphone connected to a triggered-timebase oscilloscope. The equipment weighs < 42 lb. Details are given of the circuits used. Triggering pulses at any predetermined interval between 0·5 and 30 sec are provided automatically by a transitron oscillator.

621.396.615.14 + [621.396.664 : 621.397.6 **1682**

Ultra High Instrumentation.—R. G. Peters. (*TV Engng, N.Y.*, Nov. 1951, Vol. 2, No. 11, pp. 18–20.) R.C.A. test and measurement equipment described includes a sweep marker generator and a picture monitor, r.f. load and wattmeter, and frequency and modulation monitors for television transmitters.

621.396.615.17 **1683**

Variable-Frequency Clock-Pulse Generator.—R. R. Rathbone. (*Radio & Televis. News, Radio-Electronic Engng Section*, Aug. 1951, Vol. 46, No. 2, pp. 19–20.) Description of a test unit developed at the Massachusetts Institute of Technology, providing 0·1-μs pulses with repetition frequencies from 0·2 to $4{\cdot}9 \times 10^6$/sec, the frequency being stable to within 20 parts in 10^6. The unit is used for building up complex circuits for pulse operation, and has standard 93-Ω input and output impedances.

621.396.615.17 : 621.397.6.001.4 **1684**

Linear Staircase Generator for Television Use.—A. M. Spooner & F. W. Nicholls. (*Electronic Engng*, Dec. 1951, Vol. 23, No. 286, pp. 481–482.) A cathode-coupled multivibrator is synchronized by line-suppression pulses; at the end of each pulse the multivibrator executes a train of oscillations, controllable in number between 5 and 30. This output is applied to a diode counter circuit, which generates a pulse with the corresponding number of steps. Such a waveform is useful for testing the response of television film recording apparatus.

621.396.933.2.001.4 **1685**

ILS Field Test Set.—C. L. Ellis. (*Radio & Televis. News, Radio-Electronic Engng Section*, Nov. 1951, Vol. 46, No. 5, pp. 3–5 . . 26.) Description of the G-250A set for checking the entire aircraft instrument-landing equipment. It is of rugged weatherproof construction and can be operated by non-technical personnel. Three independent crystal-controlled generators provide the marker, localizer and glide-slope frequencies, a choice of 20 being available in the localizer and glide-slope bands. A cycling unit controls the signal modulation sequence and is designed for either automatic or manual operation.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

532.137 : 621.395.61/.62 **1686**

Vibrating-Plate Viscometer.—J. G. Woodward. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 98–100.) The viscous damping exerted on a plate immersed in a liquid and oscillating in its own plane is measured by an electromechanical transducer.

534.321.9 : 620.179.1 **1687**

Ultrasonic Tyre-Testing Equipment.—(*Engineer, Lond.*, 2nd Nov. 1951, Vol. 192, No. 4997, pp. 565–566.)

A nondestructive production test developed jointly by the Dunlop and General Electric companies is based on the fact that an internal discontinuity such as imperfect bonding between rubber and fabric gives rise to an air film which reflects ultrasonic waves almost completely. The tyres are immersed in water during test. The generator used is a quartz crystal operating at 50 kc/s and located 1 in. from the rubber, in the well of the tyre; the power output is about 1 W. See also *Elect. Rev., Lond.*, 2nd Nov. 1951, Vol. 149, No. 3858, pp. 893–894.

621-52 : 621.396.611.3 **1688**

Applications of Electrical Methods of Differentiation to Control Problems.—J. Rateau. (*Rev. gén. Élect.*, Nov. 1951, Vol. 60, No. 11, pp. 451–465.) Analysis of basic differentiating circuits and discussion of their design and application in control mechanisms such as an automatic pilot.

621-526 **1689**

An Electronic Servo Simulator for Unstable and Open Loop Systems.—N. T. van der Walt. (*Electronic Engng*, Feb. 1952, Vol. 24, No. 288, pp. 52–57.) The simulator consists of a number of feedback amplifiers in cascade; its response to a square-wave input is viewed on a c.r.o. Inclusion of a clamping arrangement, operative during the timebase flyback, ensures a stable return to datum level.

621-57 : 537.228.1 **1690**

High-Speed Crystal Clutch.—(*J. Franklin Inst.*, Nov. 1951, Vol. 252, No. 5, pp. 427–428.) In a clutch developed by Codier at the National Bureau of Standards, primarily for use in high-speed computers, application of direct voltage to the electrodes of three 'bimorph' piezoelectric elements causes them to bend and press the clutch output disk against the rotating input disk.

621.317.083.7 **1691**

Simultaneous A.M. and F.M. in Rocket Telemetering.—W. C. Moore. (*Electronics*, March 1952, Vol. 25, No. 3, pp. 102–105.) Details of receiver and pressurized transmitter operating on a single carrier frequency of 183 Mc/s and providing two channels, one an a.m. channel 5 Mc/s wide suitable for rapidly varying phenomena, unusual waveforms, etc., the other a f.m. channel using a reactance modulator. The a.m. is effected by screen-grid modulation of the final stage of the transmitter.

621.365.54† **1692**

Induction Heating in the Drop-Forging Industry.—G. W. Seulen. (*Metal Treatm.*, Nov. 1951, Vol. 18, No. 74, pp. 483–489.) Various types of medium-frequency (i.e., up to 10 kc/s) generator equipment are discussed and details are given of some German-designed plants.

621.383.001.8 : 535.61-15 : 778.37 **1693**

The Application of Image Converters to High Speed Photography.—J. A. Jenkins & R. A. Chippendale. (*J. Brit. Instn Radio Engrs*, Nov. 1951, Vol. 11, No. 11, pp. 505–517.) Description of the use of a new type of tube, the Mullard ME 1201, with a Cs-Sb cathode of very low resistance and average sensitivity of 20 μA/lumen to tungsten light at 2 700°K, as a high-speed optical shutter. Exposure times shorter than 10^{-7} sec can be obtained. Numerous test photographs are reproduced.

621.384.6 **1694**

The Helix as a Linear Accelerator for Protons.—D. R. Chick & D. P. R. Petrie. (*Nature, Lond.*, 3rd Nov. 1951, Vol. 168, No. 4279, pp. 782–783.) A circular waveguide consisting of a closely wound wire helix has been proposed for accelerating protons up to about 20 MeV. To counteract the radial component of electric field due to the wave, which tends to disperse the proton beam, a hollow beam of electrons surrounding the proton beam is suggested.

621.385.833 **1695**

Ion Image of an Emissive Anode.—G. Couchet, M. Gauzit & A. Septier. (*C. R. Acad. Sci., Paris*, 5th Nov. 1951, Vol. 233, No. 19, pp. 1087–1090.) Report of observations, made with an emission-type electron microscope, of the positive-ion emission from a plane anode heated by electron bombardment. The fluorescent screen is of $CaWO_4$.

621.385.833 **1696**

The Axial Potential of [electron] Lenses with Grids.—M. Bernard. (*C. R. Acad. Sci., Paris*, 23rd July 1951, Vol. 233, No. 4, pp. 298–299.) A rigorous expression is derived for the axial potential of a lens consisting of two symmetrical cylindrical or plane equipotential elements with circular holes, and an interposed plane grid at a potential different from that of the other two elements.

621.385.833 **1697**

Gaussian Elements of [electron] Lenses with Grids.—M. Bernard. (*C. R. Acad. Sci., Paris*, 26th Nov. 1951, Vol. 233, No. 22, pp. 1354–1356.) Calculations from the formulae previously derived (1696 above) gave results in good agreement with the experimental observations of Knoll & Weichart (4542 of 1938). Both convergent and divergent lenses are treated.

621.385.833 **1698**

Rigorous Calculation of Typical Electrostatic Electron Lenses.—W. Glaser & H. Robl. (*Z. angew. Math. Phys.*, 15th Nov. 1951, Vol. 2, No. 6, pp. 444–469.) Paraxial electron trajectories are determined for configurations approximating to cylindrical lenses.

621.385.833 **1699**

Newton's Law of the Formation of Images applied to Electron Optics.—P. Funk. (*Acta phys. austriaca*, Dec. 1950, Vol. 4, Nos. 2/3, pp. 304–308.) A simpler solution than those of Hutter (1004 of 1946) and Glaser & Lammel (202 of 1942) is presented.

621.385.833 **1700**

High-Resolution Velocity Analysis with Magnetic Electron Lenses.—F. Lenz. (*Naturwissenschaften*, Nov. 1951, Vol. 38, No. 22, pp. 524–525.) A method is described for testing lens h.v. stability and investigating electron velocity losses by observations on a ring-focus line.

621.385.833 **1701**

Imaging Properties of a Series of Magnetic Electron Lenses.—G. Liebmann & E. M. Grad. (*Proc. phys. Soc.*, 1st Nov. 1951, Vol. 64, No. 383B, pp. 956–971.) An investigation of the dependence of the paraxial image-forming properties and the first-order lens aberrations on geometrical design and lens excitation. The range of gap widths, S, investigated was $S/D = 0{\cdot}2$ to $S/D = 2$, where D = lens diameter. The field distributions within the lenses were measured; the lens data derived from the measurements are given in a series of graphs applicable to the most common magnetic electron lenses. Application of the results to lens design is discussed.

621.385.833 **1702**

Technique of Electron Microscopy.—D. G. Drummond & G. Liebmann. (*Nature, Lond.*, 10th Nov. 1951, Vol. 168, No. 4280, pp. 819–821.) A report of the proceedings at a conference of the Electron Microscopy Group of the Institute of Physics held in St. Andrews University, 19th–21st June 1951, with brief indications of the subject matter of the various papers read.

621.385.833 **1703**

The Symmetrical Magnetic Electron-Microscope Objective Lens with Lowest Spherical Aberration.—G. Liebmann. (*Proc. phys. Soc.*, 1st Nov. 1951, Vol. 64, No. 383B, pp. 972–977.) In the symmetrical magnetic lenses already considered (1702 above) there is an optimum value of the lens excitation parameter which produces minimum spherical aberration. The relation between maximum obtainable axial field strength and the maximum field strength in the pole-piece gap leads to an optimum design for electron microscope objectives giving minimum spherical aberration and maximum resolving power. A modification of this optimum design for practical use is described.

621.387.4 **1704**

Self-Quenching Counters containing Small Amounts of Polyatomic Constituent.—A. D. Krumbein. (*Rev. sci. Instrum.*, Nov. 1951, Vol. 22, No. 11, pp. 821–827.)

621.395.625.3 **1705**

Ferrography.—R. B. Atkinson & S. G. Ellis. (*J. Franklin Inst.*, Nov. 1951, Vol. 252, No. 5, pp. 373–381.) 'Ferrography' is the name given to a new magnetic process, here described, for recording graphic information and reproducing it on paper in visual form. A scanning process similar to that used in facsimile produces an electrical signal which is fed to a magnetic recording head, and a record is made on a film coated with iron oxide. Printed reproductions are made using magnetic inks, either black or coloured. The record can be used repeatedly and stored indefinitely.

621.398 : 621.396.712 : 621.396.619.13 **1706**

Remote-Control System for F.M. Broadcast Stations.—P. Whitney. (*Tele-Tech*, Aug. & Sept. 1951, Vol. 10, Nos. 8 & 9, pp. 32–35 & 44–45, 80.) An effective remote-control system has enabled the WRFL transmitter, installed on a mountain peak, to be controlled from equipment in the studios at Winchester, Va., more than 20 miles away. No operators have been in regular attendance at the transmitter since April 1951. A general description is given of the equipment, with detailed circuit diagrams of the control oscillators, band-pass amplifier, automatic protective circuits and telemetry equipment. The protective equipment was found very necessary owing to frequent interruption of the transmitter power supply due to line surges caused by electrical storms, which operated the main circuit breaker. A different type of breaker, together with a recycling device operating 3–5 times within a few seconds, eliminated this difficulty.

621-52 **1707**

Fundamentals of Automatic Control. [Book Review]—G. H. Farrington. Publishers: Chapman & Hall, London, 1951, 285 pp., 30s. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, p. 453.) Gives a thorough analysis of the fundamentals of process control, but does not deal in detail with servomechanisms as such.

PROPAGATION OF WAVES

538.566 **1708**

The Jumps of Discontinuous Solutions of the Wave Equation.—H. Bremmer. (*Commun. pure appl. Math.*, Nov. 1951, Vol. 4, No. 4, pp. 419–426.) There is a correspondence between discontinuous changes in an e.m. field, with their associated wavefronts, and the amplitudes of the geometrical-optical approximations which correspond, under steady-state conditions, to ray trajectories orthogonal to the wavefronts. The theory of this correspondence is developed in terms of Dirac's impulse function for any discontinuity connected with the scalar wave equation. Modifications to the theory required in the solution of vector problems related to the application of Maxwell's equations are given.

538.566 **1709**

The Transport of Discontinuities in an Electromagnetic Field.—E. T. Copson. (*Commun. pure appl. Math.*, Nov. 1951, Vol. 4, No. 4, pp. 427–433.) Generalized solutions of the vector wave equations are obtained and the transport equations are derived, the medium being assumed isotropic, with variable dielectric constant and permeability.

538.566.029.64 **1710**

Asymptotic Solutions of a Differential Equation in the Theory of Microwave Propagation.—R. E. Langer. (*Commun. pure appl. Math.*, Dec. 1950, Vol. 3, No. 4, pp. 427–438.) "The purpose of this paper is to show that asymptotic formulas for the solutions of a differential equation that is central to the theory of microwave propagation may be readily derived from results that are available in the mathematical literature." The problem of determining the normal modes of propagation in an atmosphere in which the refractive index varies only with the height is first briefly reviewed. The differential equation under conditions applying respectively to the 'leaky' and 'transitional' modes of propagation is discussed. In each case the results are compared with the analogous results of Pekeris (2211 of 1947), obtained by power-series methods.

621.396.11 **1711**

The Propagation of E.M. Waves from Land to Sea and vice versa: Part 1.—P. Höller. (*Z. angew. Phys.*, Nov. 1951, Vol. 3, No. 11, pp. 424–432.) An analytical approximation method of investigation is used in which the wave equation is first satisfied while neglecting the boundary conditions, and the boundary conditions are then satisfied while neglecting the wave equation. It is assumed that (*a*) the earth is flat, (*b*) sea and land are individually homogeneous and the boundary is sharp, (*c*) the sea is an ideal conductor, (*d*) the complex refractive index of land is not very large, so that the simple Weyl solution is applicable, (*e*) the transmitter height is sufficient for refraction and reflection at the earth's surface to be calculated by geometrical optics, (*f*) the distances of both transmitter and receiver from the coast are large compared with the wavelength.

621.396.11 **1712**

Propagation of Very-High-Frequency Radio Waves.—E. H. Jones. (*Nature, Lond.*, 17th Nov. 1951, Vol. 168, No. 4281, pp. 870–871.) Measurements are reported of the strength of the field at a height of 90 ft due to an airborne transmitter at a height of 40 000 ft; operating frequencies of 280·2 and 386·6 Mc/s were used. Observed values are plotted against transmitter/receiver distance and compared with values calculated from conventional ray theory; the agreement is generally better when the earth's radius is taken at its actual value rather than at its four-thirds value. Where the indirect ray was reflected from the sea the observed minima were deeper than the calculated values, the difference corresponding to an apparent increase in reflection coefficient by a factor as great as 2·25 in some cases.

621.396.11 : 551.510.535 **1713**

Ionosphere Review: 1951. Greatly Reduced Rate of Decrease in Sunspot Activity and M.U.Fs.—Bennington. (See 1605.)

621.396.11 : 551.510.535 **1714**

R.F. Time-Delay Measurements.—D. Davidson: R.

Naismith & E. N. Bramley. (*Wireless Engr*, April 1952, Vol. 29, No. 343, pp. 111–112.) Comment on 473 of February and authors' reply.

621.396.81 : 523.72 **1715**

Solar Activity and Ionospheric Effects.—R. E. Burgess & C. S. Fowler. (*Wireless Engr*, Feb. 1952, Vol. 29, No. 341, pp. 46–50.) Ionospheric disturbances on short and long waves were investigated by recording the field strengths of two stations on frequencies of 18·89 Mc/s and 191 kc/s respectively. These results were compared with solar activity in the form of flares and sunspots and with simultaneous recordings of solar noise on frequencies of 30, 42, 73 and 155 Mc/s. The times of commencement of disturbances on short and on long waves and of solar flares were all coincident and varied by up to 3 min from the noise bursts which often preceded the other phenomena. 86% of the noise bursts and 50% of the observed flares occurred without any accompanying ionospheric effect.

621.396.812.029.64 **1716**

Attenuation of Radio Signals caused by Scattering.—J. B. Smyth & C. P. Hubbard: A. H. LaGrone. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1386–1387.) Comment on 2525 of 1951 and reply by one of the authors.

621.396.812.3.029.64 **1717**

An Experimental Study of Fading in Propagation at 3-cm Wavelength over a Sea Path.—D. G. Kiely & W. R. Carter. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 53–60.) The observations were made between July 1950 and January 1951 over an optical path of 10·6 nautical miles, using horizontal polarization. Associated meteorological observations were made at points not far from the transmission path. Typical signal records are shown and the data analysed in terms of the monthly mean level. No reliable prediction of fading could be made from the simple meteorological observations. The reduction in radar range due to fading reached a maximum in July and August, being up to 20% for 90% of the time. The power level of radar beacons for operation up to the radio horizon is estimated.

RECEPTION

621.396.621 : 621.396.822 **1718**

On Determining the Presence of Signals in Noise.—I. L. Davies. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 45–51.) A theoretical treatment using the concept of existence probability, which is evaluated for a signal whose form is known precisely and for a modulated carrier where only the carrier phase is unknown. The efficiency of any receiver can be derived by comparing existence probabilities at input and output; the practical implications of this are discussed. When applied to the case of a radar signal, the theory shows that the filter giving the optimum range information will also yield the greatest existence information.

621.396.621.54 **1719**

Calculation of the Sensitivity of Decimetre- and Centimetre-Wave Receivers using Diodes or Detectors as Mixers.—H. Behling. (*Arch. elekt. Übertragung*, Nov. & Dec. 1951, Vol. 5, Nos. 11 & 12, pp. 489–498 & 561–564.) Using the concept of equivalent noise resistance, formulae are developed for calculating the critical sensitivity (corresponding to unity signal/noise ratio); from these and the characteristics of the mixing circuit (mixer valve, first i.f. valve, i.f. circuit, etc.) it is possible to calculate the oscillator power required to obtain unity signal/noise ratio at the output of the i.f. amplifier. This calculation is made for a crystal mixer, and the influence on the sensitivity of the mixer and i.f. parameters is examined.

STATIONS AND COMMUNICATION SYSTEMS

517.512.2 : 621.39.001.11 : 621.396.67.012.71 **1720**

Fourier Analysis and Negative Frequencies.—M. L. Tetes. (*Wireless Engr*, March 1952, Vol. 29, No. 342, p. 80.) Discussion on 1422 of May (Shaw).

621.317.35 : 621.3.015.7 **1721**

Analysis of Non-Recurrent Pulse Groups.—L. S. Schwartz & N. P. Salz. (*Radio & Televis. News, Radio-Electronic Engng Section*, Nov. 1951, Vol. 46, No. 5, pp. 8–10.) Graphical determination of the resultant frequency spectrum of non-recurrent groups of regularly spaced pulses such as occur in teletype and p.c.m. transmissions.

621.317.35 : 621.39.001.11 **1722**

Signals of Given Duration and Minimum Spectral Width.—K. Fränz. (*Arch. elekt. Übertragung*, Nov. 1951, Vol. 5, No. 11, pp. 515–516.) Using Fourier integrals, an equation is derived whose solution gives the form of signal, for a given duration, for which the energy concentration within a given frequency band has its maximum value. The spectral concentration of energy in a rectangular pulse is very near the maximum. See also 2376 of 1951.

621.39.001.11 **1723**

The Concept of Information and Transmission Capacity in Communication Technique.—H. Weber. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, Nov. 1951, Vol. 29, No. 11, pp. 401–406. In German.) Discussion of a coding system for German text, making use of Shannon's theory.

621.39.001.11 : 519.251.6 **1724**

Information Theory and Inverse Probability in Telecommunication.—P. M. Woodward & I. L. Davies. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 37–44.) "The foundations of information theory are presented as an extension of the theory of inverse probability. By postulating that information is additive and taking suitable averages, all the essential definitions of Shannon's theory for discrete and continuous communication channels, with and without noise, are obtained. The theory is based on the idea that receiving a communication, or making an observation, merely changes the relative probabilities of the various possible messages. The whole process of reception can, therefore, be regarded as a means of evaluating *a posteriori* probabilities, and this leads to the idea that the optimum receiver in any telecommunication problem can always be specified, in principle, by inverse probability. The simplest instance is the correlation receiver for detecting very weak signals in the presence of noise, and its theory is briefly discussed."

621.395.44 **1725**

Carrier-Frequency Systems free from Linear Distortion.—J. Peters. (*Arch. elekt. Übertragung*, Nov. 1951, Vol. 5, No. 11, pp. 509–515.) Theory of the modulation and demodulation processes in a s.s.b. system is considered; rigorous conditions for freedom from distortion are derived by using the Laplace transformation. A network can be found which will transmit the envelope free from distortion for a sine-voltage input initiated at any phase over the part of the cycle when the voltage is increasing; for initiation at other parts of the cycle the output contains an additional decaying direct component. Distortion-free transmission can still be obtained in this case by a tandem arrangement of circuits designed according to the analysis given.

621.395.44 : 621.395.97 1726

New Control and Measurement Equipment for the Broadcasting [-network] Repeater Stations of the German Post Office.—E. A. Pavel & H. Liersch. (*Fernmeldetech. Z.*, Nov. 1951, Vol. 4, No. 11, pp. 513–518.) Description of station equipment for line tests, control switching and programme monitoring, comprising the test rack type 48 and associated loudspeakers. See also 1238 of May (Pavel et al.).

621.396 : 361.1 1727

Radio Communications in the Australian Flying Doctor Service.—L. N. Schultz. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1951, Vol. 12, No. 10, pp. 300–302.) Description of a system which provides subscribers distant more than 25 miles from a telephone with transceiver facilities. Three operating frequencies, of about 2, 4 and 7 Mc/s, are allotted to each of the eight bases together with the associated outposts, the actual frequencies being different for each group. Base-transmitter power ranges from 20 W to 300 W. In addition to normal communication receivers, bases have a night-alarm receiver. Outpost transmitters have an output of 3 W and are powered by vibrators, except for a few old stations with pedal-driven generators. Outposts use three fixed frequencies together with a variable tuning band for h.f. reception.

621.396.44 : 621.315.052.63] + 621.317.083.7 1728

Single Sideband Transmission and its Multiple Utilization for Carrier-Current Channels on High-Voltage Power Lines.—A. de Quervain. (*Brown Boveri Rev.*, July/Aug. 1951, Vol. 38, Nos. 7/8, pp. 208–219.) See also 801 of March (Bloch).

621.396.5 : 621.396.931 1729

450-Mc/s Mobile Radio Service.—N. E. Wunderlich. (*FM-TV*, Nov. 1951, Vol. 11, No. 11, pp. 22–25 . . 38.) Description of equipment for a dispatch system for at least 1 000 taxis operating in Chicago. Five pairs of channels 100 kc/s wide are used. Eight 15-W or 100-W ph.m. transmissions cover eight zones each roughly 5 miles square. The mobile transmitter-receiver units have a frequency stability of ± 5 parts in 10^6 and can be operated on any of four switch-selected frequencies. Communication on 452 Mc/s is superior to that on 162 Mc/s in built-up areas; the average noise level is 10 db lower and no ignition interference is experienced, while the walls of buildings apparently serve as waveguides, thus reducing attenuation.

621.396.65 1730

New Pennsylvania Turnpike U.H.F. Communications System.—D. N. Lapp. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 84–87.) Description of a system operating over a 327-mile route between the outskirts of Philadelphia and the Ohio border, with 13 intermediate stations. Frequencies of 953 Mc/s and 960 Mc/s are used along the main route, and frequencies in the band 152–162 Mc/s are used for local communication networks connected to the relay stations.

621.396.65 : 621.387.4 1731

Use of a Radio Link for studying Coincidences between Pulses from Counters separated by Large Distances.—E. Picard, A. Rogozinski & M. Surdin. (*J. Phys. Radium*, Nov. 1951, Vol. 12, No. 9, pp. 854–857.) Equipment is described for a link operating on 10 kMc/s, with a range of about 20 km. When there is no line-of-sight path between the two stations a frequency of 5 Mc/s is used.

621.396.65.029.63 : 621.316.726 1732

1 400-Mc/s Radiophone.—J. B. L. Foot. (*Wireless World*, April 1952, Vol. 58, No. 4, pp. 132–135.) An experimental radio link is described which uses a DET28 disk-seal triode valve operated to give an output of about 1 W. The range is up to about 30 miles with a signal/noise ratio of about 20 db. Each transmitter oscillator acts also as local oscillator for the double-superheterodyne receiver, a frequency difference equal to the first i.f. being maintained between the oscillators at the two stations. A 'master and slave' system is used to keep this frequency difference constant in spite of oscillator frequency drift, the frequency of the slave station being controlled by an electromechanical method in accordance with the frequency of the signal received from the master station.

621.396.7.029.62 + 621.396.621] : 621.396.619.13 1733

F.M. in Germany.—(*Wireless World*, April 1952, Vol. 58, No. 4, pp. 141–144.) Factors which have influenced the development of the f.m. v.h.f. broadcasting network in Western Germany are discussed. The disposition of transmitting stations is shown. Some details are given of simple f.m. receivers which dispense with the use of limiter and discriminator.

621.396.712 : 621.396.619.13 : 621.398 1784

Remote-Control System for F.M. Broadcast Stations.—Whitney. (See 1706.)

621.396.712.029.55 1735

The Allouis-Issoudun H.F. Group of Radiodiffusion Française.—A. Gaillard. (*Onde élect.*, Nov. 1951, Vol. 31, No. 296, pp. 420–433.) Description of equipment for the s.w. broadcasting service which in 1952 will include thirteen 100-kW transmitters, grouped at two centres. At Allouis one 100/130-kW transmitter operating in the 31, 41 and 49 m bands can be switched to any one of six dipoles oriented in N-S and E-W directions; two other units provide four simultaneous transmissions in the bands from 13 to 49 m from a system of 12 directive rhombic aerials. At Issoudun 12 simultaneous transmissions can be made. Features dealt with include the services now operating, station lay-out, power supply and valve cooling, circuit arrangement and valves, and transmission characteristics. A map and chart show the world-wide coverage achieved.

621.396.712(489) 1736

Broadcasting Installations for the Two Programmes in Denmark.—(*Teleteknik, Copenhagen*, Oct. 1951, Vol. 2, No. 3, pp. 207–249.)

The Planning of the New Broadcasting Stations.—G. Pedersen.

Propagation Conditions for the V.H.F. Range.—B. Nielsen.

The Broadcasting Station at Skive.—G. Bramslev.

The 100-kW Medium-Wave Transmitter at Kalundborg.—G. Bramslev.

Transmitters working on the International Shared Frequencies and New Stations for F.M.—P. Christensen.

Directive Aerials in Herstedvester.—J. Hansen.

Construction and Erection of Aerial Masts.—I. G. Hannemann & B. J. Rambøll.

621.396.931 : 621.395.635 1737

Selective Calling applied to Mobile Radio.—W. T. Muscio. (*Proc. Instn Radio Engrs, Aust.*, Oct. 1951, Vol. 12, No. 10, pp. 303–311.) Detailed description of the Selecto-Call system. The fixed station has a choice of 11 frequencies in the range 154–449 c/s for a.m. of a 7-kc/s subcarrier, the main carrier being frequency modulated. The mobile receiver includes a decoder comprising up to four reeds tuned to different transmitter tones; actuation of the reeds in a given sequence is required to produce the calling signal.

SUBSIDIARY APPARATUS

621-526 **1738**

Stabilization of Direct-Current Servomechanisms.—M. Cambornac & F. Lajeunesse. (*Onde élect.*, Nov. 1951, Vol. 31, No. 296, pp. 434–445.) Methods of improving stability are discussed. These include the use of phase-correcting networks and the coupling of an auxiliary dynamo to the servo motor. Circuits and characteristics of different low-power models are given, the response times of which range between 0·01 and 0·2 sec.

621-526 : 621.396.93 **1739**

Analysis and Construction of a Position-Fixing Servo-mechanism.—G. Klein. (*Ann. Télécommun.*, Nov. 1951, Vol. 6, No. 11, pp. 313–324.) The mechanism is designed to operate with the direction discriminator [895 of 1950 (Loeb et al.)] so that a direct reading of azimuth is obtained automatically. A theoretical analysis defines the specifications for the system.

621.311.6 **1740**

Carrier-Type Regulated Power.—J. Houle. (*Radio & Televis. News, Radio-Electronic Engng Section*, Nov. 1951, Vol. 46, No. 5, pp. 14–15 . . 31.) A continuously variable d.c. output from 0 to 300 V with regulation to within 0·1 V is obtained, using the following principle. An oscillator feeds a small signal to an a.c. amplifier. In series with the amplifier input is a Ge diode connected so that the d.c. control signal varies the diode bias current. At the amplifier output, the d.c. is recovered by means of a rectifier.

621.311.6 **1741**

Regulated 1 600-Ampere Filament Supply.—A. W. Vance & C. C. Shumard. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 122–123.) Description of circuit for the 6-V filaments of some 4 000 valves used in the Project Typhoon analogue-digital computer of the U.S. Navy.

621.311.6 : 621.316.722.1 **1742**

A Stabilized A.C. Supply for Lamps and Valve Heaters.—J. C. S. Richards. (*J. sci. Instrum.*, Nov. 1951, Vol. 28, No. 11, pp. 333–335.) Description of a system providing up to 150 W of a.c. power with the r.m.s. voltage stable to within $\pm$ 0·1%. A saturated diode is used as reference element and a saturable choke as control element.

621.311.62.078.3 **1743**

A High-Stability High-Voltage Power-Supply Unit.—J. Templeton. (*N.Z. J. Sci. Tech. B*, Nov. 1951, Vol. 33, No. 3, pp. 218–223.) Description of a mains-operated unit supplying 1 mA at 3 kV steady to within $\pm$ 1 part in 10^4 over periods of 30 min, and capable of modification to give greater output.

621.314.6 **1744**

Rectification of Alternating Currents with High Modulation.—J. Böhse. (*Arch. elekt. Übertragung*, Aug. 1951, Vol. 5, No. 8, pp. 363–376.) From an approximate equation for the static characteristic of rectifiers with essentially constant differential slope, equations for the demodulation products (direct-current and modulation-frequency) are derived for the case of high modulation, when the rectifier operates like a valve. The operational parameters, currents, voltages and effective resistances on the input and output sides, as well as the demodulation distortion, are represented by transcendental functions of the phase angle of the current, so that numerical values of the electrical quantities can be taken directly from tables and nomograms, whose use is explained by several examples.

621.314.6.012.6 **1745**

Exact Analysis of the Linear Rectifier Circuit: Part 1—Half-Wave Rectification with Capacitive Smoothing.—H. Niehrs. (*Frequenz*, Oct. 1951, Vol. 5, No. 10, pp. 273–279.) Formulae are derived for the amplitude and phase of the harmonics and the effective output voltage of a half-wave rectifier with sine-wave input, assuming negative half cycles to be completely blocked and internal resistance of the rectifier to be constant.

621.314.653 **1746**

The Time Lag of an Ignitron.—N. Warmoltz. (*Philips Res. Rep.*, Oct. 1951, Vol. 6, No. 5, pp. 388–400.) Measurements were made of the time lag for igniters of widely different resistance, using (*a*) liquid, (*b*) solid Hg or Sn cathodes. The effect of the gas pressure in the tube was also investigated. The results favour the thermal theory of Mierdel.

621.316.722.1 **1747**

An Electronic Voltage Stabilizer with Self-Regulated Heater Supply.—C. Morton. (*Electronic Engng*, Feb. 1952, Vol. 24, No. 288, p. 65.) The heaters of the valve cathodes are connected in series with the load, thus constituting part of the output circuit across which the stabilized voltage is developed.

621.316.722.1 **1748**

Voltage Stabilization: Demands and Methods.—A. J. Maddock. (*J. sci. Instrum.*, Nov. 1951, Vol. 28, No. 11, pp. 325–333.) Typical cases are discussed in which stabilized supplies are required. The principal types of stabilizer are described and details are given of their performance and their voltage and power ratings. Mains generator control is not considered.

TELEVISION AND PHOTOTELEGRAPHY

621.397.335 : 535.623 **1749**

N.T.S.C. Color-TV Synchronizing Signal.—R. B. Dome. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 96–97.) Discussion of the synchronizing signal required in the field tests on the band-sharing system (1750 below), in which a local oscillation synchronized to the frequency of the colour subcarrier is used for demodulating the colour signal at the receiver. The signal chosen is a train of about 10 cycles of the colour subcarrier frequency, timed to occur about midway between the end of the horizontal synchronizing pulse and the end of the blanking pulse.

621.397.5 : 535.623 **1750**

Principles of N.T.S.C. Compatible Color Television.—C. J. Hirsch, W. F. Bailey & B. D. Loughlin. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 88–95.) Discussion of the specifications formulated by the U.S. National Television System Committee to govern field tests on the system in which a separate colour signal is transmitted simultaneously with a monochrome signal on a separate subcarrier within the 4-Mc/s channel carrying the monochrome signal [826 of March (Loughlin)]. Factors affecting the choice of the colour-subcarrier frequency and the modulation system are considered.

621.397.6 : 621.396.664] + 621.396.615.14 **1751**

Ultra High Instrumentation.—Peters. (See 1682.)

621.397.611/.621].2 **1752**

Scanning-Current Linearization by Negative Feedback.—A. W. Keen. (*J. Televis. Soc.*, Oct./Dec. 1951, Vol. 6, No. 8, pp. 308–315.) "An introductory qualitative survey of the known methods of applying negative feedback to the problem of linearizing the output current of scanning systems needed in television transmitting and receiving equipment. Practical circuit developments are divided into two categories according as the fundamental process, termed 'current integration', is separable into two component operations, viz., voltage integration and voltage-to-current conversion, which are carried out consecutively by separate feedback systems connected in cascade, or performed by a single system subject to overall feedback."

621.397.611.2 **1753**

Paraxial Image Formation in the 'Magnetic' Image Iconoscope.—J. C. Francken & R. Dorrestein. (*Philips Res. Rep.*, Oct. 1951, Vol. 6, No. 5, pp. 323–346.) Describes a method of computing electron trajectories near the axis of a system of e.s. and e.m. fields having rotational symmetry. Numerical results are given for a case which approximates to conditions in the 'magnetic' image iconoscope. The mechanism of image formation is discussed; it differs considerably from that in ordinary magnetic lenses.

621.397.611.2 **1754**

The Image Iconoscope, a Camera Tube for Television.—P. Schagen, H. Bruining & J. C. Francken. (*Philips tech. Rev.*, Nov. 1951, Vol. 13, No. 5, pp. 119–133.) The different types of television camera tube are discussed and a detailed description is given of the Philips image iconoscope Type 5854. The Cs-Sb-O photocathode gives an output of about 45 μA/lumen when illuminated by an incandescent lamp with colour temperature 2 600°K. Magnetic focusing and deflection are used for the scanning beam, the resolution being 900–1 000 lines at the middle of the target and as high as 700 lines at the edges. The mica target is 25μ thick and has a thin coating of MgO to increase secondary emission. See also 1452 of May (Francken).

621.397.611.2 **1755**

Television Camera Tube.—R. Barthélemy. (*Onde élect.*, Nov. 1951, Vol. 31, No. 296, pp. 415–419.) Theory previously given (1499 of 1951) is discussed with reference to (*a*) the use of a thin-film target with its potential adjusted by an auxiliary electron stream, (*b*) the optimum thickness of the self-polarized target film. A few performance details are given of the supericonoscope with e.s. deflection and self-polarized target.

621.397.62 **1756**

Basic Circuit Description of a R.C.A. Television Receiver.—(*Radiotronics*, Oct. & Nov. 1951, Vol. 16, Nos. 10 & 11, pp. 211–224 & 228–243.) Description of the R.C.A. Victor 630TS receiver, with full circuit details and analysis of the functions of the various circuits.

621.397.62 **1757**

Wide Angle Deflection Yokes.—H. E. Thomas. (*Radio & Televis. News, Radio-Electronic Engng Section*, Sept. 1951, Vol. 46, No. 3, pp. 3–6.) A general discussion of factors to be considered in designing systems with deflection angles up to 90° and with low distortion. The length of the yoke is determined as a compromise between the requirements for high sensitivity and those for avoidance of neck shadow. Auxiliary magnetic devices for eliminating neck shadow are described. Methods of obtaining an optimum relation between spot distortion and pattern distortion are indicated.

621.397.62 : [535.623 + 535.61-29 **1758**

C.B.S.-Columbia — First Commercial Color plus Black-and-White Set.—I. J. Melman, E. S. White & S. Cuker. (*Radio-Electronics*, Nov. 1951, Vol. 23, No. 2, pp. 24–27.) Description of a receiver for 525-line black-and-white or 405-line full-colour pictures. Normal circuits are used together with a colour scanning disk with silent motor drive and associated disk-control circuit.

621.397.62 : 621.396.662 **1759**

Concentric-Lines tune U.H.F. Channels.—E. E. Harries & M. Cawein. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 108–112.) Circuit and construction details are given of a converter for use with a conventional v.h.f. television receiver; a three-section tuner of Inductuner type is used, and the frequency range 470–890 Mc/s is covered. The circuit consists of preselector, crystal mixer and oscillator, followed by an i.f. stage. The noise figure of the converter is discussed.

621.397.62 : 621.398 **1760**

Remote Controls for TV promote Viewer Comfort.—R. F. Scott. (*Radio-Electronics*, Nov. 1951, Vol. 23, No. 2, pp. 28–31.) Description of electromechanical and electronic systems incorporated in various television receivers which permit control from a convenient viewing point some distance from the set.

621.397.621 : 621.316.721.078.3 **1761**

Stabilizing Vertical-Deflection Amplifiers.—W. B. Whalley, C. Masucci & K. Hillman. (*Electronics*, March 1952, Vol. 25, No. 3, pp. 116–117.) Application of inverse feedback to the vertical-deflection amplifier makes vertical linearity and picture-height stability practically independent of valve transconductance.

621.397.621 : 621.317.35 **1762**

Television Picture Line Selector.—J. Fisher. (*Electronics*, March 1952, Vol. 25, No. 3, pp. 140–143.) Description of equipment enabling examination of the video waveform in a single selected scanning line. The oscilloscope is triggered with a single horizontal-synchronization pulse which precedes the line to be observed. Application to measurement of frequency response and transient response of television cameras and picture-generating devices, such as the monoscope and flying-spot scanner, is described.

621.397.621.2 **1763**

Evaluating Performance of TV Picture Tubes.—J. Green. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 124–129.) Methods and apparatus are described for accurately determining the vertical and horizontal dimensions of the spot, using special rasters; the significance of the measurements in relation to the design of the various parts of the c.r. tube is discussed.

621.397.621.2 : 621.385.832 **1764**

TV Picture Tubes with Iron Envelopes.—Szegho & Pohl. (See 1790.)

621.397.8 **1765**

Television Ghosts. Effect of Multi-path Propagation in Hilly Country.—J. A. Hutton. (*Wireless World*, March 1952, Vol. 58, No. 3, pp. 84–88.) An investigation of reception with 12 different aerials in hilly country round the Holme Moss transmitter. A standing-wave field due to reflection from hills two or three miles away may give rise to positive or negative ghost images according to the time delay and the modulation amplitude. The ghost image is least apparent when the aerial is between a node and antinode of the standing-wave field. The double-H aerial system consisting of two H aerials $\lambda/2$ apart, with $\lambda/4$ spacing of the dipoles, was found the best of the aerials tested.

621.397.822 : 621.397.2 **1766**

Random Noise. Rate of Occurrence of Peaks.—V. J. Francis. (*Wireless Engr*, Feb. 1952, Vol. 29, No. 341, pp. 37–40.) The number of peaks occurring above various amplitude levels on a noise trace is calculated and the results are compared with those of laboratory experiments on a system simulating the London-Birmingham television radio relay link. The results are related to the perceptible level of noise peaks on actual television pictures.

621.397.822.1 **1767**

Observer Reaction to Video Crosstalk.—A. D. Fowler. (*J. Soc. Mot. Pict. Televis. Engrs*, Nov. 1951, Vol. 57, No. 5, pp. 416–424.)

TRANSMISSION

621.396.615.16.029.55 : 621.396.933 **1768**

Brown Boveri Transmitters in the Service of Civil Aviation.—A. Vincre. (*Brown Boveri Rev.*, July/Aug. 1951, Vol. 38, Nos. 7/8, pp. 220–226.) S.w. transmitters installed for the French Ministry of Transport and Public Services meteorological service and for Air France overseas communications are described.

VALVES AND THERMIONICS

537.533.8 **1769**

The Angular Distribution of the Secondary Electrons of Nickel.—J. L. H. Jonker. (*Philips Res. Rep.*, Oct. 1951, Vol. 6, No. 5, pp. 372–387.) Description of the measurement tube, and graphical presentation of results obtained for the distribution of the secondary electrons as a function of the angle of incidence and the voltage of the primary electrons.

621.383 : 546.482.21 **1770**

Photoelectric Cells using Activated Cadmium Sulphide.—P. Goercke. (*Ann. Télécommun.*, Nov. 1951, Vol. 6, No. 11, pp. 325–331.) The photoelectric properties of CdS may be enhanced by addition of traces of Cu or Ag. The influence of this impurity content on spectral sensitivity, electrical resistance, and noise figure is studied.

621.383.4 : 546.817.231 **1771**

The Long-Wave Limit of Infra-Red Photoconductivity in PbSe.—A. F. Gibson, W. D. Lawson & T. S. Moss. (*Proc. phys. Soc.*, 1st Nov. 1951, Vol. 64, No. 383A, pp. 1054–1055.) Measurements of the photoconductivity of a PbSe photo-diode, consisting of a *p*-type crystal with a tungsten whisker, show that the long-wave limit, defined as 50% decrease from the maximum sensitivity, was $4{\cdot}7\mu$ at room temperature and $6{\cdot}8\mu$ at 90°K. Another cell with a periclase window had a long-wave limit of $8{\cdot}1\mu$ at 20°K.

621.384.5 : 621.316.722 **1772**

The Characteristics of some Miniature High-Stability Glow-Discharge Voltage-Regulator Tubes.—F. A. Benson. (*J. sci. Instrum.*, Nov. 1951, Vol. 28, No. 11, pp. 339–341.) Three types of tube were studied. The results of short- and long-term tests to determine striking-voltage and running-voltage variations are presented. Values are also given for temperature coefficient of running voltage and for the magnitudes and durations of the initial drifts. See also 3159 of 1951.

621.385 **1773**

The Measurement of Microphony in Valves.—R. Bird. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, pp. 429–431.) Arrangements are described for investigating microphony in a valve by using it as the input valve of a.f. amplifier and locating it in the sound field of a loudspeaker fed by the amplifier. Optical and electrical methods of detecting the vibrating elements are discussed.

621.385-71 **1774**

Electron-Tube Heat-Transfer Data.—B. O. Buckland. (*Elect. Engng, N.Y.*, Nov. 1951, Vol. 70, No. 11, pp. 962–966.) Essentials of 1951 A.I.E.E. Summer General Meeting paper. The method of calculating heat flow by means of equivalent electrical circuits is considered. The effects of temperature differences and cooling-fin shape on radiation- and air-cooled valves are discussed and design considerations for water-cooled and forced-air-cooled valves are summarized. Graphs are given from which the data necessary for the design of cooling systems may be determined.

621.385 : 537.525.92 **1775**

Two-Way Space-Charge Flow with Plane Electrodes.—K. Müller-Lübeck. (*Z. angew. Phys.*, Nov. 1951, Vol. 3, No. 11, pp. 409–415.) The potential in the space-charge field in the presence of electrons and positive ions was found graphically by Langmuir (1929 Abstracts, p. 511); a rigorous analytical solution is now derived for this potential.

621.385.004.15 **1776**

The Technique of Trustworthy Valves.—E. G. Rowe. (*J. Brit. Instn Radio Engrs*, Nov. 1951, Vol. 11, No. 11, pp. 525–540. Discussion, pp. 540–543.) A survey of progress in the design, manufacture and testing of radio valves to ensure high reliability. Failures occurring in manufacture and during the subsequent life of the valve are discussed, and the need for increased cooperation between user and manufacturer is stressed.

621.385.004.15 **1777**

A Survey of Quality and Reliability Standards in Electronic Valves for Service Equipment.—G. L. Hunt. (*J. Brit. Instn Radio Engrs*, Nov. 1951, Vol. 11, No. 11, pp. 519–524. Discussion, pp. 540–543.) Conditions of use of valves in Service equipment are described and also methods adopted for ensuring supplies of satisfactory valves.

621.385.012 : 621.317.755 **1778**

Electron Tube Curve Tracer.—Kuykendall. (See 1678.)

621.385.029.6 : 538.311 : 621.318.423 : 513.647.1 **1789**

Properties of the Electromagnetic Field of Helices.—Roubine. (See 1580.)

621.385.029.63/.64 **1780**

Equivalent Temperature of an Electron Beam.—M. E. Hines: P. Parzen. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1385–1386.) Comment on 2580 of 1951 and reply by one of the authors.

621.385.032.216 **1781**

The Life of Oxide Cathodes in Modern Receiving Valves.—G. H. Metson, S. Wagener, M. F. Holmes & M. R. Child. (*Proc. Instn elect. Engrs*, Part III, March 1952, Vol. 99, No. 58, pp. 69–81. Discussion, pp. 82–87.) A summary of existing information, with some results of original research. If mechanical faults, the effect of gas on cathode emission, and excessive interface feedback can all be avoided, valve life is probably limited only by evaporation of the activated oxide. 24 references.

621.385.032.216 : 539.16 **1782**

Contribution to the Study of Electronic Tubes by the Use of Radioactive Elements.—J. Debiesse & G. Neyret. (*Le Vide*, Nov. 1951, Vol. 6, No. 36, pp. 1098–1102.) Radioactive isotopes are used for getter and cathode materials to facilitate the investigation of the distribution and migration of the materials and the origin of the electron emission. Radioactivity of the metal base in general reduces the thermionic emission from the cathode. See also 2301 of 1951 (Debiesse et al.).

621.385.2 **1783**

Influence of Initial Velocities on Electron Transit Times in Diodes.—J. T. Wallmark. (*Phys. Rev.*, 1st Nov. 1951, Vol. 84, No. 3, p. 598.) Barut (2057 of 1951) has described a method for calculating the transit time of electrons in diodes with partial space-charge, assuming a uniform initial-velocity distribution. A first-order perturbation method has been applied to the calculation of transit times for electrons with nonuniform initial-velocity distribution. Curves are given showing the spread in transit time (*a*) for a diode with anode voltage 100 V and anode-cathode distance of 5 mm, as a function of current density in the diode for a difference in initial velocity of 0·1 V (roughly corresponding to conditions for an oxide cathode at 1 100°K), (*b*) for a reflected beam under the same conditions, the spread in transit time being in this case much reduced. A complete report is in preparation.

621.385.2 **1784**

The Transit-Time Effect in a Cylindrical Diode.—Way Dong Woo. (*J. appl. Phys.*, Nov. 1951, Vol. 22, No. 11, pp. 1333–1339.) The impedance offered to a small superposed alternating current by a space-charge-limited cylindrical diode is expressed in terms of the variational anode conductance, the transit-time angle, and a group of constants which are functions of the ratio of the anode and cathode radii. The result covers the range of transit-time angle up to π radians and of the ratio of the anode and cathode radii from unity up to 100.

621.385.2 : 546.289] + 621.314.7 **1785**

Germanium Crystal Valves.—B. R. A. Bettridge. (*Electronic Engng*, Nov. 1951, Vol. 23, No. 285, pp. 414–417.) Characteristics of Ge diodes are outlined, and television circuit applications described. Ge triodes are mentioned briefly.

621.385.2 : 621.318.572 **1786**

R.F. Bursts actuate Gas-Tube Switch.—H. J. Geisler. (*Electronics*, Feb. 1952, Vol. 25, No. 2, pp. 104–105.) Description of technique for using simple gas-filled diodes as switches in storage and other circuits of electronic computers. Pulses of r.f. voltage applied to external metal bands round the diode envelopes cause the diodes to strike at reduced d.c. voltage. Examples of circuit applications are given.

621.385.5.032.212 : [621.318.572 + 681.142 **1787**

The Single-Pulse Dekatron.—J. R. Acton. (*Electronic Engng*, Feb. 1952, Vol. 24, No. 288, pp. 48–51.) Description of a new type of gas-filled cold-cathode counting valve which differs from the earlier dekatrons [2066 of 1950 (Bacon & Pollard)] in requiring only a single input pulse to move the cathode glow on a complete step. The tentative specification is given for the GC10D development valve, which is reliable for pulse rates up to 20 000 per sec. Input and output circuits are discussed in relation to the nature of the pulses dealt with.

621.385.832 **1788**

Cathode-Ray Tubes. A Review of Progress.—L. F. Broadway. (*Proc. Instn elect. Engrs*, Part I, Nov. 1951, Vol. 98, No. 114, pp. 316–320.)

621.385.832 : 621.318.572 **1789**

New Electronic Tubes employed as Switches in Communication Engineering: Part 2—Switch Tubes.—J. L. H. Jonker & Z. van Gelder. (*Philips tech. Rev.*, Oct. 1951, Vol. 13, No. 4, pp. 82–89.) Experimental multicontact valves are described in which the 'contacts' are effected by means of secondary emission, the primary electron beam being directed electrostatically on to the various secondary-emission elements. The use of a ribbon-shaped primary beam permits currents of several milliamperes with voltages of 200 to 300 V, and valve dimensions can be kept small. Part 1: 1173 of April.

621.385.832 : 621.397.621.2 **1790**

TV Picture Tubes with Iron Envelopes.—C. S. Szegho & R. G. Pohl. (*TV Engng, N.Y.*, Nov. 1951, Vol. 2, No. 11, pp. 8–9 . . 27.) An envelope with an iron cone and Cr/Fe-alloy beads for sealing the cone to the glass parts is cheaper to make than one with a Cr/Fe-alloy cone. As a further development the Cr-Fe-alloy beads were eliminated and screen glasses were developed suitable for sealing to the iron cones, using an intermediate glaze at the sealing area; technique for this operation is described. The requirements for the neck glass are also discussed.

621.396.615.141.2 : 537.533.8 **1791**

Influence of Secondary Emission on the Oscillation Process in Whole-Anode Magnetrons.—F. W. Gundlach & K. Schörken. (*Z. angew. Phys.*, Nov. 1951, Vol. 3, No. 11, pp. 416–424.) Measurements are reported of the cathode resistance, the back-heating current and the h.f. voltage of the oscillating magnetron as functions of anode voltage. Oscillation is maintained even when the external cathode-heater circuit is completely cut off, the cathode temperature being then such that no appreciable thermionic emission can occur. Analysis of the results leads to the conclusion that the magnetron is maintained by secondary emission.

621.385 + 621.396.6 **1792**

Introduction to Electronic Circuits. [Book Review]—R. Feinberg. Publishers: Longmans, Green & Co., London, 163 pp., 18s. (*Wireless Engr*, April 1952, Vol. 29, No. 343, p. 113.) "The treatment is, on the whole, satisfactory although some will find it rather compressed."

MISCELLANEOUS

061.3 : 621.396.029.63 **1793**

Report on the First I.R.E. U.H.F. Symposium.—B. M. Ely. (*TV Engng, N.Y.*, Oct. 1951, Vol. 2, No. 10, pp. 10–13, 28.) Subjects dealt with at this symposium, held in Philadelphia, included transmission tests at 850 Mc/s, apparatus for frequency and impedance measurements, a side-fire helical transmitting aerial, and the design of u.h.f. receivers.

6 : 061.4 **1794**

British Instrument Industries Exhibition — London, 1951.—M. W. Thring. (*J. sci. Instrum.*, Oct. 1951, Vol. 28, No. 10, pp. 293–300.) The development of the scientific instrument industry in Britain is reviewed, and some of the more important exhibits at the first exhibition are discussed.

621.396 **1795**

Radio Handbook. [Book Review]—Publishers: Editors and Engineers Ltd, Santa Barbara, California, 13th edn, 736 pp., \$6.00. (*Electronics*, Dec. 1951, Vol. 24, No. 12, pp. 322 . . 326.) "Progressive and thorough in its coverage of the newest and most helpful developments."

621.396 **1796**

Radio Amateur's Handbook. [Book Review]—Publishers: American Radio Relay League, West Hartford, Conn., 20th edn, 618 pp., \$3.00. (*Electronics*, Dec. 1951, Vol. 24, No. 12, pp. 322 . . 326.) "Contains what is still probably the most complete listing of communication type tubes to be found anywhere."

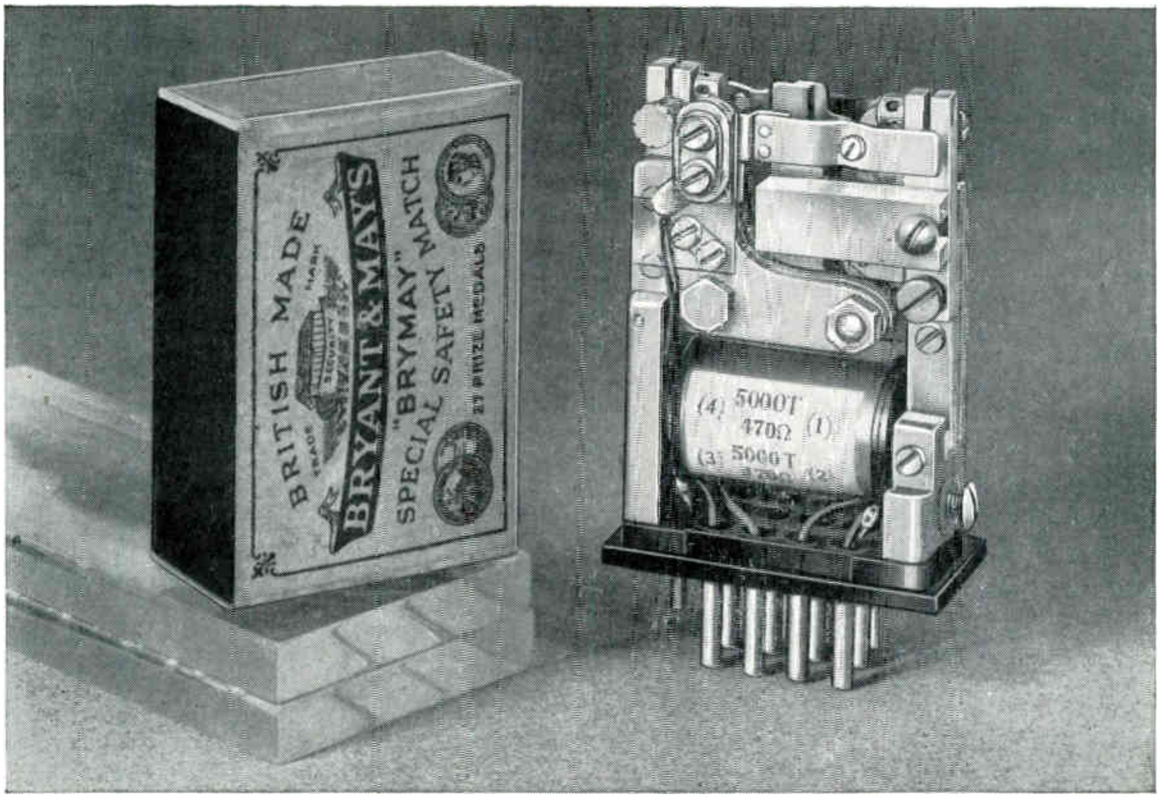
EDISWAN

CLIX

United for better Radio, Television and Electronic component service

By adding the Clix range of radio, television and electronic components to its existing list of products, The Edison Swan Electric Co., Ltd. is able to offer an improved components service to the Radio Industry. Future enquiries and orders for these products, and others in the Ediswan range, will be welcomed.

THE EDISON SWAN ELECTRIC COMPANY LIMITED
155 Charing Cross Road, London, W.C.2, and branches
Member of the A.E.I. Group of Companies



SMALL BUT VERY GOOD!

Considering it is only the size of a matchbox the Type 5 Carpenter Polarized Relay is capable of a surprisingly high performance providing answers to problems in many fields of electrical engineering. Outstanding features of the Type 5 include:

HIGH OPERATIONAL SPEED • FREEDOM FROM CONTACT REBOUND
IMMUNITY FROM POSITIONAL ERROR • GOOD CONTACT PRESSURES
HIGH SENSITIVITY • ACCURACY OF SIGNAL REPETITION
RUGGED DESIGN • EXCEPTIONAL THERMAL STABILITY

Plug or solder tag base optional.

Dimensions—(With cover. Excluding connecting pins.) $2\frac{3}{16}$ *ins. high.* $1\frac{7}{16}$ *ins. wide.* $\frac{3}{4}$ *in. deep. Weight (including socket)* 4.8 *oz.* (137 *gm.*)

Complete specification and further details of the complete range of Carpenter Relays may be had on request.

TMC CARPENTER POLARIZED RELAY TYPE 5

Manufactured by the Sole Licensees:

TELEPHONE MANUFACTURING CO. LTD

Contractors to Governments of the British Commonwealth and other Nations.

HOLLINGSWORTH WORKS · DULWICH · LONDON S.E.21 *Telephone: Gipsy Hill 2211 (10 lines)*

APPOINTMENTS—The engagement of persons answering these advertisements must be made through the local office of the Ministry of Labour and National Service, etc., if the applicant is a man aged 18-64 or a woman 18-59 inclusive, unless he or she or the employer is excepted from the provisions of The Notification of Vacancies Order, 1952.

Murphy Radio Ltd., wish to appoint a chief radio and television receiver designer to take charge of their electrical design laboratory. The successful applicant must possess good academic qualifications, preferably with some research experience and followed by several years in an industrial organisation engaged in television or allied fields.

The post is pensionable and the salary paid will be commensurate with the high standard of qualifications required. Applications, which will be treated in strict confidence, should be addressed to the Director of Engineering, Murphy Radio, Ltd., Welwyn Garden City, Herts.

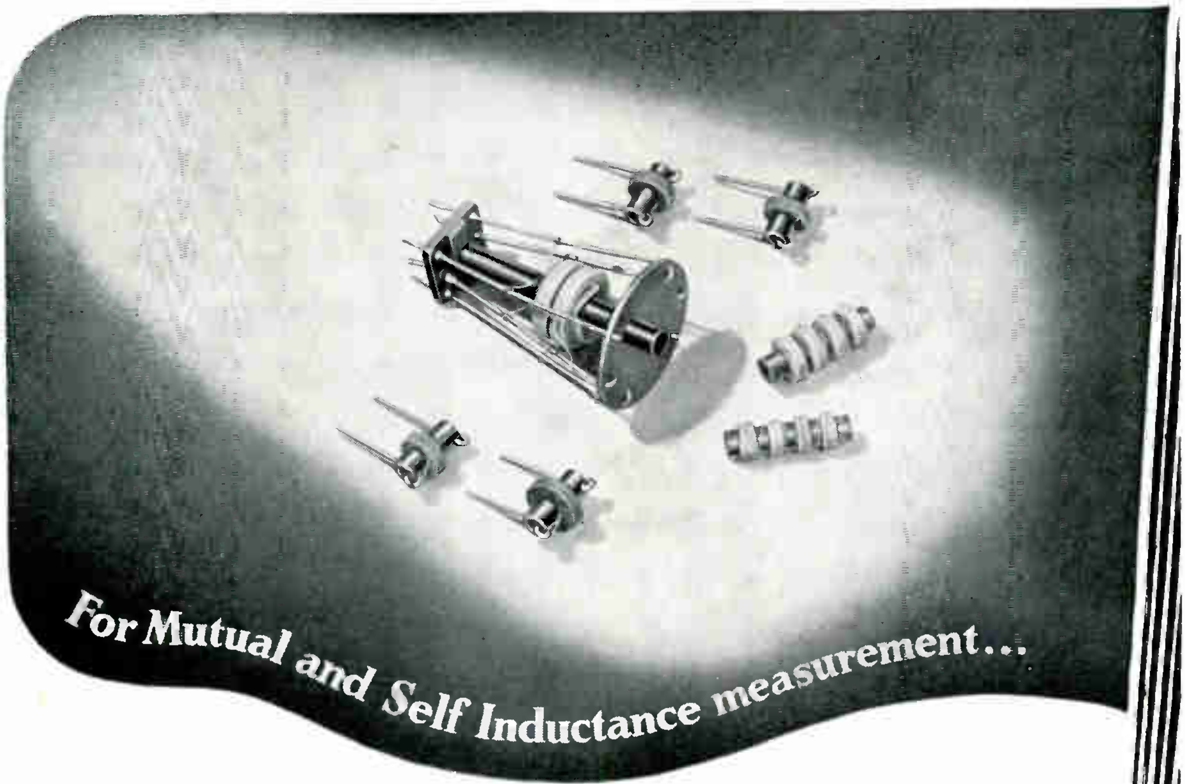
Cottage Laboratories Ltd., require the following staff:—

(*a*) Project leader to control a division of electronic and electro-mechanical engineers. The applicant should be capable of leading a team of research and development engineers in a highly technical field. Starting salary will be according to ability, experience and age.

(*b*) Senior electronic and electro-mechanical physicists or engineers. A physical and experimental outlook is desirable. The applicants must be capable of designing modern electronic circuits and of leading a group of assistant engineers. Starting salary according to age and experience.

(*c*) Assistant Engineers interested in electronics and small mechanisms.

There are prospects of rapid promotion in an expanding laboratory. Five-day week, excellent canteen and active sports club. Apply stating age, nationality, previous experience and salary required to the Manager, Cottage Laboratories Ltd., Portsmouth Road, Cobham, Surrey.


For Mutual and Self Inductance measurement...


CINTEL
MUTUAL AND SELF INDUCTANCE BRIDGE

THE WORLD'S GREATEST BOOKSHOP

FOYLES

• FOR BOOKS •

New and secondhand Books on every subject.
Stock of over 3 million volumes.

119-125, CHARING CROSS ROAD, LONDON, W.C.2

Gerrard 5660 (16 lines) ★ *OPEN 9—6 (inc. Sat.)*

Engineer of degree standard, B.Sc., engineering desirable, required for design and development of power capacitors. Should have knowledge of di-electrics and measurements. Apply Box 0009, c/o *Wireless Engineer*.

Development Engineer with knowledge of metallurgy and electronics required for experimental work on ultrasonics in Research Department of scientific instrument makers in N.E. London area. Age 25/35.—Write stating experience and salary required to Box 0010, c/o *Wireless Engineer*.

Specialist required with expert knowledge of development and production of carbon potentiometer tracks.—Apply Chief Engineer, Box 0011, c/o *Wireless Engineer*.

S. Smith & Sons (England), Ltd., Bishops Cleeve, Cheltenham, require the following staff:—

Senior Engineer—to lead a section in the design and development of small servo systems including the application of gyroscopic and magnetic amplifier technique. This is a permanent post and carries superannuation benefits, and housing assistance will be given to successful applicant. Commencing salary £900 to £1,200 dependent on age, qualifications and experience. Ref. 4/EN/G.

Senior Mechanical Engineer—to control design and development of a wide range of Aircraft Instruments and allied equipment. Applicants should have a sound technical education with previous design experience in pressure-sensitive elements and intricate mechanisms. Salary £900 to £1,200 depending upon age and experience. Ref. 5/EN/M.

Development Engineers—for development of high-grade test equipment in connection with manufacture of automatic pilots and aircraft instruments. Work involves application of electronic techniques over a frequency range from zero to approximately 100 kc. p.s. together with light electrical and mechanical engineering. Preference will be given to applicants with experience of applied measurements in one or more of the above engineering fields but the foremost qualification is a sound appreciation of fundamental engineering principles. Salary according to qualifications and experience. Ref. 1/EN

Development Engineers—senior and junior engineers experienced in the design of electronic test equipment including valve voltmeters, C.R.O., oscillators, etc. Applicants will be responsible for the development of research models to the production stage and must have a good fundamental knowledge of electrical theory and practical experience of design. Salary dependent on qualifications and experience. Ref. 2/EN.

Development Engineers—for work on auto control and servo systems. Applicants with experience in low frequency electronic techniques, including the use of magnetic amplifiers. Preference given to those with previous experience in designing equipment for aviation requirements. Salary according to qualifications and experience. Ref. 3/EN.

Write quoting Reference Numbers and giving qualifications and experience to the Personnel Manager.

Belling & Lee, Ltd., Cambridge Arterial Road, Enfield, Middlesex, require research assistants in connection with work on electronic components, fuses, interference suppressors and television aerials. Applicants must be graduates of the I.E.E. or possess equivalent qualifications together with similar laboratory experience. Salary will be commensurate with previous experience; five-day week, contributory pension scheme. Applications must be detailed and concise, and will be treated as confidential.

Ferranti, Ltd., have immediate vacancies for men with electrical engineering qualifications to undertake the advanced testing of Naval Anti-Aircraft Fire Control Equipment involving electronics and servo mechanisms either in firms workshops or on board H.M. Ships in Home Ports. Salary in accordance with age and experience between £356 and £650 per annum. Normal expenses plus a generous allowance are paid when working out.

Previous experience of this type of work, though desirable, is not essential. Forms of application from Mr. R. J. Hebbert, Staff Manager, Ferranti, Ltd., Hollinwood, Lancs. Please quote ref. H.G.N.

The General Electric Co., Ltd., Brown's Lane, Coventry, have vacancies for Development Engineers, Senior Development Engineers, Mechanical and Electronic, for their Development Laboratories on work of national importance. Fields include microwave and pulse applications. Salary range £400-£1,250 per annum. Vacancies also exist for Specialist Engineers in component design, valve applications, electro-mechanical devices and small mechanisms. The Company's Laboratories provide excellent working conditions with social and welfare facilities. Superannuation Scheme. Assistance with housing in special cases. Apply by letter stating age and experience to The Personnel Manager (Ref. CHC.).

One or two experienced radar or electronic engineers, also engineers experienced in servo mechanisms are wanted for a Guided Weapons project by English Electric Laboratory, Luton. Progressive post with a starting salary of £600 to £1,000 per annum according to experience. Write, giving full details quoting Ref. 456H, to Central Personnel Services, English Electric Co., Ltd. 24/30 Gillingham Street, London, S.W.1

Post Graduate and Final Year university students in physics, electrical and mechanical engineering and metallurgy are invited to send details of their records to the Staff Manager (Ref. GBLC/S/876), Research Laboratories of The General Electric Co., Ltd., Wembley, Middlesex. A number of openings in interesting experimental research will be available during the coming months for men with outstanding ability and qualifications.

Engineers and physicists with experience in the following fields are required by the G.E.C. at the Stanmore Laboratories: (*a*) servo-mechanisms, (*b*) microwave aerials, transmitters or receivers, (*c*) small mechanisms, (*d*) electronic circuitry, (*e*) D.C. amplifiers, (*f*) test gear for field trials. Preference will be given to men with good academic qualifications. Applications should be sent to the Staff Manager (Ref. GBLC/875), G.E.C. Research Laboratories, Wembley, Middlesex.

Two graduates, one senior, one junior, required for work in Electronics Section of laboratory engaged in Physics Research. The function of the Electronics Section is primarily to design and build control and measuring gear for the laboratory, including Sections working on Electron Microscopy, Nuclear Physics, Semi-Conductors and Physical Metallurgy.

Two other posts are vacant for work on new project requiring men with Ph.D. or honours degree and at least two years' post-graduate experience and having interest in work on radio frequency power circuits.

Applications in writing stating age, qualifications and experience to Personnel Officer, Associated Electrical Industries Limited, Research Laboratory, Aldermaston Court, Aldermaston, Berkshire.

Scientific Computer required to assist with design, development and research work in connection with electrical networks. Minimum qualifications: Inter degree or Higher School Certificate including Mathematics. Experience in Computing, Mathematics, Physics or Engineering desirable. Initiative and interest in Mathematics and numerical work essential. Good salary and conditions of service. Apply Personnel Department, Telephone Manufacturing Co., Ltd., Sevenoaks Way, St. Mary Cray, Kent.

G.E.C.

QUARTZ CRYSTAL ACTIVITY TEST SET

UNSURPASSED AS A SIMPLE & ACCURATE INSTRUMENT

FOR THE MEASUREMENT OF CRYSTAL PERFORMANCE

The G.E.C. Quartz Crystal Activity Test Set measures the equivalent parallel resistance of a quartz crystal when oscillating in a circuit having an input capacity of either 20 pF, 30 pF or 50 pF, the alternative capacities being selected by a switch.

The dial is calibrated and has a range of 4 kilohms to 130 kilohms and is direct reading. No calculation is necessary. Measurements can be made at any convenient amplitude of oscillation up to 10V. R.M.S. at the crystal terminals for crystals of normal activity.

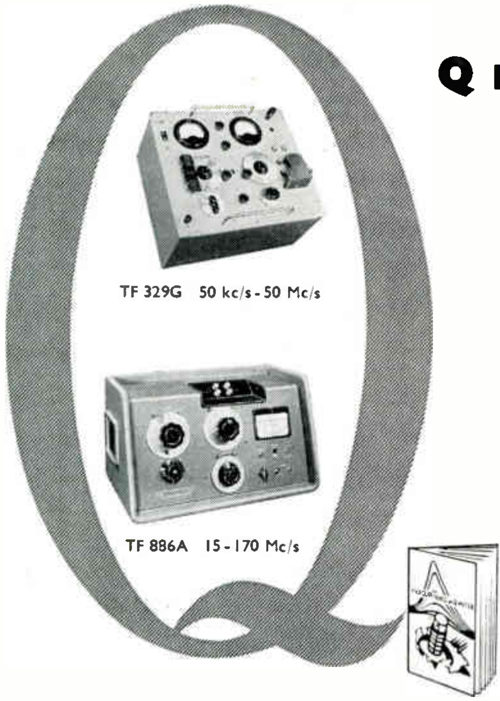
The accuracy of the loss dial calibration is $\pm$ 2%.

WRITE FOR DESCRIPTIVE LEAFLET—

SALFORD ELECTRICAL INSTRUMENTS LTD.

PEEL WORKS : SILK STREET : SALFORD : LANCS.

Subsidiary of THE GENERAL ELECTRIC CO. LTD. OF ENGLAND



Index to Advertisers

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.

Electrolytics for chassis mounting

Since the introduction of these chassis mounting condensers, thousands of Amplifiers and P.A. equipments have been designed incorporating them.

The surge limiting feature of Wet Electrolytics is now of less interest owing to the greater load currents in common use today, but their indestructibility and freedom from possibility of short circuit are advantages which are often sought.

Dry Electrolytics have other advantages; greater capacity for a given size, larger ripple rating, ability to work at higher temperatures and adaptability for mounting in any position.

WET ELECTROLYTIC CONDENSERS

CAP. in µF.	PEAK WORKING VOLTS	SURGE VOLTS	BODY		TYPE NUMBER	LIST PRICE EACH
			L'gth	Dia.		
4	440	460	3	1⅜	812	8/6
8	440	460	4½	1⅜	802	10/-
8	500	525	4½	1⅜	805	12/-
16	440	460	4½	1⅜	802	12/-
32	320	400	4½	1⅜	809	12/-

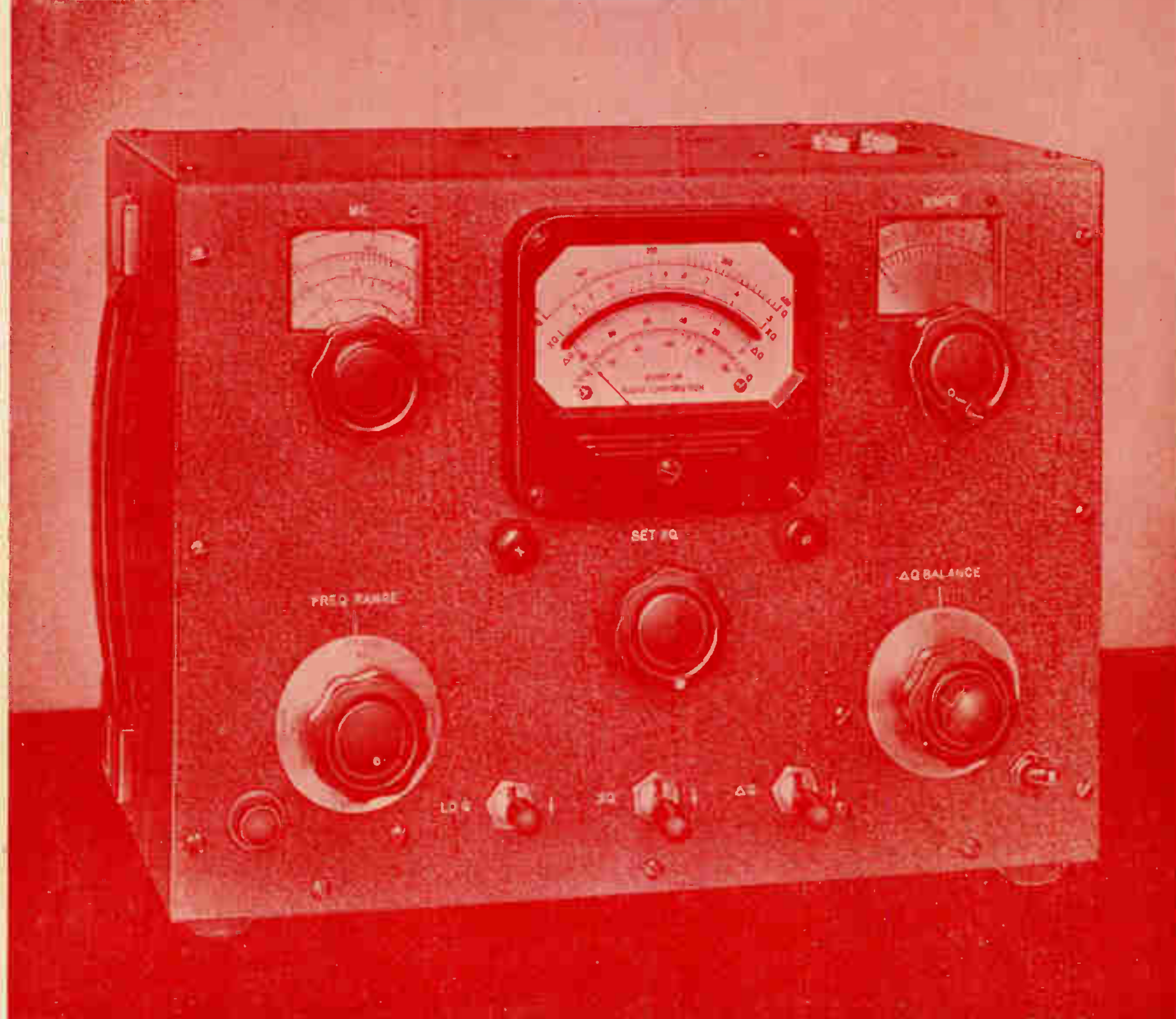
DRY ELECTROLYTIC CONDENSERS

CAP. in µF.	PEAK WORKING VOLTS	SURGE VOLTS	BODY		TYPE NUMBER	LIST PRICE EACH
			L'gth	Dia.		
32	350	400	2⅞	1	312	9/-
4	500	600	2⅞	1	512	7/-
8	500	600	4½	1	512	8/-
16	500	600	4½	1½	512	11/6
32	500	600	4½	1½	512	17/6
8	600	700	4½	1½	922	15/-

All types of Wets and Drys can be supplied with insulating washers to isolate negative can from chassis where necessary.

THE TELEGRAPH CONDENSER CO. LTD

RADIO DIVISION: **NORTH ACTON · LONDON · W.3 · Tel: ACORN 0061**


MC
SET ΔQ
ΔQ BALANCE
FREQ RANGE