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In this issue

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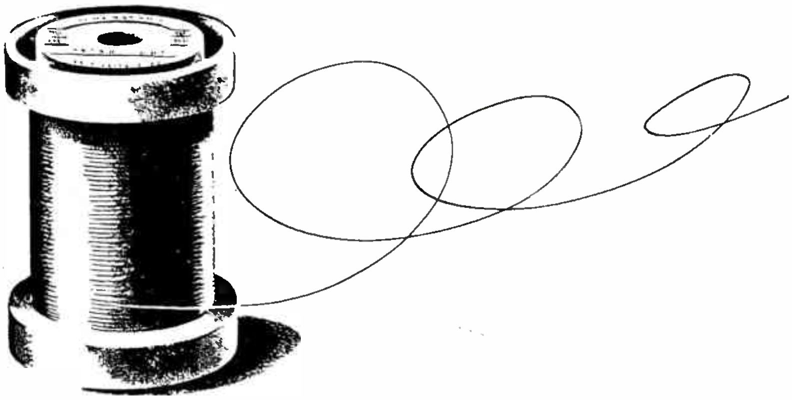
AUGUST 1959 Vol 36 *new series* No 8



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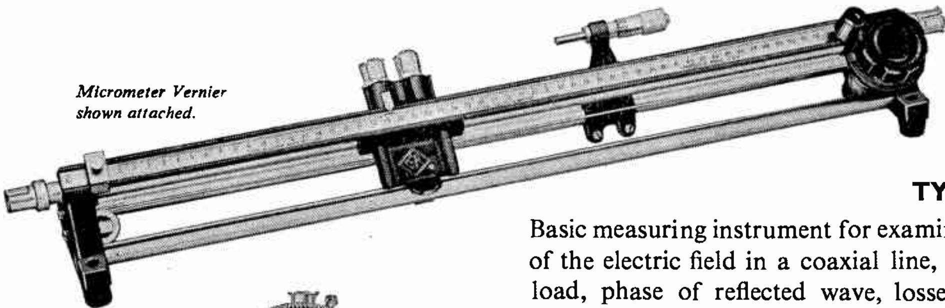

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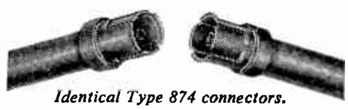

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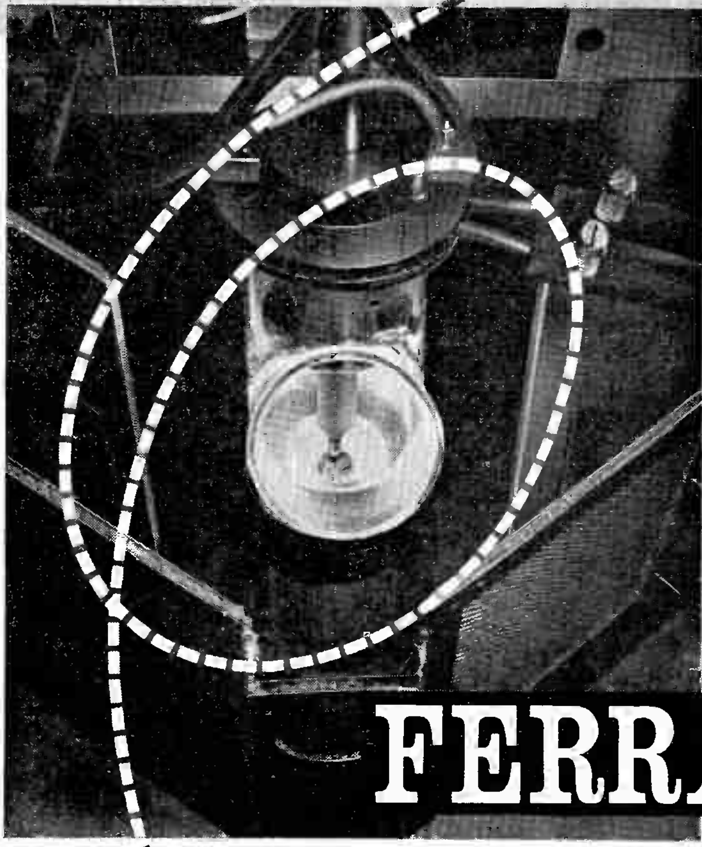

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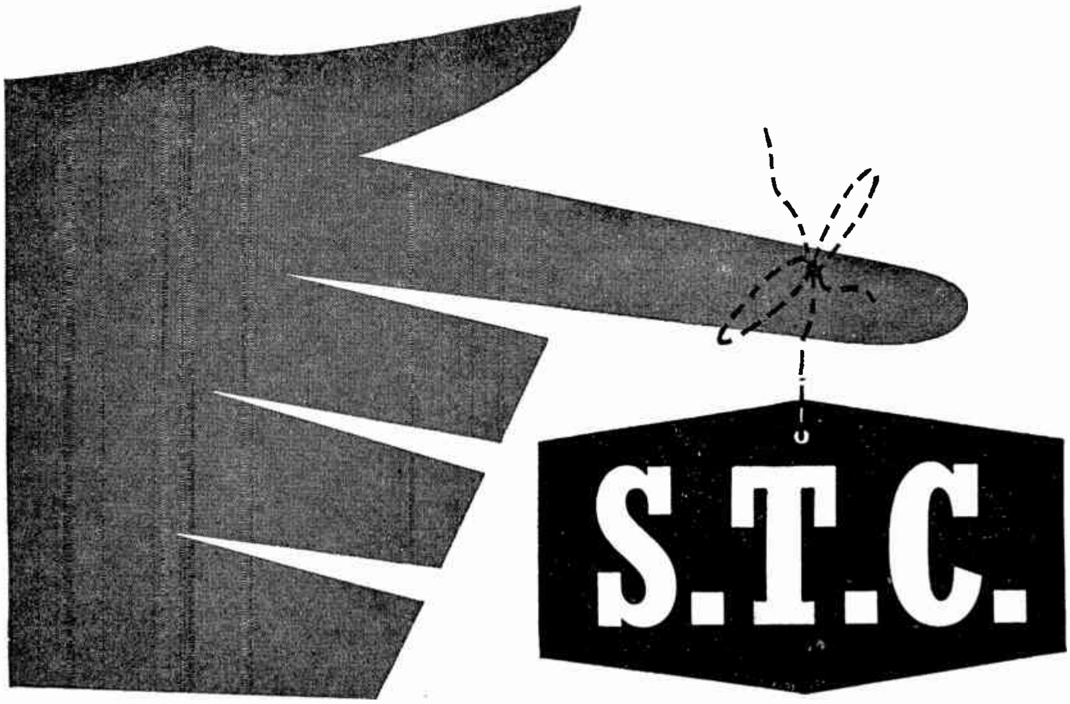

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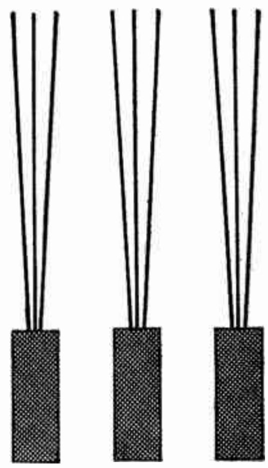

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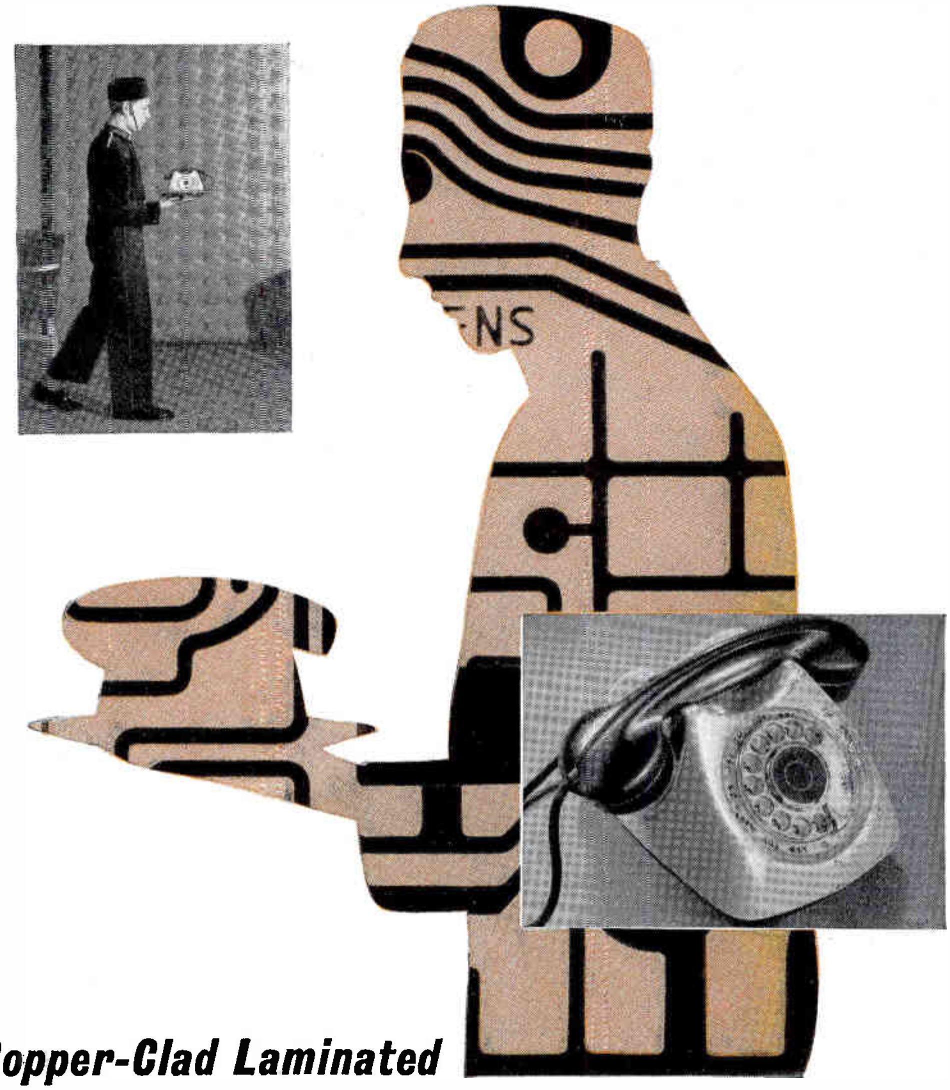

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type	zener voltage Vz @ Iz volts	zener current Iz mA	zener impedance Zz (max) @ Iz ohms	reverse current LIb @−10V 25° C μA	power dissipation P(max) Ts=50°C watts	typical temp. coefficient %/°C
1S501	22	150	4	10	8	0.08
1S502	24	150	4	10	8	0.08
1S503	27	150	4	10	8	0.08
1S504	30	150	5	10	8	0.08
1S505	33	150	5	10	8	0.08
1S506	36	150	6	10	8	0.09
1S507	39	150	6	10	8	0.09
1S508	43	100	7	10	8	0.09
1S509	47	100	8	10	8	0.09
1S510	51	100	10	10	8	0.10
1S511	56	100	11	10	8	0.10
1S512	62	50	14	10	8	0.10
1S513	68	50	16	10	8	0.10
1S514	75	50	24	10	8	0.11
1S515	82	50	26	10	8	0.11
1S516	91	50	40	10	8	0.12

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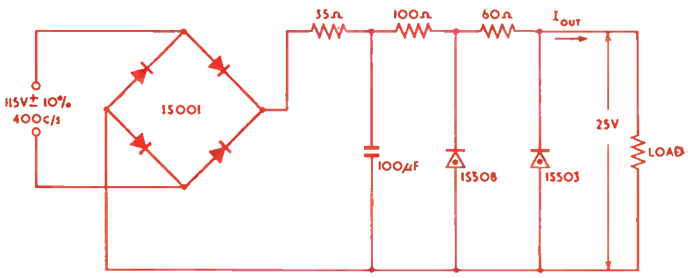

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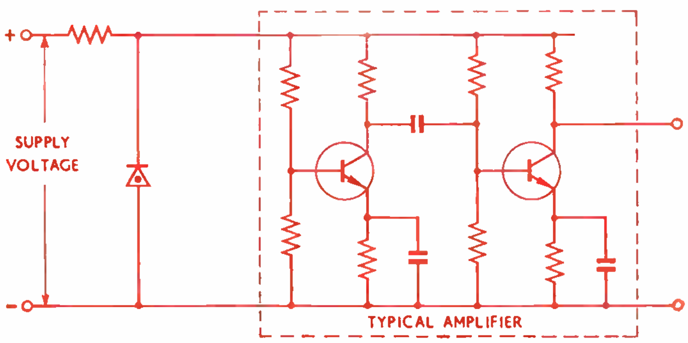

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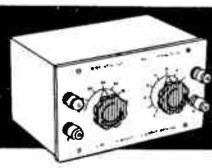

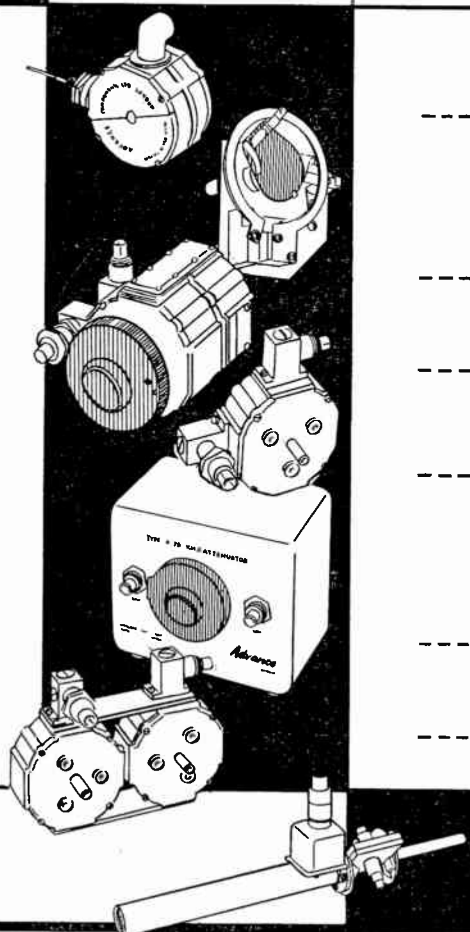

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Tentative Ratings and Characteristics

MAXIMUM DESIGN CENTRE RATINGS

Anode Dissipation (watts)	$p_{a(max)}$	3*
Screen Dissipation (watts)	$p_{g2(max)}$	1*
Anode Voltage (volts)	$V_{a(max)}$	250
Screen Voltage (volts)	$V_{g2(max)}$	250
Heater to Cathode Voltage (volts r.m.s.)	$V_{h\text{-}k(max)r.m.s.}$	200**

*With a grid to cathode circuit resistance not exceeding 10,000 ohms.
**Measured from cathode to higher potential heater pin.

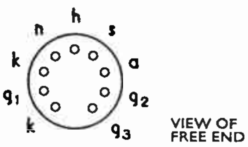
INTER-ELECTRODE CAPACITANCES (pF)†

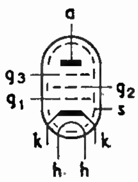
Grid 1 to Earth	c_{in}	9.0
Anode to Earth	c_{out}	3.7
Grid 1 to Anode	$c_{g1\text{-}a}$	0.007

†Measured in fully shielded jig, without can.

MAXIMUM DIMENSIONS (mm)

Overall Length	67.5
Seated Height	60.5
Diameter	22.2



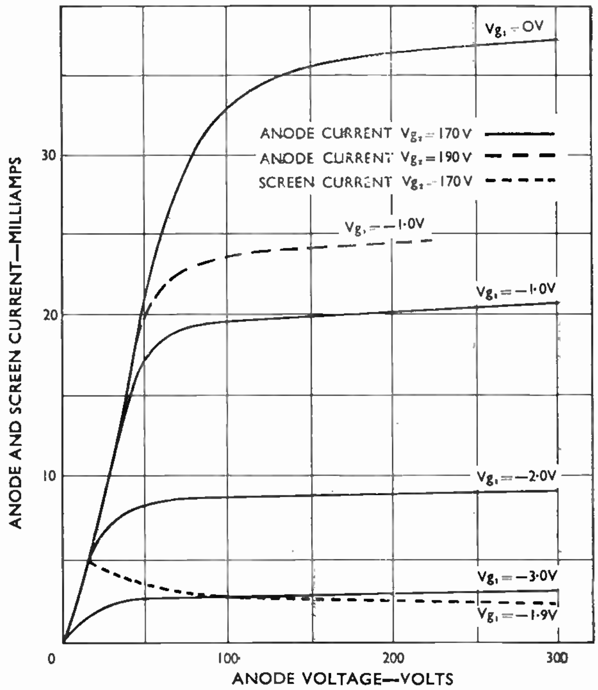


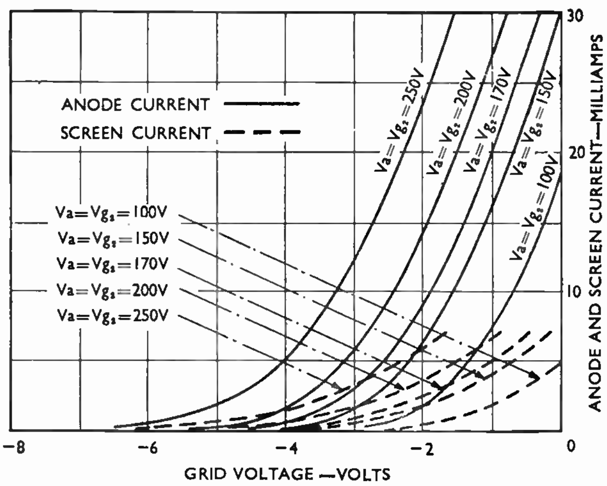
TYPICAL OPERATION

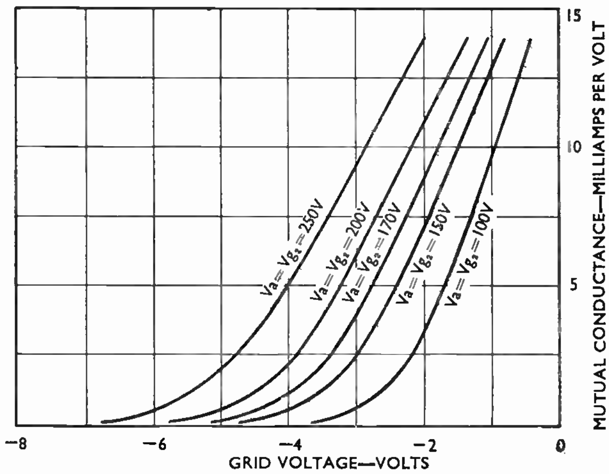
Anode Voltage (volts)	V_a	170
Screen Voltage (volts)	V_{g2}	170
Self Bias Resistance (ohms)	R_k	150
Anode Current (mA)	I_a	10
Screen Current (mA)	I_{g2}	2.6
Mutual Conductance (mA/V)	g_m	9.2
Inner Amplification Factor	$\mu_{g1\text{-}g2}$	64
Equivalent Grid Noise Resistance (ohms)	R_{eq}	670
Input Loss at 38 Mc/s (ohms)	$r_{g1\text{-}k}(W)$	8500*
Input Capacity Working (pF)	$c_{in}(W)$	12.1†Δ
Change in Input Capacity produced by biasing valve to cut-off (pF)	$\Delta c_{in}(W)$	2.45Δ
Figure of Merit (valve only) (Mc/s)		210‡
Effective Figure of Merit (valve and		130

*The two cathodes strapped and joined directly to earth.
Δ Measured at 38 Mc/s.
† Inter-electrode capacity with holder capacity balanced out.
‡ Given by $\dfrac{g_m \times 10^3}{2\pi\sqrt{c_{in}\,c_{out}}}$

Tentative Characteristic Curves of Ediswan Mazda Valve Type 6F23.







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CONTENTS

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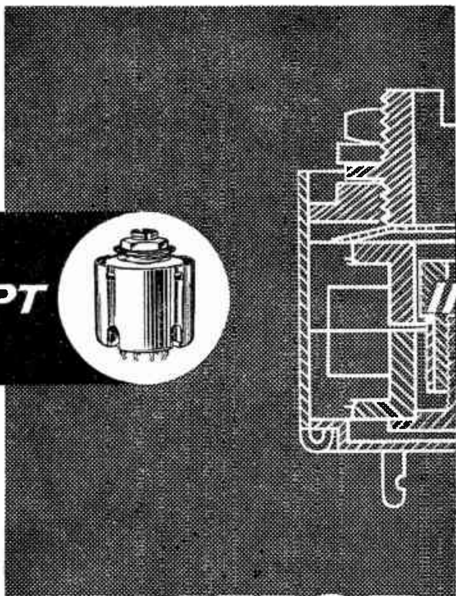

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AUGUST 1959 *incorporating WIRELESS ENGINEER*

Colour Vision

OUR attention has recently been drawn to some experiments with the reproduction of coloured pictures which have been carried out by E. H. Land, of Polaroid fame. The results of these experiments cannot be explained by the ordinary theory of colour vision, and it thus appears that this theory is in need of amendment or, at least, extension.

This month's "Fringe of the Field" contains an account of some of Land's work, and we were interested enough to carry out one experiment ourselves. We took two photographs of a scene on ordinary black-and-white film, one through a green filter and the other through a red filter. After development, positives were prepared and we projected and superposed the two pictures, using white light for the one taken through the green filter and red light for the other.

The resulting picture contained much more colour than the red, pink and white that one would expect. There were definite yellow, orange and green and an excellent brown.

By chance our picture was seen by a colleague, who at once said that it was not nearly as good as Land's! It turned out that he had been present at one of Land's demonstrations and we had from him a first-hand account of it. We need say here only that he confirmed the report, which we had seen elsewhere, that Land's pictures, projected with only white and red light, give a colour picture of quality comparable to that of modern colour transparencies, save only for the blue. We need say no more about this, for he recorded his impressions at the time in *Amateur Photographer* for 18th June 1958 (Vol. 115, p. 803).

Our interest in this, of course, arises out of its possible effect on colour television. We do not necessarily mean that we think the method could be directly applied to it. It might be, and there might be some saving of bandwidth. In our view, the importance is an indirect one. A new theory of colour vision seems necessary to explain Land's results. When this is found, it may well open the way to further developments, and it would be surprising if these did not have repercussions on colour television. At the moment, we can only say that there is much more in colour than meets the eye!

Drift Transistor

SIMPLIFIED ELECTRICAL CHARACTERIZATION

By J. te Winkel*

SUMMARY. *Equivalent circuits for a drift transistor are developed starting from a set of parameters derived from the physical principles underlying the device. It is shown what approximations are possible if limited frequency ranges or large values of the drift field are considered. Further simplifications are obtained from the introduction of suitably chosen frequency parameters. The resulting equivalent circuits appear to be simply related to those commonly used for a normal alloy transistor. The form is the same and the values of the circuit elements can be found by means of a number of multiplying factors that depend on the drift field only. These are given in graphical form.*

The derivation of an equivalent circuit for any transistor should start from the physical principles underlying the working of the device. For a drift transistor, where the effect of the built-in electric field has to be accounted for, the derivation itself presents no particular difficulty, provided that the field is assumed constant throughout the base region. However, the resulting expressions for the elements of an equivalent circuit appear to be quite involved. So there is a definite need for simplification, if necessary at the cost of exactness.

The purpose of the present paper is to develop two basic equivalent circuits, one for common-base operation and another for common-emitter operation. Special attention will be paid to the simplest possible representation of the circuit elements. Another paper will deal with the implications of this approach as regards the choice of parameters to characterize transistor performance and the measurement of these quantities.

Fundamental Parameters for a Common-Base Circuit

Among the small-signal parameters which are obtained from physical reasoning those associated with the transport of minority carriers through the base should be considered first. Other parameters to take account of are the junction capacitances and the resistances appearing between the junctions and the electrodes.

The base transport phenomena can be described by a set of fourpole parameters, which have been published in various forms [1,2]. The complete equivalent circuit is then obtained by adding to this fourpole the other elements mentioned above.

Of the four base transport parameters, as previously given by the author[2], two are related to the flow of minority carriers from the emitter through the base to the collector. The two others represent the control of this flow by the collector-to-base voltage via a change in the effective base width (Early-effect). Considering the two latter parameters first, it appears that they cannot be put in a simple and generally applicable form; moreover, actual values will differ considerably from the theoretical ones when the drift field is only approximately constant. However, as has been shown by Krömer[1], the Early-effect becomes small when the drift field is large. Therefore, when a simple equivalent circuit is aimed at, one is justified in neglecting the two parameters in question and in leaving what residual effects there are to be determined by measurement.

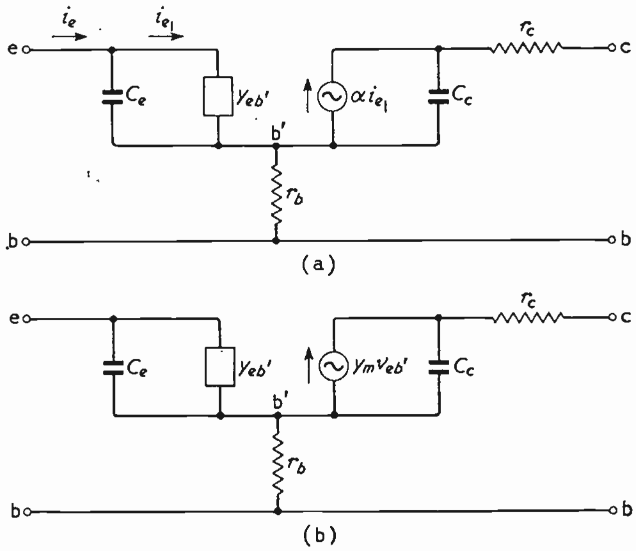


Fig. 1. *Fundamental common-base equivalent circuits*

The two remaining parameters then are the ratio of the minority carrier currents at the collector and emitter junctions, in other words the internal alpha, and the ratio of the emitter current and the voltage across the emitter–base junction, the internal input admittance. As an alternative to either of these parameters one may take the ratio of the collector current and the emitter-to-base voltage, the internal transconductance. The emitter efficiency (i.e., the ratio of the minority carrier current at the emitter junction and the total current crossing this junction) is assumed to be unity.

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

The complete common-base equivalent circuits thus obtained are shown in Fig. 1(a) and (b); α, y_{eb}' and y_m constituting the internal parameters. The junction capacitances are denoted by C_e and C_c, the series resistances by r_b and r_c. As shown, one has to distinguish between the internal base b′ and the base connection b and also between the current flowing in the emitter terminal i_e and the minority carrier current across the emitter junction i_{e1}. Expressions for the internal parameters can be found in Ref. 2 or, in another form, in Ref. 1.

The expression for the internal alpha in Fig. 1(a) is:

$$\alpha = \frac{e^{\frac{1}{2}A}}{\dfrac{\frac{1}{2}A}{\sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}} \sinh \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}} + \cosh \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}} \quad \ldots \quad (1)$$

and for the internal input admittance in Fig. 1(a) and (b)

$$y_{eb}' = \frac{1}{r_e} \cdot \frac{\frac{1}{2}A + \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}} \coth \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}}{\frac{1}{2}A + \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2}} \coth \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2}}} \quad \ldots \quad (2)$$

The transconductance in Fig. 2(b) is:

$$y_m = \frac{1}{r_e} \frac{e^{\frac{1}{2}A}}{\frac{1}{2}A + \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2}} \coth \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2}}} \cdot \frac{\sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}}{\sinh \sqrt{\frac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}} \quad \ldots \quad (3)$$

in these three expressions.

$$A = \frac{F\,W}{kT/q}, \quad \ldots \quad (4)$$

where F represents the drift field and W the base width. Thus A is equal to the drift potential across the base expressed in units of the thermal voltage kT/q. The quantity A is also simply related to the ratio of the impurity concentrations at the emitter and collector junctions[1,2].

$$A = \log_e \frac{N_e}{N_c} \quad \ldots \quad (5)$$

The quantity W/L is the ratio of base width to diffusion length; it is small and the square is usually negligible. Furthermore

$$\omega_0 = \frac{2D}{W^2}, \quad \ldots \quad (6)$$

with D equal to the diffusion constant of minority carriers in the base. This expression may be recognized as one which is sometimes used to define the cut-off frequency of a transistor for the case where the drift field is zero. However, the definition which is used in the present article would give a value 1·21 times larger. Thus, for the present, Equ. (6) should be looked upon as defining merely a frequency parameter ω_0.

Finally

$$r_e = \frac{kT}{q\,I_e} \quad \ldots \quad (7)$$

represents the emitter diode differential resistance, I_e being the emitter direct current.

Approximations for the three internal parameters are derived in the succeeding sections.

Fundamental Parameters for a Common-Emitter Circuit

For common-emitter operation one would prefer an equivalent circuit with the current generator connected between emitter and collector. The necessary transformations can be readily made. As internal parameters, one may have either the input admittance $y_{b'e}$ and the ratio of collector and base current: α' [Fig. 2(a)] or the input admittance and the transconductance [Fig. 2(b)]. Again, one must distinguish between internal and external base and between the current i_b flowing into the base terminal and that fraction i_{b1}, associated with the base transport only. The transformation from Fig. 1(a) to Fig. 2(a) yields

$$y_{b'e} = y_{eb'}(1 - \alpha) \quad \ldots \quad (8)$$

and

$$\alpha' = \frac{\alpha}{1-\alpha} \quad \ldots \quad (9)$$

The transconductance in Fig. 2(b) appears to have the same absolute value as that in Fig. 1(b). However, as shown, the phase is reversed.

Approximations for $y_{b'e}$ and α' will be discussed below.

Approximations for α.

The purpose of this section is to show in what ways the expression (1) for α can be simplified.

Large Values of A

When the argument is large the hyperbolic functions in (1) both tend to become equal to half the exponential.

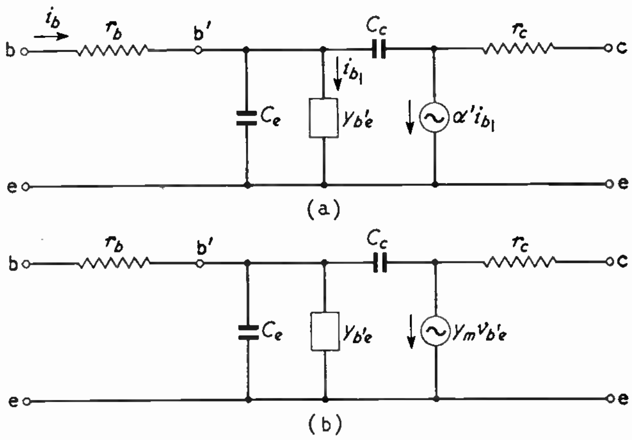


Fig. 2. *Fundamental common-emitter equivalent circuits*

For large values of A and ω/ω_0, neglecting W^2/L^2 at the same time, Equ. (1) can be simplified to

$$\alpha = \frac{2\sqrt{\tfrac{1}{4}A^2 + 2j\dfrac{\omega}{\omega_0}}}{\tfrac{1}{2}A + \sqrt{\tfrac{1}{4}A^2 + 2j\dfrac{\omega}{\omega_0}}} \exp.\left[-\left(\sqrt{\tfrac{1}{4}A^2 + 2j\frac{\omega}{\omega_0}} - \tfrac{1}{2}A\right)\right] \quad \ldots \ldots \ldots \ldots \ldots (10)$$

The range of validity of this expression must be found by computation. It may be noted first that (10) again approaches (1) at very low frequencies, if one neglects W^2/L^2 in (1) as well. Computing the difference between (10) and (1) as a function of frequency for a given value of A, one therefore finds a maximum. Taking $A = 3$, for instance, this occurs at $\omega/\omega_0 = 1{\cdot}34$ and amounts to less than 1%. The difference appears to decrease rapidly if one takes larger values for A. Thus the approximation (10) may be used for all values of ω/ω_0 provided that $A \geqslant 3$.

Evaluation of α' and Introduction of the Frequency Parameter ω_1

The equation (10) for α still contains ω_0 as the unit of frequency, which is only a meaningful quantity when the drift field is zero. However, a change of the frequency scale may be obtained by considering the ratio of collector and base current α', as given by (9).

An expression for α' valid at low frequencies can be found by taking (1) and expanding the square root and the hyperbolic functions. To obtain a useful result the expansion will be restricted to values of A such that the condition

$$\left|\frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right| \ll \tfrac{1}{4}A^2$$

is fulfilled. Then, with this restriction,

$$\alpha = \frac{e^{A/2}}{\left(1 - \dfrac{2W^2}{A^2L^2} - j\dfrac{4\omega}{A^2\omega_0}\right)\left\{\sinh\tfrac{1}{2}A + \dfrac{\cosh\frac{1}{2}A}{A}\left(\dfrac{W^2}{L^2} + 2j\dfrac{\omega}{\omega_0}\right)\right\} + \cosh\tfrac{1}{2}A + \dfrac{\sinh\frac{1}{2}A}{A}\left(\dfrac{W^2}{L^2} + 2j\dfrac{\omega}{\omega_0}\right)} \quad (11)$$

The substitution of this expression in (9) leads to

$$\alpha' = \frac{\tfrac{1}{2}A^2}{A - 1 + e^{-A}} \, \frac{\dfrac{2L^2}{W^2}}{1 + j\dfrac{\omega}{\omega_0}\cdot\dfrac{2L^2}{W^2}} \quad \ldots (12)$$

Considering now a transistor for which the drift field is zero, Equ. (1) will reduce to

$$\alpha(0) = \frac{1}{\cosh\sqrt{\dfrac{W^2}{L^2} + 2j\dfrac{\omega}{\omega_0}}}, \quad \ldots \ldots (13)$$

which expression at low frequencies, under the condition

$$\left|\frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right| \ll 1,$$

can be expanded to

$$\alpha'(0) = \frac{\dfrac{2L^2}{W^2}}{1 + j\dfrac{\omega}{\omega_0}\dfrac{2L^2}{W^2}} \quad \ldots \ldots \ldots \ldots (14)$$

When comparing (12) and (14), it is seen that the presence of the drift field increases the zero-frequency value of α'; it is multiplied by the first factor that appears in (12). The cut-off frequency of α' is not affected, however. Denoting the zero-frequency values of α' and $\alpha'(0)$ by α'_0 and $\alpha'_0(0)$ respectively and the cut-off frequency of both quantities by $\omega_{c\alpha'}$, one has from (14) the well-known relations for a normal field-free transistor

$$\omega_{c\alpha'} = \omega_0 \frac{W^2}{2L^2} \text{ and } \alpha'(0) = \frac{2L^2}{W^2}$$

or

$$\alpha'_0(0)\,\omega_{c\alpha'} = \omega_0 \quad \ldots \ldots \ldots \ldots (15)$$

The corresponding relation for the drift transistor, taken from (12) would be

$$\alpha'_0\,\omega_{c\alpha'} = \frac{\tfrac{1}{2}A^2}{A - 1 + e^{-A}}\,\omega_0 \quad \ldots \ldots (16)$$

When A is so small that the condition for (12) is no longer fulfilled and one has instead

$$\left|\tfrac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right| \ll 1,$$

it can be shown that the relation (16) must be replaced by

$$\alpha'_0\,\omega_{c\alpha'} = \frac{1}{1 - \tfrac{1}{3}A}\,\omega_0.$$

This, however, is also the expression obtained by expanding (16) for small values of A.

It is now proposed to use the quantity $\alpha'.\omega_{c\alpha'}$, given by (16) and henceforward to be denoted by ω_1, as a new unit of frequency that should replace ω_0 in (1) and (10). For a field-free transistor, ω_0 and ω_1 are obviously equal; for a drift transistor, one should have from (16)

$$\frac{\omega_1}{\omega_0} = \frac{\frac{1}{2}A^2}{A - 1 + e^{-A}} \quad \ldots \quad (17)$$

As will appear from the following, the quantity ω_1 can be used to describe the frequency behaviour of the alpha of a drift transistor in much the same way as ω_0 can for a normal transistor. Thus, comparing transistors of the same base width, the relation (17) may also be interpreted as showing the measure in which the high-frequency properties may be improved by the incorporation of a drift field. It also will represent the increase in the zero-frequency value of α'.

The relation (17) is given in a graphical form by Fig. 3. As shown by Krömer[1] the highest value of A obtainable in practice is about 8, for the upper limit in this figure and in those to follow a value of 9 has been taken.

The fact that the cut-off frequency of α' is not affected by the drift field needs some comment. For a given alternating collector current the base current can be divided into two parts, one which supplies the majority carriers necessary for recombination and another, in phase quadrature with the former, which provides the majority carriers needed to maintain space charge neutrality. Considering now the minority carriers in the base, it may be assumed that at low frequencies their concentration will vary at a uniform rate over the base region. In this case, the two components of the base current will each be proportional to the amplitude of the varying part of the total minority carrier charge; the proportionality factor being the reciprocal of carrier life-time $1/\tau$ for the first component and the angular frequency ω for the other. At the cut-off frequency of α' both currents will be equal in magnitude. Thus

$$\omega_{c\alpha'} = \frac{1}{\tau}$$

a result that also could have been found from (6) and (15), with the help of $L^2 = D\tau$. It is seen that the cut-off frequency $\omega_{c\alpha'}$ does not depend either on the total number of minority carriers in the base or on their distribution, quantities which will depend on the drift field.

In most transistors, one has to consider other components of the base current besides those discussed above, which were the only ones taken account of in the derivation of Equ. (1). Additional components will result from a surface recombination current and a flow of majority carriers from the base to the emitter. As long as these currents are proportional to the bulk recombination current discussed above, all that has been said earlier will still hold, provided τ is replaced by some other quantity. If L is changed accordingly, the relations represented by Equs (11) to (17) will thus also be valid in this more general case; provided, of course, that the term replacing W^2/L^2 remains sufficiently small. Similar remarks apply when τ has different values in different parts of the base.

Representation of α with ω₁ as a Frequency Parameter

Replacing ω_0 by ω_1 and neglecting W^2/L^2 the expression (1) for α becomes

$$\alpha = \frac{e^{\frac{1}{2}A}}{\dfrac{\sinh \frac{1}{2}A\sqrt{1 + j\dfrac{\omega}{\omega_1}\dfrac{4}{A - 1 + e^{-A}}}}{\sqrt{1 + j\dfrac{\omega}{\omega_1}\dfrac{4}{A - 1 + e^{-A}}}} + \cosh \frac{1}{2}A\sqrt{1 + j\dfrac{\omega}{\omega_1}\dfrac{4}{A - 1 + e^{-A}}}} \quad \ldots \quad (18)$$

When $A \geqslant 3$, the simplified expression (10) may be used. In this case, the exponential in (17) may be neglected and the introduction of ω_1 results in

$$\alpha = \frac{2\exp.\left[-\frac{1}{2}A\left(\sqrt{1 + j\dfrac{\omega}{\omega_1}\dfrac{4}{A-1}} - 1\right)\right]}{1 + \dfrac{1}{\sqrt{1 + j\dfrac{\omega}{\omega_1}\dfrac{4}{A-1}}}} \quad . \quad (19)$$

For a normal transistor $A = 0$ and $\omega_0 = \omega_1$; neglecting W^2/L^2 here also, one has from (13)

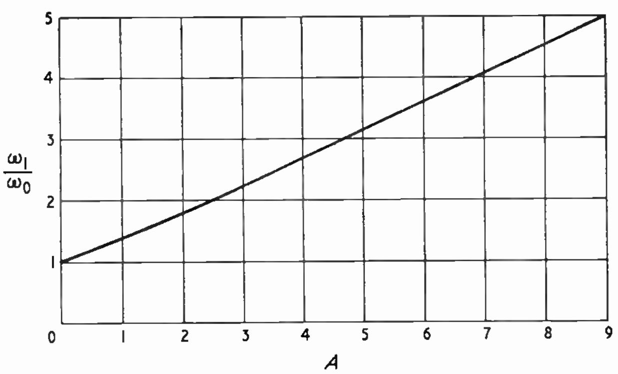
$$\alpha = \frac{1}{\cosh\sqrt{2j\dfrac{\omega}{\omega_1}}} \quad \ldots \quad (20)$$

A plot of α for various values of the drift potential A using ω_1 as a unit of frequency is shown in Fig. 4. Also shown is the half circle

$$\frac{1}{1 + j\dfrac{\omega}{\omega_1}}$$

which approximates all curves at low frequencies.

Fig. 3. *The ratio of the frequency parameters ω_1 and ω_0*



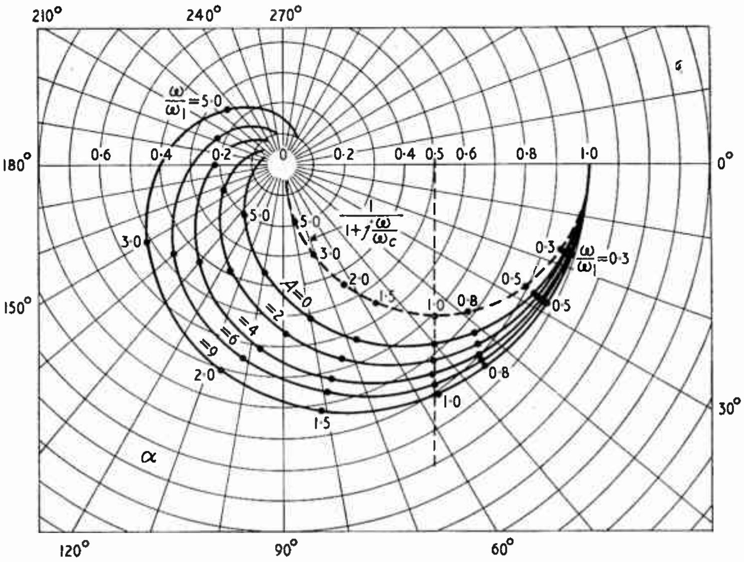


Fig. 4. *α-curves with ω_1 as the frequency parameter*

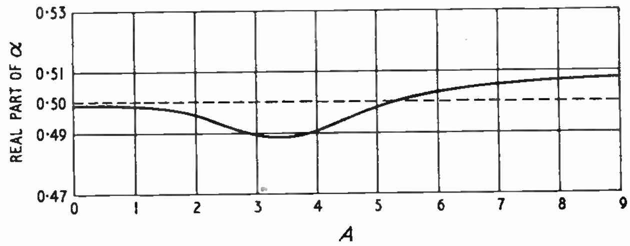


Fig. 5. *The real part of α at $\omega = \omega_1$*

It may be noted that W^2/L^2 has been neglected purely as a matter of convenience. With the help of (12) and (9) a correction can easily be applied; for not too small values of α'_0, however, it will only be important at low frequencies.

When examining Fig. 4 it will be seen that the points where $\omega/\omega_1 = 1$ lie very nearly on the vertical line that passes through the point 0·5 on the horizontal axis. This suggests that, to a very close approximation, one might also define ω_1 as the frequency where the real part of α equals 0·5†. The error made can be seen from Fig. 5, which shows the real part of α at $\omega = \omega_1$ for various values of A.

Cut-off Frequency of α

A parameter commonly used to describe the frequency behaviour of a transistor is the α cut-off frequency ω_c. By definition, it is the frequency at which the magnitude of α equals $\frac{1}{2}\sqrt{2}$. When computing the curves of Fig. 4 ω_c can be found by interpolation, and so one can determine a factor K given by

$$K = \frac{\omega_c}{\omega_1} \quad \ldots \quad (21)$$

The phase angle of α at the cut-off frequency can be calculated at the same time. For reasons which will become obvious from the next section, the phase angle itself will not be used to characterize transistor performance; it will be replaced by a quantity ϕ which is 45 degrees or $\pi/4$ radians less than the negative of this angle. Figs. 6 and 7 show ϕ as a function of the drift potential A.

An interesting fact which can be observed from these figures is that, for values of A of practical interest, both K and ϕ are very nearly linear functions of A. One has, with an error of less than 1% and up to $A = 9$

$$K = 1{\cdot}21 + 0{\cdot}90\,A \quad \ldots \quad (22)$$

and

$$\phi = 0{\cdot}221 + 0{\cdot}98\,A \quad \text{in radians,}$$

or

$$\phi = 12{\cdot}7 + 5{\cdot}59\,A \quad \text{in degrees.} \quad \ldots \quad (23)$$

For a normal transistor $A = 0$ and $K = 1{\cdot}21$, hence $\omega_c = 1{\cdot}21\,\omega_0$, as mentioned when discussing Equ. (6).

Representation of α with ω_c as a Frequency Parameter

Changing the frequency parameter ω_1 to ω_c using values of K specified by (22), Fig. 4 has been redrawn in Fig. 8. The half-circle is now meant to represent

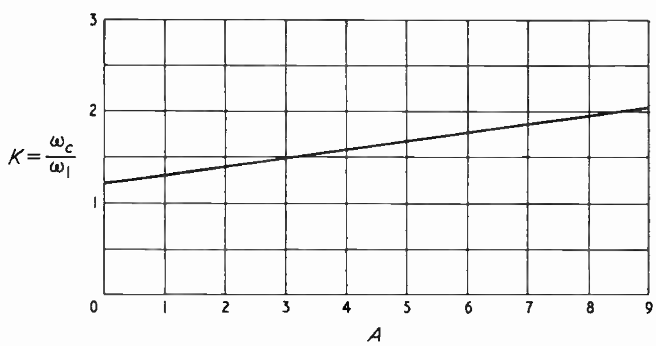
$$\frac{1}{1 + j\dfrac{\omega}{\omega_c}} \quad \ldots \quad (24)$$

An examination of Fig. 8 will show that the points on the α-curves and on the half-circle that correspond to fixed values of ω/ω_c smaller than unity lie very nearly on circles that have their centres at the origin. So, provided that $\omega/\omega_c \leqslant 1$, the magnitude of α will be practically independent of A and equal to the absolute value of (24).

When the values computed for the phase angle of α are examined, another interesting property is revealed. If one takes a particular α-curve and the half-circle and compares points that have equal values of ω/ω_c, it appears that the phase difference is very nearly proportional to frequency.

The two observations combined suggest the following

Fig. 6. *The ratio of the frequency parameters ω_c and ω_0*



† For this suggestion the author is indebted to Mr. L. G. Cripps of Mullard Research Laboratories, Salfords, who will discuss its implications in a succeeding article.

approximation for α‡

$$\alpha = \frac{\exp.\left[-j\phi\frac{\omega}{\omega_c}\right]}{1+j\frac{\omega}{\omega_c}} \quad \ldots \quad (25)$$

If $A \geqslant 3$ a simplified expression can be found when one introduces α, as given by (1), in the equation (2) and subsequently substitutes this quantity by the simplified expression (10). Neglecting exp. $(-A)$ in respect to A, one gets

$$y_{eb}' = \frac{1}{2r_e}\left(1+\sqrt{1+j\frac{\omega}{\omega_1}\frac{4}{A-1}}\right)\frac{1-\exp.[-A]}{1-\exp.\left[-A\sqrt{1+j\frac{\omega}{\omega_1}\frac{4}{A-1}}\right]} \quad \ldots \quad (27)$$

The quantity ϕ has been defined in the preceding section. It is seen that by this definition and by that of ω_c the equation (25) will give the correct value when $\omega = \omega_c$. A final check on its validity can be obtained by comparing computed values with (18), (19) or (20). Considering first the frequency range $\omega = 0$ to $\omega = \omega_c$ it is found that the error has a maximum at some intermediate value of ω and that it increases with A. Taking $A = 9$ as the highest value of practical interest the maximum appears to occur at $\omega/\omega_c = 0{\cdot}4$ and to amount to $1{\cdot}5\%$. At frequencies above ω_c the error becomes larger, as may be seen from Fig. 8, where Equ. (26) is represented by the dashed curves. A general criterion for the validity of (25) in this frequency region can be given with the help of the frequency parameter ω_1. For all values of A the error is about 3% for $\omega/\omega_1 = 2{\cdot}5$ and 10% for $\omega/\omega_1 = 3$.

Conclusion

It is shown that to a good approximation the internal alpha of a drift transistor can be characterized in a simple form by two quantities only. These may be either the drift potential A and the particular frequency ω_1 where the real part of α equals 0·5 or the α cut-off frequency ω_c and a quantity ϕ associated with the phase angle of α at this frequency.

The first approximation is valid over a wide frequency range, it is given in two forms depending on the magnitude of A. For frequencies well below ω_1 and for not too large values of A, the parameter ω_1 alone might suffice. The second approximation, simpler in form, is only valid somewhat beyond the cut-off frequency ω_c.

As an alternative, one might also take as one characterizing parameter the frequency ω_1, and as the other, instead of A, either the frequency ω_c or the phase angle ϕ.

Approximations for the Common-Base Input Admittance

Introduction of ω_1

An expression for the input admittance with ω_1 as the frequency parameter is obtained when one takes the original Equ. (2), neglecting W^2/L^2, and introduces ω_1 according to (17):

Admittance at Low Frequencies

For small values of ω/ω_1 the product of the square root and the hyperbolic function in (26) can be expanded with the result

$$y_{eb}'_0 = \frac{1}{r_e}\left\{1+j\frac{\omega}{\omega_1}\frac{1}{A-1+e^{-A}}\left(1-\frac{2A}{A-1}+e^{-A}\right)\right\} \quad \ldots \quad (28)$$

Thus, at a low frequency, the admittance can be regarded as consisting of the parallel combination of a resistance r_e and a capacitance

$$C_{eb}'_0 = \frac{1}{\omega_1 r_e}\frac{1}{A-1+e^{-A}}\left(1-\frac{2A}{A-1}+e^{-A}\right) \quad (29)$$

If $A \geqslant 3$ one may neglect exp. $(-A)$ in this expression.

For a normal transistor one has $A = 0$ and $\omega_0 = \omega_1$, in this case Equ. (2) will reduce to

$$y_{eb}'\ (0) = \frac{1}{r_e}\sqrt{2j\frac{\omega}{\omega_1}}\coth\sqrt{2j\frac{\omega}{\omega_1}}\ . \quad \ldots \quad (30)$$

The capacitance at low frequencies again will follow from the expansion, its value is

$$C_{eb}'_0(0) = \frac{2}{3\omega_1 r_e}\ . \quad \ldots \quad (31)$$

One may put Equs. (29) and (31) in a general form:

$$C_{eb}'_0 = \frac{1}{M\omega_1 r_e}\ . \quad \ldots \quad (32)$$

where M depends on A only. Its value ranges from 1·5 for $A = 0$ to about 8 for $A = 9$, see Fig. 9.

Admittance of Other Frequencies

For fixed values of A one may calculate the real and imaginary parts of the admittance as given by (26) or (27) and compare these to the low-frequency values specified by (28). The result is shown in Fig. 10, the change in conductance by the product $g_{eb}'r_e$ and the change in capacitance by the ratio $C_{eb}'/C_{eb}'_0$. It appears that up to $\omega/\omega_1 = 1$ the change with frequency is rather small; therefore, it would not be worth while to attempt a better approximation by replacing the parallel combination of r_e and $C_{eb}'_0$ by a more elaborate network.

$$y_{eb}' = \frac{1}{r_e}\cdot\frac{1+\sqrt{1+j\frac{\omega}{\omega_1}\frac{4}{A-1+e^{-A}}}\coth\tfrac{1}{2}A\sqrt{1+j\frac{\omega}{\omega_1}\frac{4}{A-1+e^{-A}}}}{1+\coth\tfrac{1}{2}A} \quad \ldots \quad (26)$$

‡ The approximation given in this section agrees in form with the one that Thomas and Moll have derived from experimental data[3]. However, the analysis given by these authors requires that $\phi = K - 1$, which is only approximately true, as may be seen from (22) and (23).

As may be seen from the equivalent circuit (Fig. 1) the junction capacitance C_e must be added to the input capacitance $C_{eb'}$ derived above. It should be noted that

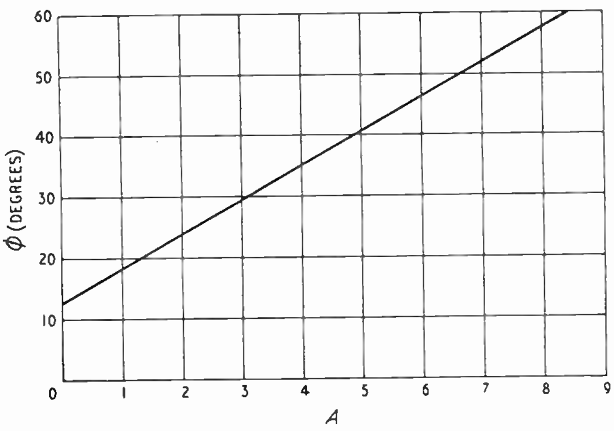


Fig. 7. *The phase-angle parameter ϕ*

the former is nearly independent of the emitter direct current, while the latter is directly proportional to it.

Approximations for Transconductance

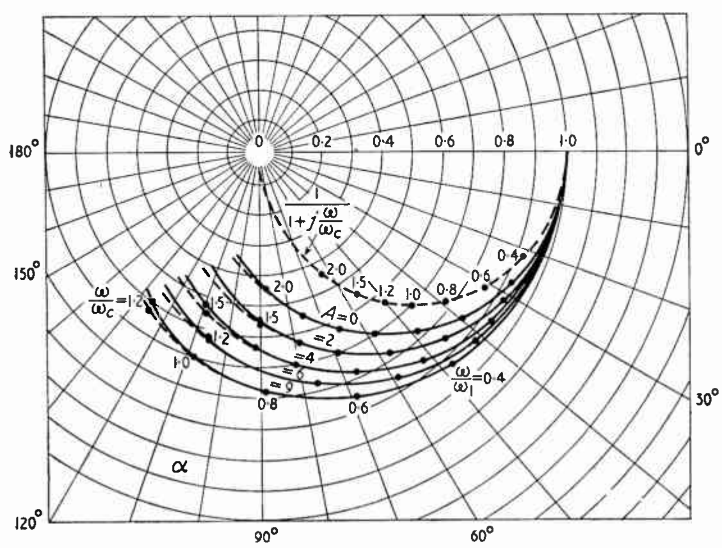
This quantity, which can be used in both the common-emitter and the common-base circuit, is given in full by Equ. (3). Neglecting W^2/L^2, the introduction of ω_1 according to (17) will simplify this expression to

$$y_m = \frac{1}{r_e} \frac{\sqrt{1 + j\frac{\omega}{\omega_1}\frac{4}{A-1+e^{-A}}}}{\sinh \frac{1}{2}A\sqrt{1 + j\frac{\omega}{\omega_1}\frac{4}{A-1+e^{-A}}}} \sinh \tfrac{1}{2}A \,. \quad \ldots \quad (33)$$

At low frequencies, Equ. (33) may be approximated by the first terms of the series expansion, as follows

$$y_{m0} = \frac{1}{r_e} \frac{1}{1 + j\frac{\omega}{\omega_1}\frac{A\coth\frac{1}{2}A - 2}{A-1+e^{-A}}} \quad \ldots \quad \ldots \quad (34)$$

Fig. 8. *α-curves with ω_c as the frequency parameter. Full lines: exact values from Fig. 4; dashed lines: approximate values according to Equ.* (25)



For a normal transistor one can put $A = 0$ and $\omega_0 = \omega_1$ and Equ. (3) will reduce to

$$y_m(0) = \frac{1}{r_e} \cdot \frac{\sqrt{2j\frac{\omega}{\omega_1}}}{\sinh\sqrt{2j\frac{\omega}{\omega_1}}}, \quad \ldots \quad \ldots \quad \ldots \quad (35)$$

the low-frequency value of which is

$$y_{m0}(0) = \frac{1}{r_e} \cdot \frac{1}{1 + j\frac{\omega}{3\omega_1}} \,. \quad \ldots \quad \ldots \quad \ldots \quad (36)$$

The relations (34) and (36) may be put in the general form

$$y_{m0} = \frac{1}{r_e} \frac{1}{1 + j N\frac{\omega}{\omega_1}}, \quad \ldots \quad \ldots \quad \ldots \quad (37)$$

where N will depend only on A. Its value varies from 0·333 for $A = 0$ to 0·87 for $A = 9$, see Fig. 11.

When computing the difference between (33) and (37) it is found that the approximation (37) can be used up to $\omega/\omega_1 = 0{\cdot}5$; the error is about 6% in magnitude and 1° in phase angle. With $\omega/\omega_1 = 0{\cdot}25$ the error is less than 1%.

Approximation for α'

A suitable expression has been derived already in the form of Equ. (12); i.e.,

$$\alpha' = \frac{\frac{1}{2}A^2}{A-1+e^{-A}} \frac{\frac{2L^2}{W^2}}{1 + j\frac{\omega}{\omega_0}\cdot\frac{2L^2}{W^2}},$$

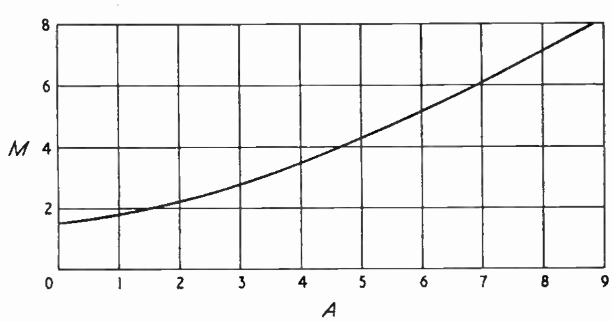
or in another form, introducing ω_1 from (17) and the zero-frequency value of α',

$$\alpha' = \frac{\alpha'_0}{1 + j\alpha'_0\frac{\omega}{\omega_1}} \,. \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (38)$$

The validity of this approximation will be realized by recalling the condition under which Equ. (12) was derived:

$$\left|\frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right| \ll \tfrac{1}{4}A^2 \,.$$

Fig. 9. *The multiplying factor M for the common-base input capacitance*



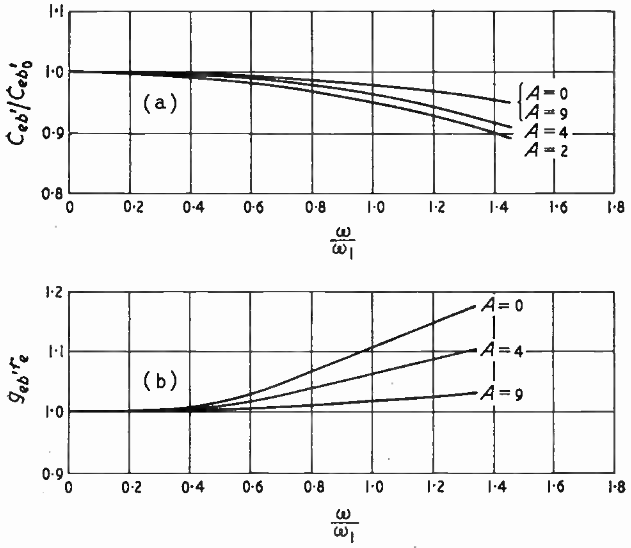


Fig. 10. *The change with frequency of the common-base input capacitance (a) and conductance (b)*

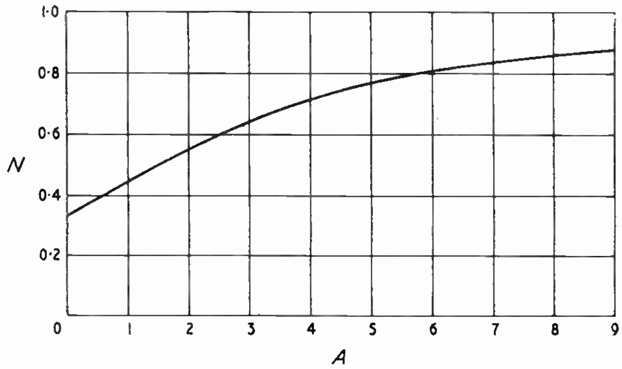


Fig. 11. *The multiplying factor N associated with the transconductance*

Inserting α' from (12) in this relation, it becomes

$$\left|\alpha'\right| \gg \frac{4}{A - 1 + e^{-A}} \cdot \qquad \ldots \quad \ldots \quad \ldots (39)$$

For very small A or $A = 0$ the condition is

$$\left|\tfrac{1}{4} A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right| \ll 1 \ ,$$

with the result

$$\left|\alpha'\right| \gg 2 \ . \qquad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots (40)$$

Using (12) and (17) one can also write

$$y_{b'e0} = \frac{1}{\alpha'_0 r_e} + j\frac{\omega}{\omega_1 r_e} \cdot \qquad \ldots \quad \ldots \quad \ldots (43)$$

Thus, the admittance is the parallel combination of a resistance $\alpha_0' r_e$ and a capacitance $C_{b'e} = 1/\omega_1 r_e$. In a similar manner one obtains for a normal field-free transistor

$$y_{b'e0} = \frac{1}{\alpha'_0(0)\, r_e} + j\frac{\omega}{\omega_0 r_e} \cdot$$

It is seen that this expression is also applicable to a drift transistor when ω_0 is replaced by ω_1.

Admittance at Other Frequencies

To this end one can take Equ. (41), neglect W^2/L^2 directly (which is equivalent to putting $\alpha'_0 = \infty$), and introduce ω_1 from (17). The result is

$$y_{b'e} = \frac{1}{r_e} \cdot \frac{1 - e^{-A}}{2}\left\{1 + \sqrt{1 + j\frac{\omega}{\omega_1}\frac{4}{A - 1 + e^{-A}}}\ \frac{\cosh \tfrac{1}{2}A\sqrt{1 + j\frac{\omega}{\omega_1}\frac{4}{A-1+e^{-A}}} - e^{\frac{1}{2}A}}{\sinh \tfrac{1}{2}A\sqrt{1 + j\frac{\omega}{\omega_1}\frac{4}{A-1+e^{-A}}}}\right\} \cdot \qquad \ldots \quad \ldots (44)$$

Approximations for the Common-Emitter Input Admittance

The full expression for this admittance, substituting (1) and (2) in (8), is

$$y_{b'e} = \frac{1}{r_e} \cdot \frac{\tfrac{1}{2}A}{\tfrac{1}{2}A + \sqrt{\tfrac{1}{4}A^2 + \frac{W^2}{L^2}}\ \coth\sqrt{\tfrac{1}{4}A^2 + \frac{W^2}{L^2}}}\left\{1 + \frac{\sqrt{\tfrac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}\left(\cosh\sqrt{\tfrac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}} - e^{\frac{1}{2}A}\right)}{\tfrac{1}{2}A \sinh\sqrt{\tfrac{1}{4}A^2 + \frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}}}\right\} \qquad \ldots (41)$$

Admittance at Low Frequencies

The expansion of the square root and the hyperbolic functions in (41), under the condition

$$\left|\frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right| \ll \tfrac{1}{4}A^2$$

and neglecting W^2/L^2 with respect to $\frac{1}{4}A^2$, will yield

$$y_{b'e0} = \frac{1}{r_e}\left(\frac{W^2}{L^2} + 2j\frac{\omega}{\omega_0}\right)\frac{A - 1 + e^{-A}}{A^2} \cdot \qquad \ldots (42)$$

The admittance can be computed for various values of A and plotted in the complex plane. It is then found that, up to $\omega/\omega_1 = 2$, the points lie very nearly on half-circles that pass through the origin and have their centres on the horizontal axis. This suggests that a good approximation would be obtained by considering the impedance to consist of a resistance and a capacitance in series. So one might put

$$\frac{1}{y_{b'e}} = Pr_e + \frac{1}{j\omega C_{b'e0}} \cdot \qquad \ldots \quad \ldots \quad \ldots (45)$$

The factor P in the real part has been introduced to

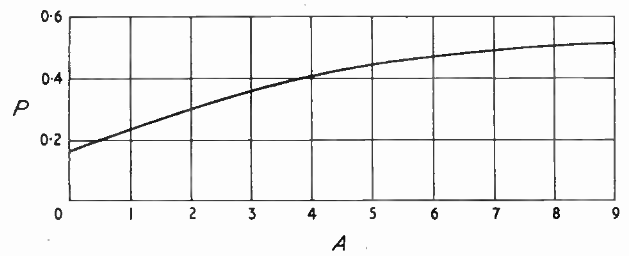
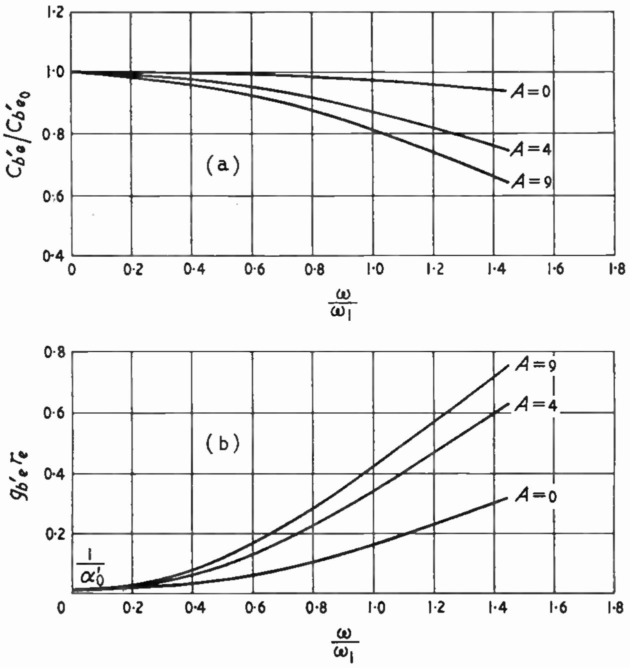


Fig. 12. *The multiplying factor P for the common-emitter input resistance*

Fig. 13. *The change with frequency of the common-emitter input capacitance (a) and conductance (b)*



represent the effect of the drift field, the imaginary part is assumed to be determined by the low-frequency capacitance derived from Equ. (43)

$$C_{b'e0} = \frac{1}{\omega_1 r_e} .$$

Computed values of (45) can be compared with the reciprocal of (44). It is then found that the approximation Pr_e is correct within 1% up to $\omega/\omega_1 = 1$ and $A = 9$. The values of P following from this calculation range from 0·167 for $A = 0$ to 0·51 for $A = 9$, see Fig. 12. The difference between the series capacitance deduced from (44) and $C_{b'e0}$ is about 2% at $\omega/\omega_1 = 1$ and $A = 9$ and still less at lower values of A.

To obtain an approximation that will be usable at very low frequencies as well, one may, of course, connect a resistance $\alpha'_0 r_e$ in parallel with the series combination.

For many practical purposes, one would prefer a representation of the common-emitter input admittance in the form of the parallel combination of a resistance and a capacitance. To this end Fig. 13 has been derived from the real and imaginary parts of (44). The change with frequency of the conductance is shown by the product $g_{b'e}r_e$, that of the capacitance by the ratio $C_{b'e}/C_{b'e0}$.

The change of the parallel capacitance is seen to be small when moderately low frequencies are considered. The conductance shows a considerable increase, however, in particular when the field is large. Its value will no longer depend on α'_0, provided that the latter quantity is reasonably large; a value of 100 has been assumed for Fig. 13.

Conclusion

By the introduction of suitable multiplying factors and by a proper choice of the frequency parameter small-signal equivalent circuits for a drift transistor can be obtained that are essentially the same as those in common use for the normal alloy transistor.

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SPUTTERED RESISTORS INCREASE COMPONENT DENSITY

According to a paper presented by D. A. McLean of Bell Telephone Laboratories at the recent Western Electronics Conference, San Francisco, U.S.A., sputtered thin-film resistors, formed from refractory metals such as tantalum and titanium, may be one of the more important developments in microminiature electronics. Such resistors can be produced on glass or ceramic bases in lines as narrow as 0·001 in., spaced 0·001 in. apart.

These newly developed resistors rely on a high precision masking process which makes it possible to produce thin films in specifically restricted locations.

In producing a resistor, an over-all thin film of copper is first deposited on the ceramic or glass base. Then the desired pattern is etched into the copper surface by standard photo-etching techniques, leaving the base material exposed. Tantalum or other refractory metals are then deposited on to the etched copper pattern and the whole unit placed in an etching bath. The copper with its overlay of tantalum is removed, leaving behind only the tantalum which is in direct contact with the bare surface. Since the masks are extremely thin, fine detail is possible. Also, since the sputtered materials adhere to the base material itself, support considerations are not necessary and complex patterns can be produced.

In one experiment, a three-stage flip-flop circuit which occupied a standard printed-circuit card about 3½ in. × 7 in. was reduced to a ceramic base 2 in. × 2 in. The experiment did not attempt to achieve the maximum reduction possible, but the board still contained 24 resistors in values up to 121 kΩ, nine capacitors of 10,000 pF and plug-in arrangements for six transistors and nine diodes. In this arrangement, the density of the passive components is about 275,000 per cubic foot, including the 0·05-in. thick base.

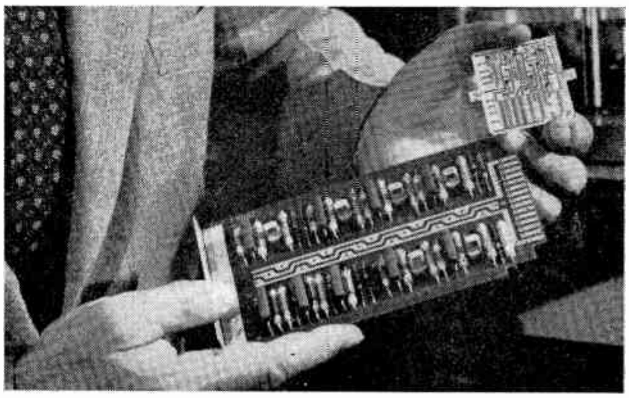


Illustration shows comparison of conventional printed-circuit card with the ceramic circuit card using sputtered resistors and capacitors. The small card incorporates all the components appearing on the conventional card except the diodes

Power in an Angle-Modulated Wave

By W. C. Vaughan, M.B.E., B.Sc., Ph.D., M.I.E.E.

It is commonly accepted that no net power is involved in the process of modulating the angle of a sinusoidal carrier wave by a sinusoidal signal. Indeed, one of the advantages claimed for frequency and phase modulation is that these are methods which, in contrast with amplitude modulation, operate at a constant power level. One might therefore expect that most textbooks treating the theory of modulation would include at least some indication of the proof of this assertion. It is, however, almost always given as an unsubstantiated statement of fact. To conclude, as is sometimes done, that the modulator supplies no power because it does not alter the amplitude of the modulated wave is erroneous, since to do so takes no account of the modification to the shape of the wave.

In a search for information on this subject, the writer has found only one textbook which considers the power aspect of angle modulation. For an essential part of the treatment, however, this refers the reader to "A Course of Modern Analysis' by Whittaker and Watson. Even in this authoritative mathematical treatise the desired information is still elusive, for formal proof is set as an exercise for the student. Fortunately the authors give a hint regarding the method.

The practice of referring readers to advanced works on pure mathematics is not uncommon in communications-engineering textbooks. Although this is often unavoidable when the mathematical analysis is difficult or tedious, it is frequently frustrating to the assiduous student who may not find mathematical texts the easiest of reading. In the case of propositions such as that now to be considered, where the proof is short and does not involve unfamiliar mathematical processes, there is much to be said for including the analysis in the engineering textbook. Either of the following methods of estimating the power in an angle-modulated wave would, it is suggested, suffice for this purpose.

Modulation of the phase-angle $(\omega_c t + \psi)$ of a periodic voltage having an instantaneous value

$$v_c = V_c \cos(\omega_c t + \psi) \qquad \ldots \quad \ldots \quad \ldots \quad (1)$$

by a sinusoidal signal of amplitude V_m and a frequency $\omega_m/2\pi$ is expressed in the relation

$$v = V_c \cos(\omega_c t + m \sin \omega_m t) \qquad \ldots \quad \ldots \quad (2)$$

where m is a pure number termed the modulation index. Equation (2) describes frequency modulation by a signal wave

$$v_m = V_m \cos \omega_m t$$

if m is proportional to V_m/ω_m. Alternatively, it describes phase modulation by a signal wave,

$$v_m = V_m \sin \omega_m t$$

if m is proportional to V_m but independent of ω_m. If, as it is claimed, no power is expended in the modulation process, the average power developed in a resistive load R when the voltage v is applied across it must be the same as that due to the unmodulated sinusoidal voltage v_c, viz., $V_c^2/2R$.

Power in the Modulation Products

The function on the right-hand side of Equ. (2) may be expanded to show the equivalence of the modulated wave to a symmetrical pattern of sinusoidal modulation products centred on the carrier frequency $\omega_c/2\pi$. These sidebands are spaced about the centre component at intervals of frequency which are multiples of the modulation frequency $\omega_c/2\pi$. In both phase and frequency modulation the amplitudes of the sideband components are proportional to integral-order Bessel functions of the modulation index m. The centre component has an amplitude proportional to $J_0(m)$. These modulation products are expressed in the familiar mathematical relation,

$$\begin{aligned} v_c &= V_c \cos(\omega_c t + m \sin \omega_m t) \\ &= V_c \{J_0(m) \cos \omega_c t + J_1(m) [\cos(\omega_c + \omega_m)t \\ &\qquad - \cos(\omega_c - \omega_m)t] \\ &\qquad + J_2(m) [\cos(\omega_c + 2\omega_m)t \\ &\qquad + \cos(\omega_c - 2\omega_m)t] \\ &\qquad + J_3(m) [\cos(\omega_c + 3\omega_m)t \\ &\qquad - \cos(\omega_c - 3\omega_m)t] \\ &\qquad + \ldots\ldots\} \qquad \ldots \quad (3) \end{aligned}$$

The average power supplied by a sinusoidal voltage $V \cos \omega t$ to a resistive load R in any interval of time τ is given by,

$$\begin{aligned} \overline{W} &= \frac{1}{\omega\tau R} \int_0^\tau [V \cos \omega t]^2 \, d(\omega t) \\ &= \frac{V^2}{2R} \left[1 + \frac{\sin 2\omega\tau}{2\omega\tau}\right] \end{aligned}$$

It follows from this relation that if τ greatly exceeds the period $2\pi/\omega$, $\overline{W}$ tends towards $V^2/2R$. The pair of nth-order sidebands described in the expression on the right-hand side of Equation (3) will therefore contribute to the load, in any interval which is long compared with the frequency period of the member of lower frequency, an average power $[V_c J_n(m)]^2/R$ To take account of the infinite spread of the sideband spectrum, it is necessary to average over an indefinitely long time interval to obtain the total contribution of all the modulation products. The average power then tends towards a value,

$$\overline{W}_\infty = \frac{V_c^2}{2R} \left\{ [J_0(m)]^2 + 2([J_1(m)]^2 + [J_2(m)]^2 + \ldots) \right\} \qquad \ldots \quad (4)$$

Fortunately, it is a simple matter to show that the extremely complicated function,

$$\{[J_0(m)]^2 + 2\sum_{n=1}^{\infty}[J_n(m)]^2\}$$

has unit value. To do this, use is made of the expansion of the exponential function $\exp.\left[\frac{m}{2}\left(x-\frac{1}{x}\right)\right]$ viz.,

$$\begin{aligned}
\exp.\left[\frac{m}{2}\left(x-\frac{1}{x}\right)\right]
&= \left\{\left[1+\frac{mx}{2}+\frac{1}{2!}\left(\frac{mx}{2}\right)^2+\ldots\right]\times\left[1-\frac{m}{2x}+\frac{1}{2!}\left(\frac{m}{2x}\right)^2\ldots\right]\right\}\\
&= 1-\left\{\left[\frac{1}{1!1!}\left(\frac{m}{2}\right)^2+\frac{1}{2!2!}\left(\frac{m}{2}\right)^4\ldots\right]\right.\\
&\quad + x\left[\frac{1}{1!}\left(\frac{m}{2}\right)-\frac{1}{1!2!}\left(\frac{m}{2}\right)^3+\frac{1}{2!3!}\left(\frac{m}{2}\right)^5\ldots\right]\\
&\quad + x^2\left[\frac{1}{2!}\left(\frac{m}{2}\right)^2-\frac{1}{1!3!}\left(\frac{m}{2}\right)^4+\frac{1}{2!4!}\left(\frac{m}{2}\right)^6\ldots\right]\\
&\quad + \ldots\ldots\\
&\quad - \frac{1}{x}\left[\frac{1}{1!}\left(\frac{m}{2}\right)-\frac{1}{1!2!}\left(\frac{m}{2}\right)^3+\frac{1}{2!3!}\left(\frac{m}{2}\right)^5\ldots\right]\\
&\quad + \frac{1}{x^2}\left[\frac{1}{2!}\left(\frac{m}{2}\right)^2-\frac{1}{1!3!}\left(\frac{m}{2}\right)^4+\frac{1}{2!4!}\left(\frac{m}{2}\right)^6\ldots\right]\\
&\quad \left.+\ldots\right\}\\
&= \left\{J_0(m)+\left(x-\frac{1}{x}\right)J_1(m)+\left(x^2+\frac{1}{x^2}\right)J_2(m)\ldots\right\} \qquad \ldots (5)
\end{aligned}$$

Similarly, expansion of $\exp.\left[-\frac{m}{2}\left(x-\frac{1}{x}\right)\right]$ gives,

$$\begin{aligned}
\exp.\left[-\frac{m}{2}\left(x-\frac{1}{x}\right)\right]
&= \left\{J_0(m)-\left(x-\frac{1}{x}\right)J_1(m)+\left(x^2+\frac{1}{x^2}\right)J_2(m)\ldots\right\} \qquad \ldots (6)
\end{aligned}$$

Now the product of $\exp.\left[\frac{m}{2}\left(x-\frac{1}{x}\right)\right]$ and $\exp.\left[-\frac{m}{2}\left(x-\frac{1}{x}\right)\right]$ is unity, so that if (5) and (6) are multiplied together it follows that,

$$\begin{aligned}
1 &= \left\{\left[J_0(m)+\left(x-\frac{1}{x}\right)J_1(m)+\left(x^2+\frac{1}{x^2}\right)J_2(m)\ldots\right]\right.\\
&\quad \left.\times\left[J_0(m)-\left(x-\frac{1}{x}\right)J_1(m)+\left(x^2+\frac{1}{x^2}\right)J_2(m)\ldots\right]\right\}\\
&= \left\{J_0(m)\,[J_0(m)+\text{terms in } x]\right.\\
&\quad + J_1(m)\,[2J_1(m)+\text{terms in } x]\\
&\quad + J_2(m)\,[2J_2(m)+\text{terms in } x]\\
&\quad \left.+\ldots\right\} \qquad \ldots\ \ldots\ \ldots\ \ldots (7)
\end{aligned}$$

Since the left-hand side of Equ. (7) is independent of x, the sum of all terms containing x on the right-hand side must be zero. Accordingly,

$$1 = \{[J_0(m)]^2 + 2\sum_{n=1}^{\infty}[J_n(m)]^2\}$$

It is now evident that $\overline{W}_\infty$ tends to a value $V_c^2/2R$ as the time interval is indefinitely extended.

The frequency spectrum of a sinusoidal carrier wave, angle-modulated by a sinusoidal signal, is a complicated pattern of modulation products distributed over an infinite frequency band. The frequencies of these components are dependent on the modulation frequency; their amplitudes are functions of the modulation index m. Such patterns have one feature in common, namely that with rising order there is a progressive decrease in amplitude of components whose orders are greater than the modulation index. In fact, the practicability of communication by angle-modulation methods depends upon this. Experience has shown that the exclusion of components having orders greater than $(m + 4)$ does not prejudice reconstruction of a tolerable replica of the modulation signal at the receiver. The components in the frequency band $1/2\pi\ [\omega_c \pm (m+4)\ \omega_m]$ therefore suffice for satisfactory communication. If m is a whole number, the excluded component having the largest amplitude will then be that of order $(m + 5)$. Examination of a table of Bessel functions will confirm that when m is less than 20, the ratio $\frac{J_{m-1}(m)}{J_{m+5}(m)}$ always exceeds 20. This means that, with this bandwidth limitation, there will be at least one transmitted component whose amplitude is more than twenty times that of the most prominent excluded component, if the modulation index is kept below 20. In these circumstances, extremely small power is associated with each excluded component. It therefore seems likely that the power dissipated by the modulation products within the pass-band will not differ substantially from $\overline{W}_\infty$. This power is carried by a group of sinusoidal components which have, in practice, frequencies very much higher than the modulation frequency $\omega_m/2\pi$. It may be inferred, then, that the average power delivered to the load in a time interval comparable with the modulation period is very nearly $V_c^2/2R$. Negligible power is therefore required of the modulator.

Average Power in the Modulation Period

The conclusion drawn from the above analysis, although reasonable, is not completely satisfying since it assumes the sum of a rapidly converging series of very small terms to be a negligible fraction. Rigorous justification of this assumption would be difficult. The following method, which yields an exact expression for the average power dissipated in an interval equal to the modulation period, is more elegant. It assumes the carrier frequency to be a multiple of the modulation frequency. The consequent slight loss of generality cannot be regarded as serious, since it is hardly likely that the modulation process will be dependent on fortuitous choice of the modulation frequency. Moreover, the conclusion from the second method supports that derived from the first.

Suppose that $\omega_c = K\omega_m$, where K is an integer. Then since $\cos\,[K\,\omega_m t + m \sin \omega_m t]$

$$= \cos\,[K\,(\omega_m t + 2\pi) + m \sin\,(\omega_m t + 2\pi)]$$

for all values of t, the function representing v in Equation (2) is seen to have a period $\omega_m/2\pi$. In these circumstances, the instantaneous voltage v may be represented by,

$$v = V_c \cos\,(K\theta + m \sin\theta) \qquad \ldots\ \ldots\ \ldots (8)$$

where $\theta = \omega_m t$. During each modulation cycle, this voltage will dissipate in a resistance R an average power,

$$\overline{W}_m = \frac{V_c^2}{2\pi R}\int_0^{2\pi} \cos^2 (K\theta + m \sin \theta)\, d\theta$$

$$= \frac{V_c^2}{2\pi R}\int_0^{2\pi} \frac{1}{2}\,[1 + \cos 2\,(K\theta + m \sin \theta)]\, d\theta$$

$$= \frac{V_c^2}{2\pi R}\left[\pi + \frac{1}{2}\int_0^{2\pi} \cos 2\,(K\theta + m \sin \theta) d\theta\right] \quad (9)$$

When n is an integer, the definite integral

$$\frac{1}{2\pi}\int_0^{2\pi} \cos\,(n\phi + x \sin \phi)\, d\phi$$

defines, as will be shown in Appendix 1, the Bessel function[1] $(-1)^n J_n(x)$. The result at (9) may therefore be written as

$$\overline{W}_m = \frac{V_c^2}{2R}\left[1 + \frac{1}{2} J_{2K}(2m)\right] \quad \ldots \quad \ldots \quad (10)$$

since $2K$ is an even number.

Bessel functions always have values less than unity; their values fall rapidly with ascending order for $n > x$. In practice, it is unlikely that the factor K will be less than a hundred times the modulation index m. As a consequence, the Bessel function $J_{2K}(2m)$ will always be a very small fraction. Thus $\overline{W}m$, the average power delivered to the resistive load in a time interval equal to the modulation period $2\pi/\omega_m$ will differ inappreciably from $V_c^2/2R$. The average power supplied by the modulator during this interval will therefore be substantially zero.

APPENDIX 1

The definite integral

$$\int_0^{2\pi} \cos\,(n\phi + x \sin \phi)\, d\phi$$

may be written as

$$\int_0^{2\pi} \cos n\phi \cos\,(x \sin \phi)\, d\phi - \int_0^{2\pi} \sin n\phi \sin\,(x \sin \phi)\, d\phi$$

These two integrals may be evaluated from the following well-known expansions in Bessel functions,

$$\cos\,(x \sin \phi) = J_0(x) + 2\,\{J_2(x) \cos 2\phi + J_4(x) \cos 4\phi + \ldots\}\ldots \quad \ldots\ (11)$$

$$\sin\,(x \sin \phi) = 2\,\{J_1(x) \sin \phi + J_3(x) \sin 3\phi + \ldots\}\ldots \quad (12)$$

If (11) and (12) are multiplied respectively by $\cos n\phi\, d\phi$ and $\sin n\phi\, d\phi$ and integrated term by term between 0 and 2π, then

$$\int_0^{2\pi} \cos n\phi \cos\,(x \sin \phi)\, d\phi = J_0(x)\int_0^{2\pi} \cos n\phi\, d\phi + 2J_2(x)\int_0^{2\pi} \cos n\phi \cos 2\phi\, d\phi + 2J_4(x)\int_0^{2\pi} \cos n\phi \cos 4\phi\, d\phi + \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

and
$$\int_0^{2\pi} \sin n\phi \sin\,(x \sin \phi)\, d\phi = 2J_1(x)\int_0^{2\pi} \sin n\phi \sin \phi\, d\phi + 2J_3(x)\int_0^{2\pi} \sin n\phi \sin 3\phi\, d\phi + 2J_5(x)\int_0^{2\pi} \sin n\phi \sin 5\phi\, d\phi + \ldots \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

[1] This integral is closely related to that encountered in the Fourier analysis of the beam current of a velocity-modulated tube.

The integrals on the right-hand side of Equ. (13) have the general form,

$$\int_0^{2\pi} \cos n\,\phi \cos k\phi\, d\phi$$

where $k = 0, 2, 4$, etc. This may be evaluated as follows,

$$\int_0^{2\pi} \cos n\phi \cos k\phi\, d\phi = \frac{1}{2}\int_0^{2\pi}\left[\cos\,(n + k)\,\phi + \cos\,(n-k)\,\phi\right] d\phi$$

$$= \frac{1}{2}\left[\frac{\sin\,(n+k)\,\phi}{(n+k)} + \frac{\sin\,(n-k)\,\phi}{(n-k)}\right]_0^{2\pi}$$

If n is zero or is a whole number and is not equal to k, the terms in the bracket disappear at both limits. When n is equal to k and the latter is not zero, the general term becomes,

$$\int_0^{2\pi} \cos^2 n\phi\, d\phi = \frac{1}{2}\left[\phi + \frac{\sin 2n\phi}{2n}\right]_0^{2\pi} = \pi$$

If, however, n and k are both zero, the general term reduces to $\int_0^{2\pi} d\phi$ which has a value 2π. Consequently,

$$\int_0^{2\pi} \cos n\phi \cos\,(x \sin \phi)\, d\phi = \begin{cases} 2\pi J_0(x) & \text{when } n = 0 \\ 0 & \text{when } n = 1, 3, 5, \text{etc.} \\ 2\pi J_n(x) & \text{when } n = 2, 4, 6, \text{etc.} \end{cases} \quad \ldots\ (15)$$

In a precisely similar manner it can be shown that,

$$\int_0^{2\pi} \sin n\phi \sin\,(x \sin \phi)\, d\phi = \begin{cases} 0 & \text{when } n = 0 \\ 2\pi J_n(x) & \text{when } n = 1, 3, 5, \text{etc.} \\ 0 & \text{when } n = 2, 4, 6, \text{etc.} \end{cases}$$

The results at (15) and (16) may now be combined to give

$$\frac{1}{2\pi}\left\{\int_0^{2\pi} \cos n\phi \cos\,(x \sin \phi)\, d\phi - \int_0^{2\pi} \sin n\phi \sin\,(n \sin \phi)\, d\phi\right\}$$

$$= \frac{1}{2\pi}\int_0^{2\pi} \cos\,(n\phi + x \sin \phi)\, d\phi = (-1)^n J_n(x)$$

when $n = 0, 1, 2$, etc.

FORTHCOMING COURSES

Lectures on Solid State Physics

A course of ten special lectures dealing with modern developments in solid state physics is to be held at the Enfield Technical College, Queens Way, Enfield.

The lectures, which will be delivered by members of Mullard Research Laboratories, will begin on Monday 5th October 1959. The course is suitable for physics graduates or others with comparable background knowledge and the fee is £1.

Application forms may be obtained from the Head of the Department of Pure and Applied Science at the College.

I.E.E. Part III and I.Mech.E. Part C Courses

South East London Technical College are running a six-months' full-time course leading up to the Institution of Electrical Engineers Part III examination and/or the Institution of Mechanical Engineers Part C examination. Lectures start on 19th October 1959 and end in May 1960 and will be held Monday to Friday 9.0 a.m.–1.15 p.m. and 2–5 p.m.

The London fee for the complete course is £17.

Applications to attend the course should be sent to the Head of Electrical Engineering and Applied Physics Department, Lewisham Way, London, S.E.4.

TWO CO-ORDINATE COLOUR

Some long time ago (February and March 1957), I attempted an analysis of the processes of colour mixing and colour reproduction. Not at any learned level, of course, but simply to try and work out the connection between the principles of colour-television transmission and Newton's colour circle. The conclusion seemed to be that there are three attributes to colour, which require three separate pieces of information for their transmission or recording, but that in many cases we can get away with two. Considering why this should be so, one wonders whether the three attributes are interwoven rather than independent; whether one of them is rated less highly than the other two; or whether we are simply content with an imperfect specification. And the answer would appear to be that all three of these reasons are relevant. None of them, however, tells us anything about the nature of the colour-vision process.

The three characteristics of colour are:

Subjective, describing a sensation: hue, luminosity, and saturation;

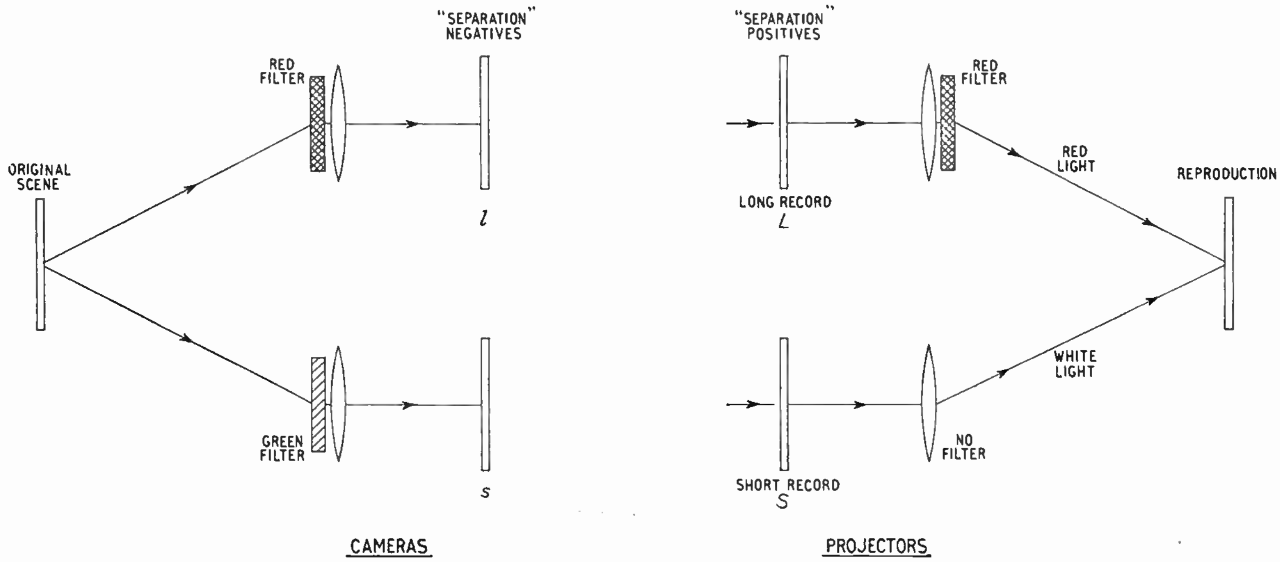
Objective, specifying a stimulus: dominant wavelength, luminance, purity.

In Newton's colour-circle, the second of these two factors is left out of account; it is a plane diagram, and two co-ordinates locate a point in a plane. The same is true of the colour triangle and the C.I.E. chromaticity diagram. Three co-ordinates are in fact used, R, G, B, or X, Y, Z, but this is an algebraic device to avoid negative values, on the principle that the eye can only add; for they are not independent. In all these diagrams chrominance (the colour factor) has been effectively separated from luminance (the intensity factor). The same kind of thing (though without the restriction of purely positive quantities) is done in preparing colour information for television transmission. The luminance signal Y is separated from the chrominance signals I and Q; though there are additional subtleties in that some chrominance is in fact packed into Y, and the minimum of information is allotted to Q, which makes do with a niggardly bandwidth on the score that people can manage without it in regions of fine detail.

I said the same kind of thing but there is, in fact, a vast difference in principle if you try to bring the eye into it; for the diagrams dissociate luminance from chrominance, and the signals insert some chrominance into luminance. There seems the basis for a profitable slogan here, when somebody gets round to promoting these tergents of mine; "takes chrominance from luminance" should send you all flocking to exchange your valuable vouchers at the tallow-chandler's.

What has brought this up again is the recent work of Dr. Edwin Land on what can be called two co-ordinate colour. This has caused such a stir that I felt I ought to say something about it. He has described his experiments admirably in an article in *Scientific American* (May 1959), which I expect many of you will have read. I first met the work in a more popular account by Francis Bello in *Fortune*, also for May 1959; and a review of the work,

Fig. 1. *Colour reproduction using red light and white light only*



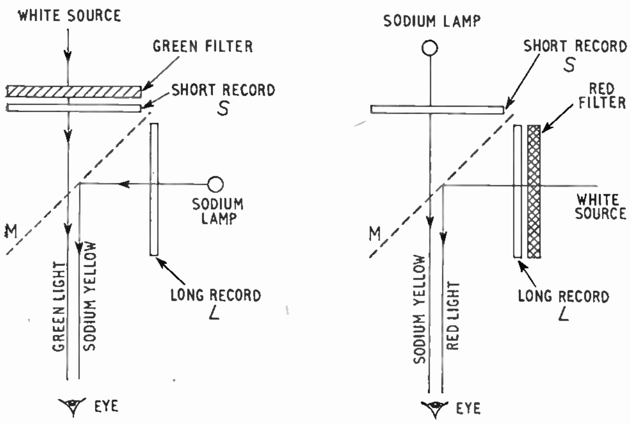


Fig. 2. *Colour reproduction using lights of two colours only: (a) short-record in green, long-record in sodium yellow; (b) short-record in sodium yellow, long-record in red. M is a lightly silvered 'semi-mirror'*

with its possible implications, was given in *The Manchester Guardian* for 14th July 1959.

Before getting on to this, however, a nearly irrelevant analogy in the matter of co-ordinates or parameters intrudes itself. If you want to be able to write down the behaviour of a circuit completely, you must know three things about it, namely R, L, and C. If you are only interested in the proportion of the incident power that it dissipates, as a fraction of the power that a purely resistive circuit of the same impedance would dissipate in the same circumstances, then X and R alone suffice. It all depends on whether you are interested in absolute power dissipation, or merely in power factor.

Land's Experiments

Since these are fully described in the *Scientific American* article, and beautifully illustrated there, I need only summarize them briefly here.

1. *The Red-White Experiment*

Two photographs of the same scene are taken through red and green filters respectively. The black-and-white negatives, l and s, are printed out to give black-and-white positives L and S, the long-wavelength-record and short-wavelength-record. Projected in register, using red light for L and white light for S (Fig. 1), a satisfactory full-colour picture is seen. This is, as Land mentions, a rediscovery of an effect tried out in the early days of colour cinematography.

2. *The Two-Colour Experiments*

With the same two records, viewing S through a green filter and L in sodium yellow light, superposing the two pictures (Fig. 2) reproduces full colour; the same result is found if S is viewed with sodium yellow, and L using a red filter. No spectral light apart from sodium yellow and the transmission band of the filter is present in either of these two cases [Fig. 3(a)].

3. *Monochromator Experiments*

These are essentially a repetition of (2), using monochromatic light. Within limits, provided that S is illuminated by the shorter wavelength, and L by the longer, it does not matter what the actual wavelengths are; a good colour picture is seen. Wavelengths between 4,300 and 6,150 Å can function either in the capacity of shorter or longer wavelength. The full range of colours is seen only if both lie within 5,750–6,150 Å, but *they can be as close as* 100 Å *to one another*. A reversal effect is shown in the short-wave part of the spectrum, below 4,900 Å, where the shorter of the two wavelengths must illuminate L and the longer of the two serve for S. These results are summarized in Fig. 3(b).

4. *Quantitative*

Whatever the two stimuli are, the action of the two record pictures is in each case to vary the percentage of

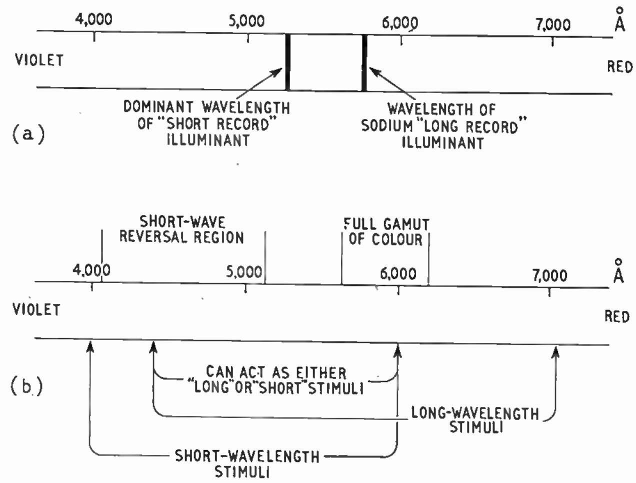


Fig. 3. *(a) Spectral lights used in Fig. 2 (a); (b) ranges of the spectrum in which the long and short wavelength stimuli can lie*

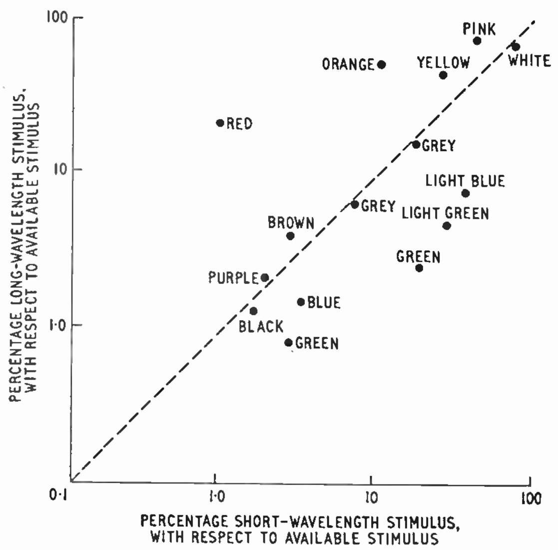


Fig. 4. *The quantitative factor. For each stimulus, the operative quantity is the intensity expressed as a percentage of the maximum available stimulus. The colour observed is determined by the two percentages; the graph (plotted logarithmically) shows the distribution black-grey-white along the 45-degree line, with red, orange, yellow (the 'warm' colours) above it, and blue and green (the 'cold' colours) below it*

each stimulus at different parts of the pattern. If I have understood the argument correctly, Land attributes this to two acts of evaluation by the eye. The first is an appreciation of the percentage of each stimulus in terms of the maximum available—assessing what we will call P_S and P_L. The second is a response to the combination of P_L and P_S. Fig. 4 illustrates the quantitative results obtained. Note that the stimuli themselves simply have to fulfil the conditions of being 'long' and 'short' relative to one another. In discussing this, Land goes back to the old Newtonian idea of "taking moments and finding the centre of gravity", and suggests that the eye acts in some such way in performing the evaluations. He makes one rather interesting point, that we enjoy a very wide waveband for the transmission of colour information; and that in a world in which pigments were designed to act in keeping with the minimum needs, the full gamut of colour sensations could be furnished from a very much narrower visible spectrum.

Goethe

But you will doubtless be hearing a great deal more about Land's experiments and their interpretation as time goes on. My preoccupation at the moment is with the mystic number two, and it crops up in what might be called the two co-ordinate theory of colour published by Goethe in 1808. A paper on this, given by M. H. Wilson and R. W. Brocklebank at the 1958 Physical Society Exhibition and printed in the Society's Yearbook for 1958, shows what can be done with edge spectra in producing subjective colour effects. If we have either a white-illuminated slit, or a brightly-illuminated white line on a black card, and view it through a prism [Fig. 5(a)] the normal sequence of spectral colours is seen, with violet nearest the edge of the prism, since this is a virtual-image spectrum. With a black band on a white card, however, the upper edge as seen through the prism has red, orange, yellow (reading down), and the lower edge violet, blue, cyan (reading up), the complementaries to the colours of the upper edge [Fig. 5(b)]. Goethe considered that these 'edge spectra' were the really fundamental colour effect, and that Newton's spectrum, that of Fig. 5(a), was a secondary effect given by the overlapping of the edge spectra from the two sides of the slit. By observing black-and-white patterns through a prism, Goethe concluded that there were two attributes only to a colour—light and darkness.

Wilson and Brocklebank show the physical basis for this idea. They plot out the edge-spectra colours on the C.I.E. chromaticity diagram, and show that every possible chromaticity can be produced from the two-edge spectra, or by additive combinations of them. They also demonstrate the wide variety of colours obtained using black-and-white patterns and a prism. Their most interesting picture is a subjective record showing a wide colour range and including pink, yellow, and green, and obtained by projecting in register three separate transparencies through *three different blue filters*. It seems to me, on my very recent acquaintance with them both, that there are some common features between Goethe's ideas and Land's experiments, though at a very general level. Goethe thought of colour as the interplay of light and darkness; Land's experiments seem to translate this into another dimension, as the interplay of two lots of interplays between light and darkness. But there is one important feature about Goethe's ideas which we might look into further. He did not regard 'white' as a sort of artefact built up of Newton's spectrum. To him, *white was a colour*. I think he would have considered that Land was doing things the hard way, using, say, yellow-and-black with green-and-black when he himself managed quite well with white-and-black. Do not misunderstand me here. I am speaking quite seriously, and with sufficient innocence of the subject to be duly guarded and

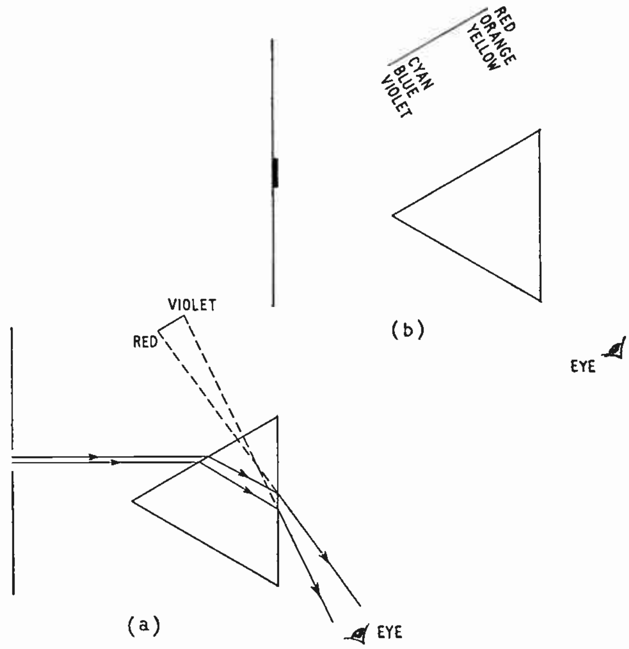


Fig. 5. *Virtual-image spectra observed by eye through a prism. (a) Newton's spectrum, from a white-illuminated slit or a white stripe on a black card; (b) the two edge-spectra from the edges of a black stripe on a white card*

circumspect. For the last piece of two co-ordinate work I want to discuss, that of Ragnar Granit, faces the problem as to whether white is perceived as a sensation in its own right, and concludes that it is.

Dominator and Modulator

Granit's 1945 Thomas Young Oration (*Proceedings of the Physical Society*, Vol. 57, p. 448, November 1945) describes experiments on the responses obtained by inserting microelectrodes in the living eye of rat, guinea-pig, snake, and cat. The electrode picked out a single nerve fibre proceeding from a ganglion fed by several retinal receptors, and the intensity of the response to light was measured by the rate of generation of electrical impulses. As a rule, with the eye light-adapted, the variation of response with wavelength followed a very similar pattern to that of the relative luminosity curve of the human eye.

The point here is that, whatever the wavelength of the light, the basic signal sent to the brain is a series of impulses, the number per second conveying only one piece of information, which Granit interpreted as the sensation of brightness or whiteness. No evidence for three different and separate fundamental response curves was found; but fibres were found which had a preferential response in one region of the spectrum, not unlike the response curves found subjectively in human observers. There appeared to be four types, with peaks in the red, yellow, green, and blue. But these were not sharers in generating the 'whiteness' signal; their function was to superpose an additional signal upon it. Granit called the generators of the whiteness signal *dominators*; and the selective generators *modulators*. Here, the two co-ordinate aspect has perhaps involved us in a certain amount of difficulty; we are just about where we came in, with an eye which according to the physiologists, can separate luminance from chrominance —or rather, which responds to 'white' as a sort of carrier wave, and accepts 'colour' as a signal superposed on this. Unless I have miscounted, we seem to have lost one of the co-ordinates of chrominance somewhere. Or else cheated in putting all the modulators together as one signal generator. Although "some kind of local sign" is as near as Granit will go towards describing their action.

Conclusion

To say that all this has me going round in colour-circles is a mild understatement. We know what the chromaticity diagram does; it enables us to compound and analyse things of the same kind, namely stimuli of the same kind, expressed in absolute energy terms if need be. Except that it involves the relative luminosity curve of the eye, it is as objective as algebra. Land's experiments suggest that there are subjective processes at work which involve comparison and assessment, and that though the stimuli may be of the same kind their operation is not a simple addition effect. Granit's experiments suggest that there are two different kinds of response to stimuli. Goethe, on the other hand, regarded the sense of vision as something much more than a mere response to stimuli; to him it was an active and creative sense. It may well be that all these different approaches will eventually converge.

Concertina Phase-Splitter – 2

LOW FREQUENCIES

From the discussion in Part 1 it will be evident that valve capacitances normally have appreciable effect upon the concertina phase-splitter only at high audio frequencies. Their effect is usually negligible under some 20 kc/s. The analysis of the basic circuit, given in Part 1, is thus highly accurate at low frequencies.

The condition for the anode output to be equal in amplitude and opposite in phase to the cathode output is merely that the anode load impedance Z_a shall be equal to the cathode load impedance Z_k. These impedances purport to be resistances at low frequencies, for then shunt capacitances are negligible. However, even if the resistances are made precisely equal, Z_a and Z_k usually become unequal at a sufficiently low frequency because Z_a contains the impedance of the power supply or anode decoupling circuit whereas Z_k does not.

As it is normally applied, therefore, the concertina phase-splitter becomes unbalanced at very low frequencies. It is important, therefore, to investigate the magnitude of this effect. With anode decoupling the circuit takes the form shown in Fig. 5. We assume that the impedance of the power supply is negligible compared with R_d, and we can then write

$$Z_k = R_k$$

$$Z_a = R_a + \frac{R_d}{1 + j\omega T_d}$$

where $T_d = C_d R_d$.

For balanced outputs at the higher frequencies it is necessary to have $R_k = R_a$. In all that follows, therefore, we shall take $Z_k = R_a$. From equation (3) of Part 1

$$\frac{-v_a}{v_k} = \frac{Z_a}{Z_k} = 1 + \frac{R_d/R_a}{1 + j\omega T_d} \quad \ldots \quad \ldots \quad (24)$$

It is obvious that the unbalance is large at zero frequency. Typically, R_d is of the order of twice R_a and so the anode output becomes about twice the cathode output. The usual concertina phase-splitter is thus useless if the response must be maintained down to zero frequency; however, the circuit can be modified in a manner which will be described later to remove this limitation.

The phase angle between $-v_a$ and v_k can be written from (24) as

$$\theta = \tan^{-1} - \omega T_d \frac{R_d/R_a}{1 + R_d/R_a + \omega^2 T_d^2} \quad \ldots \quad (25)$$

By differentiating and equating to zero we find that θ is a maximum when $\omega T_d = \sqrt{(1 + R_d/R_a)}$. Inserting this value in (25) we get

$$\theta_{max} = \tan^{-1} - \frac{R_d/R_a}{2\sqrt{(1 + R_d/R_a)}}$$

With $R_d = R_a$, a normal minimum value for R_d, the maximum phase angle is $\tan^{-1} -0{\cdot}3535 = -19{\cdot}45°$. With $R_d = 22\ \text{k}\Omega$, $C_d = 8\ \mu\text{F}$, this occurs at a frequency

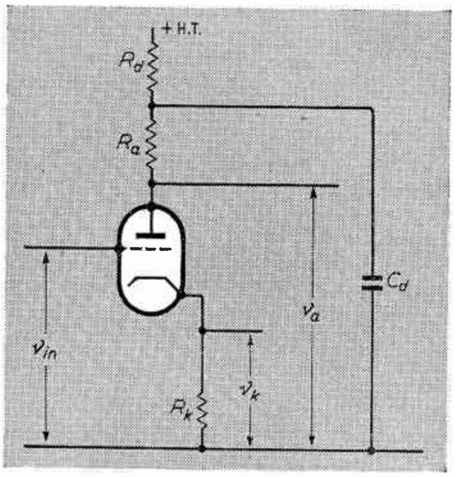


Fig. 5. *Circuit of concertina phase splitter with anode decoupling*

of 1·28 c/s. At 25 c/s we have $\omega T_d = 27{\cdot}6$ and so $\theta = -2{\cdot}07°$. There is thus no difficulty in keeping unbalance negligibly small down to the lowest audio frequency with quite reasonable values of decoupling capacitance. If 2° phase unbalance is considered excessive at 25 c/s it can be reduced almost as much as one wants. This value is for 8 μF only. The use of 100 μF would not be impracticable and would reduce the phase shift to $2/12{\cdot}5 = 0{\cdot}16°$ and shift the 2° frequency to 2 c/s. With such a large capacitance it would usually be possible to reduce R_d and so obtain still less unbalance.

When satisfactory balance is obtained over the working range of frequencies the variations of gain will usually be negligible. This comes about because satisfactory balance demands $Z_a \approx R_k$. If the stage is connected in a feedback loop, however, the performance at frequencies below the working range may be important from the point of view of stability.

One output is v_k, the other is $-v_a$. Both are subject to further amplification and we shall assume that they are equally amplified, and are then combined in a transformer, say. The total output is thus proportional to $v_k - v_a$ and a fraction of this is fed back to the input. Now

$$v_k - v_a = i_a (Z_k + Z_a)$$

since by definition $-v_a = i_a Z_a$.

Equation (2) of Part 1 gives i_a in terms of v_{in} and from these we get, after a little algebra,

$$\frac{v_k - v_a}{v_{in}} = 2A_0 \frac{1 + x + j\omega T_d}{1 + y + j\omega T_d} \qquad \text{.. .. (26)}$$

where $x = R_d/2R_a$

$$y = \frac{R_d}{r_a + R_a(2 + \mu)} = 2xA_0/\mu$$

$$A_0 = \frac{\mu R_a}{r_a + R_a(2 + \mu)}$$

From (26)

$$\phi = \tan^{-1} -\omega T_d \frac{x - y}{(1 + x)(1 + y) + \omega^2 T_d^2} \qquad (27)$$

This is a maximum when

$$\omega T_d = \sqrt{[(1 + x)(1 + y)]}$$

and is

$$\phi_{max} = \tan^{-1} - \frac{x + y}{2\sqrt{[(1 + x)(1 + y)]}}$$

The maximum phase angle is independent of the size of the decoupling capacitor; altering the capacitor value merely changes the frequency at which the maximum angle occurs. Since $y = 2xA_0/\mu$ it follows that if $1 \gg 2A_0/\mu$, which will often be the case, the maximum phase angle is approximately

$$\tan^{-1} - \frac{x}{2\sqrt{(1 + x)}}$$

assuming that x is not much larger than unity. If $x = 1$ (i.e., $R_d = 2R_a$) the angle is $\tan^{-1} -1/2\sqrt{2} \approx 19{\cdot}5°$. If $x = 0{\cdot}5$ (i.e., $R_d = R_a$) the angle is $\tan^{-1} -0{\cdot}25/\sqrt{1{\cdot}5} \approx -11{\cdot}8°$. In this latter case the more accurate formula with $y = 0{\cdot}044$ leads to an angle of $-10{\cdot}35°$, so the simple approximation is normally adequate.

The concertina phase-splitter can thus be seen to introduce appreciable overall phase shift at some very low frequency and may thus affect the stability of a negative-feedback amplifier. The magnitude of the phase shift depends mainly on R_d/R_a and falls as this ratio is reduced.

The obvious way of eliminating low-frequency unbalance is to make $Z_k = Z_a$ by adding cathode components as in Fig. 6. With equal component values in anode and cathode circuits perfect balance is obtained down to zero frequency. Practically, however, it is impossible to match the capacitors closely, for they will normally be electrolytic types and will operate with different polarizing voltages.

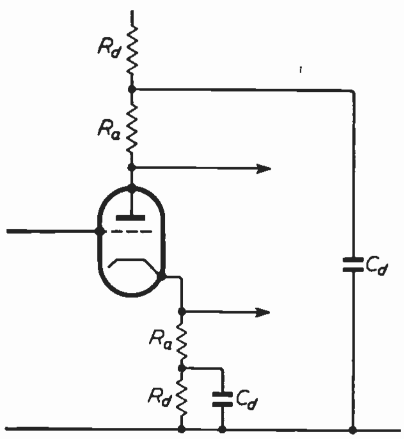
Although this circuit will give proper balance, the impedances of the loads vary with frequency and so the gain varies to some extent. This is readily calculable but it is not worked out here since the circuit is not one likely to have much application.

An alternative way of maintaining balanced outputs down to zero frequency is shown in Fig. 7. The anode output is taken via a frequency-compensated potential divider. This is a well-known technique in other spheres and the requirement is to have

$$R_d/R_a = R_1/R_2 \text{ and } C_dR_d = C_1R_1$$

Additionally it is clearly necessary to have

Fig. 6. *Modified phase splitter with equal anode and cathode loads*



$R_k = R_a R_2/(R_a + R_2)$ for, at high frequencies, the capacitances are effectively short circuits and the anode load comprises R_a and R_2 in parallel and must obviously equal the cathode load R_k.

The ratio of R_2 to R_a is arbitrary and it is usually convenient to make R_2 about 10 times R_a; R_a, R_d and C_d can be chosen in the ordinary way. The other values then follow from the relations given. Thus, for

$$R_a = R_d = 22\ \text{k}\Omega,\ C_d = 8\ \mu\text{F and } R_2/R_a = 10$$

we have:

$$R_2 = R_1 = 220\ \text{k}\Omega,\ R_k = 20\ \text{k}\Omega,\ C_1 = 0{\cdot}8\ \mu\text{F}.$$

Again the gain, as distinct from the balance, is frequency dependent. The impedance of the anode load as a whole increases as frequency falls. For constant input voltage, therefore, the anode current falls somewhat as frequency is reduced and so the output falls. This anode current effect also occurs with Fig. 6, of course, but as in that circuit the voltages are developed across the rising impedances the output rises, and quite considerably.

When the above relations hold, Equ. (2) of Part 1 applies with

$$Z_k = R_k \text{ and } Z_a = R_k\left[1 + \frac{R_d/R_a}{1 + j\omega C_d R_d}\right].$$

At high frequencies the denominator of (2) is $r_a + R_k(2 + \mu)$; at zero frequency it is $r_a + R_k(2 + \mu + R_d/R_a)$. With $r_a = 15\ \text{k}\Omega$, $R_k = 22\ \text{k}\Omega$ and $\mu = 20$ the total is 499 kΩ at high frequencies and 543 kΩ at zero frequency. The drop in gain at zero frequency thus amounts to about 0·7 dB only. The important factors in minimizing the variations are μ, which should be as high as possible, and R_d/R_a, which should be as small as possible.

Although in this way the circuit of Fig. 7 can be utilized satisfactorily down to zero frequency, there is a mean voltage difference between the output terminals. With Fig. 6 this is 50–100 volts; with Fig. 7 it is lower and it can be made zero. This is not usually realised.

For a response effective at zero frequency decoupling becomes impracticable and there is no point in including decoupling components. The circuit can thus take the form of Fig. 8, in which C_2 represents unavoidable capacitance across R_2. If $C_1R_1 = C_2R_2$ the potential divider is frequency compensated at high frequencies.

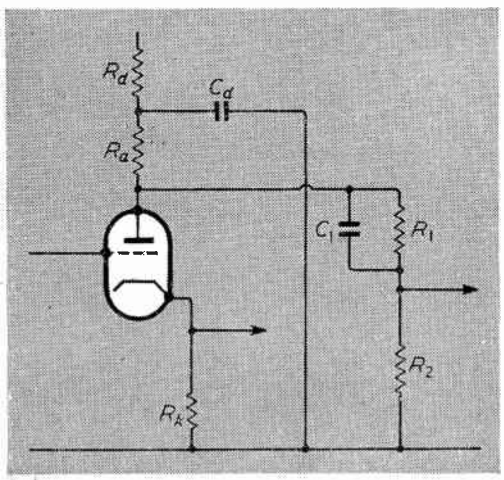


Fig. 7. *Circuit with compensation for the low-frequency effects of anode decoupling*

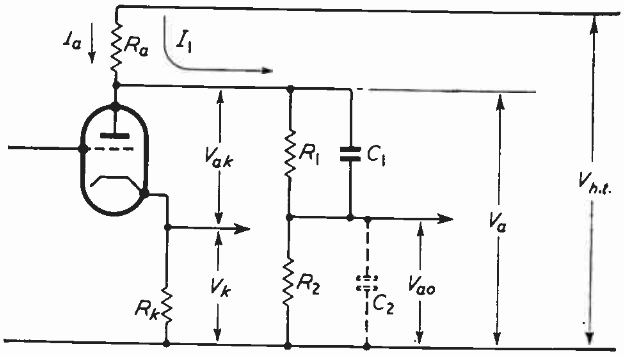


Fig. 8. *Circuit designed not only for equal and opposite signal outputs but for zero potential difference between the output terminals under no-signal conditions*

Using capital letters for no-signal voltages and currents, if the potential difference between the two output terminals is to be zero we must have

$$I_a R_k = I_1 R_2$$

Equality of signal outputs obviously demands

$$R_k = \frac{R_a(R_1 + R_2)}{R_a + R_1 + R_2} \cdot \frac{R_2}{R_1 + R_2}$$

In addition we have

$$V_{ht} = I_a R_a + I_1.(R_a + R_1 + R_2)$$

Normally one will start by knowing the desired values of I_a, V_{ak} and V_{ht}. By inspection of Fig. 8

$$V_{ak} = I_1 R_1$$

and it is convenient to assume an arbitrary value for $I_1/I_a = \alpha$ such as, 0·1. A little algebra then gives

$$\alpha = \frac{I_1}{I_a} = \frac{R_k}{R_2} = \frac{V_{ak}/I_a}{R_1}$$

$$V_{ht}/I_a = 2R_a$$

whence

$$R_a = V_{ht}/2I_a$$

$$R_1 = V_{ak}/\alpha I_a = V_{ak}/I_1$$

$$R_2 = \frac{R_a(1 - \alpha) - \alpha R_1}{\alpha}$$

$$R_k = \alpha R_2$$

As an example, let $V_{ht} = 300$ V, $V_{ak} = 100$ V, $I_a = 2$ mA, $\alpha = 0{\cdot}1$. Then $R_a = 75\ \text{k}\Omega$, $R_2 = 175\ \text{k}\Omega$, and $R_k = 17{\cdot}5\ \text{k}\Omega$. The no-signal output voltage is $I_aR_k = 35$ V and so the anode–earth voltage is 135 V; as a check this should equal the h.t. voltage less the drop across R_a, or $300 - 75 \times 2{\cdot}2 = 135$ V.

The anode load of the valve is $75 \times 675/750 = 67{\cdot}5$ kΩ. The signal voltage at the anode is thus $67{\cdot}5/17{\cdot}5 = 3{\cdot}85$ times as great as at the cathode. At the anode output terminal it is reduced by the factor $175/675 = 1/3{\cdot}85$ times by the potential divider.

The circuit is not one which is likely to be of great utility for the mean potential between the output terminals is obviously affected by h.t. and bias voltage variations. It has not previously been described, as far as the writer knows, and it has seemed worth while putting its characteristics on record.

Signal Flow Graphs

APPLICATION TO LINEAR CIRCUIT ANALYSIS

By **R. F. Hoskins M.Sc.** *

SUMMARY. *This paper is an account of the Signal Flow Graph technique of circuit analysis developed by S. J. Mason in two* "Proc. I.R.E." *papers. This technique can be applied with advantage to transistor-amplifier circuits, and the treatment in this article enables the results to be directly related to Bode's classical feedback theory. Detailed accounts are given of the formal operations necessary to reduce flow graphs to simpler forms, to which Mason's gain formula can be easily applied.*

In a great many cases, the steady-state analysis of linear networks involves the calculation of the ratio of output signal to input signal. Thus we may require the transfer impedance of a network, which is simply the ratio of output voltage to input current; alternatively calculation of the gain of a voltage amplifier amounts to determining the ratio of output voltage to input voltage, and so on. In general, this type of problem is tackled by expressing the input and output signals in terms of intermediate voltages and currents existing in various parts of the network. This gives a system of linear simultaneous equations which must be solved to obtain the required ratio.

A signal flow-graph is a topological model of such a system of equations and, by manipulation of the graph, it is possible to obtain a solution of the equations without performing the usual algebraic operations. The formal algebra of such flow-graphs, together with a general method of solution, is presented by Mason in Refs. 1 and 2. The treatment in this paper differs from that of Mason in that a more algebraic approach is used.

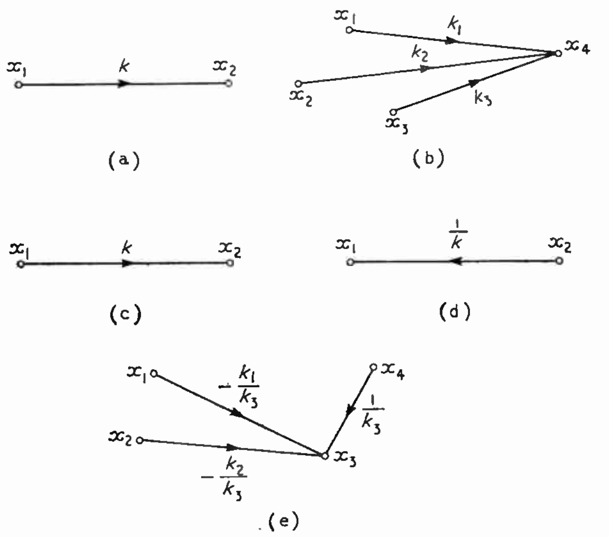


Fig. 1. *Simple graph (a) and more complex arrangement, (b) showing three paths to x_4. A graph showing x_2 in terms of x_1 appears at (c) and the inverse of this to show x_1 in terms of x_2 at (d). A reversal of the direction of the x_3, x_4 branch of (b) leads to (e)*

Basic Principles of Flow-Graphs

Currents and voltages are represented as nodes, x_i, in the graph, while the linear relations existing between them appear as directed branches connecting the nodes. Thus, the equation $x_2 = k.x_1$ is to be expressed by the graph shown in Fig. 1 (a). Here, x_1 may be a current and x_2 a voltage, in which case the branch k will have the dimensions of an impedance. If, on the other hand, x_1 and x_2 are both voltages, for example, k will represent a linear amplifier. More generally, the equation: $x_4 = k_1 x_1 + k_2 x_2 + k_3 x_3$, is to be represented by the configuration shown in Fig. 1 (b); i.e., each node is to be the algebraic sum of all contributions which enter it from other nodes.

It should be noted at this stage that all branches are *directed*: that is to say, we cannot reverse the direction of a given branch without making appropriate alterations in the form of the linear relations which the graph represents. For example, the graphs shown in Fig. 1(c) and (d) are equivalent since the first graph represents the equation $x_2 = kx_1$ and the second represents the equivalent equation $x_1 = x_2/k$.

In the more complicated configuration shown in Fig. 1(b), suppose we wish to reverse the direction of the branch which connects node x_3 to x_4. This means that we wish to express x_3 as a function of x_4 rather than x_4 as a function of x_3; i.e., instead of an equation: $x_4 = k_1x_1 + k_2x_2 + k_3x_3$, we wish to represent the equivalent equation, $x_3 = \frac{1}{k_3}x_4 - \frac{k_1}{k_3}x_1 - \frac{k_2}{k_3}x_2$.

Hence we see that the result of such a reversal leads to the new, but equivalent, graph of Fig. 1(e). This process is discussed more fully in the later section on *Inversion*. The above formal rules for setting up flow-graphs give us two basic laws of combination: *Rule* (1). If several branches entering a node x_2 all

* *Siemens Edison Swan Ltd.*

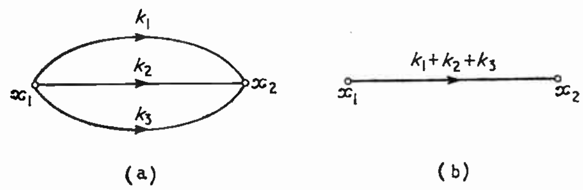


Fig. 2. *Several branches like (a) may be replaced (b) by one*

start from the same node x_1, then they may be replaced by a single branch linking x_1 to x_2 which is the algebraic sum of the original branches. Thus, the graphs of Fig. 2(a) and (b) are equivalent since

$$x_2 = k_1x_1 + k_2x_1 + k_3x_1 = (k_1 + k_2 + k_3).\, x_1$$

Rule (2). Since the pair of equations, $x_2 = k_1x_1$, $x_3 = k_2x_2$ is evidently equivalent to the single relation $x_3 = k_1k_2x_1$, the two graphs which represent, first, the two equations, and second, the resultant single equation, must be equivalent. This is illustrated in the graphs of Fig. 3(a) and (b).

Finally, we must consider the situation in which two nodes are linked by branches going in both directions. The nodes are then said to be linked by a *feedback loop*. Consider the flow-graph of Fig. 4(a). The corresponding equations will be:

$$x_2 = k_1x_1 + K_2x_3 \quad \ldots \quad (1)$$

$$x_3 = k_2x_2 \quad \ldots \quad (2)$$

$$x_4 = k_3x_3 \quad \ldots \quad (3)$$

If we eliminate x_2 from equations (1) and (2), we reduce the system to the two equations:

$$\left.\begin{aligned} x_3 &= k_1k_2x_1 + k_2K_2x_3 \\ x_4 &= k_3x_3 \end{aligned}\right\} \quad \ldots \quad (4)$$

From these we can solve directly for x_4 in terms of x_1 and get:

$$x_4 = k_1k_2k_3\left(\frac{1}{1-k_2K_2}\right)x_1 \quad \ldots \quad (5)$$

However, we can equally well represent the pair of equations (4) by introducing the concept of a *self-loop*. The flow-graph representation is shown in Fig. 4(b).

Here, x_3 is expressed as the result of a contribution from node x_1 together with a fedback portion, k_2K_2, of its own output. Since this graph must also express the relation between x_4 and x_1 shown in equation (5), we have the rule:

Rule (3). If a path goes from a node x_1 to a node x_3 by way of an intermediate node x_2, then a self-loop at node x_2, of magnitude g, will introduce a factor $1/(1-g)$ into the gain from x_1 to x_3.

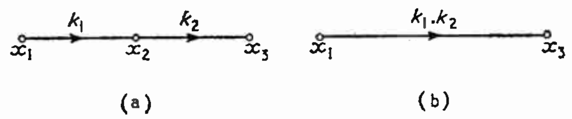


Fig. 3. *Tandem-connected branches (a) may be replaced by one as in (b)*

The quantity k_2K_2 above is called the *loop gain* of the feedback loop which connects nodes x_2 and x_3. In the graph of Fig. 4(b), thc quantity would be described as the *loop gain of the self-loop* at x_3.

Illustration: Earthed-Base Configuration

As an example of the setting up and manipulation of a flow-graph, consider the equivalent circuit of an earthed-base p–n–p transistor. The circuit equations are:

$$V_i = (r_e + r_b)\, i_e - r_b\, i_c$$

$$(r_m + r_b)\, i_e = (r_b + r_c + R_L)\ .\ i_c$$

$$V_o = i_c\, R_L.$$

The corresponding flow-graph is shown in Fig. 5(b).

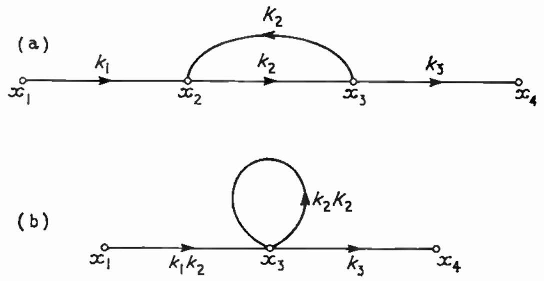


Fig. 4. *A graph with a feed back loop (a) may be transformed to one with a self-loop (b)*

In this form, the graph will give the input resistance V_i/i_e. Here, V_i appears as the sum of one branch coming direct from node i_e, and another coming by way of i_c. Combining these branches by the basic operations already defined, we have:

$$\frac{V_i}{i_e} = (r_e + r_b) + (-r_b)\cdot\frac{r_m + r_b}{r_b + r_c + R_L}$$

$$= \frac{(r_e + r_b)\,(r_b + r_c + R_L) - r_b\,(r_m + r_b)}{r_b + r_c + R_L}$$

$$= r_e + r_b\cdot\frac{(r_c - r_m + R_L)}{r_b + r_c + R_L}.$$

Suppose we now wish to obtain the voltage amplification; i.e., to express V_o in terms of V_i. This requires that V_i should appear as a source, and this can be done by inverting the path from i_e to V_i. This gives a new graph, Fig. 5(c), corresponding to a rephrased system of equations:

$$i_e = \frac{1}{r_e + r_b}\cdot V_i + \frac{r_b}{r_e + r_b}\, i_c$$

$$i_c = \frac{r_m + r_b}{r_b + r_c + R_L}\cdot i_e$$

$$V_o = R_L\ .\ i_c$$

This is the same form as the feedback graph discussed in the preceding section and reduces to the self-loop form of Fig. 5(d), in the same way as before, the feedback loop $\dfrac{r_b(r_m + r_b)}{(r_e + r_b)(r_b + r_c + R_L)}$

now appearing as a self-loop.

Hence,

$$\frac{V_o}{V_i} = \frac{(r_m + r_b) \, . \, R_L}{(r_e + r_b)(r_b + r_c + R_L)} \left[\frac{1}{1 - \dfrac{r_b(r_m + r_b)}{(r_e + r_b)(r_b + r_c + R_L)}} \right]$$

$$= \frac{(r_m + r_b) \, . \, R_L}{r_b(r_c + r_e + R_L - r_m) + r_e(r_c + R_L)}$$

This very simple case illustrates a general method of attack:

A graph is first set up from the circuit equations, and suitable branches are then inverted, where necessary, to set up the appropriate nodes as source and sink. The next step is to number the intermediate nodes in order, so that each is defined in terms of the preceding ones. In the case of the above example, the second graph, Fig. 5(c), shows node i_e defined in terms of the source V_i; i_c is then expressed in terms of i_e and finally, V_o in terms of i_e. In the general case, this means that we would consider a set of equations in the form:

$$x_1 = a_{01}x_0 + a_{11}x_1 + a_{21}x_2 + \ldots + a_{n1}x_n,$$
$$x_2 = a_{02}x_0 + a_{12}x_1 + a_{22}x_2 + \ldots + a_{n2}x_n,$$
$$\cdots$$
$$x_n = a_{0n}x_0 + a_{1n}x_1 + a_{2n}x_2 + \ldots + a_{nn}x_n.$$

Here, the source is represented by x_0 and the sink by x_n; the intermediate nodes are $x_1, x_2, \ldots, x_{n-1}$. The coefficient a_{ij} represents the branch going from x_i to x_j; similarly a_{ji} represents the return branch from x_j to x_i. With the nodes numbered in this order, we may speak of a *forward* branch a_{ij}, with $i < j$, and a *backward* branch, with $j < i$. Lastly, the coefficient a_{ii} evidently represents the self-loop at node x_i. Note that in the general case, the source x_0 feeds every one of the nodes $x_1, \ldots, x_n$. In practice, it is usually only one node, x_1, which receives a contribution from x_0—as in the case of the example worked above.

In solving such a system of equations we can proceed in two ways. First we can eliminate the intermediate variables $x_1, x_2, \ldots, x_{n-1}$, one by one, until we derive a final equation expressing x_n in terms of x_0. Alternatively, we could solve for x_n in terms of x_0 by using determinants and Cramer's rule. Similarly, we can either reduce the flow-graph stage by stage, by successively eliminating the intermediate nodes (as was done in passing from the feedback graph of the example to the self-loop graph by eliminating node i_e), or use a general formula derived by Mason[2]. This formula allows the gain from source to sink to be read directly from a graph without any preliminary manipulations. In practice, it is easier to use a combination of both methods; i.e., first reduce the graph to a simpler form by node elimination and then apply Mason's formula. In the next section, graph reduction is discussed in detail.

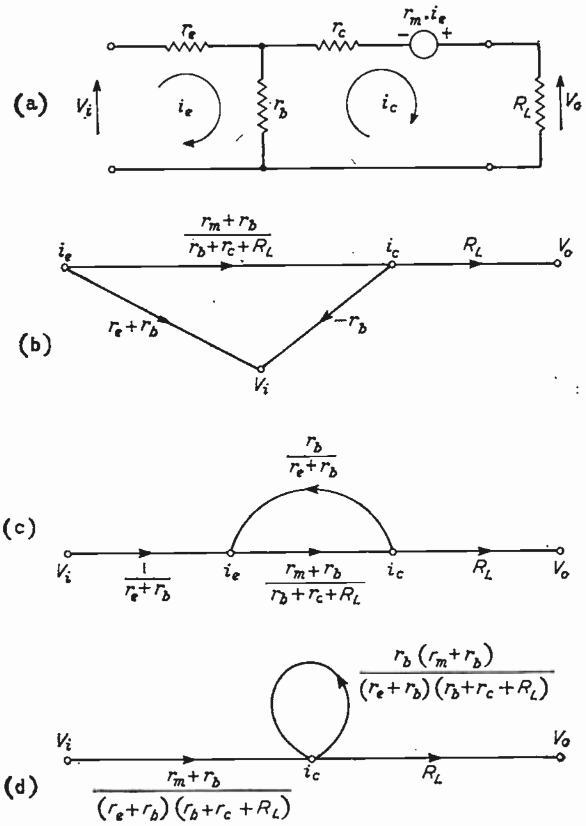


Fig. 5. *Equivalent circuit of an earthed-base transistor (a) and its graph (b). Transformations of the graph are shown in (c) and (d)*

Reduction of Flow-Graphs

It has already been shown how a simple feedback loop can be replaced by a self-loop. In effect, what has been achieved is the elimination of a backward path from the graph. We consider more complicated examples of such a process. The equations shown below are represented by the flow-graph of Fig. 6(a).

$$x_1 = a_{01}x_0 + a_{21}x_2 + a_{31}x_3$$
$$x_2 = a_{12}x_1 + a_{32}x_3$$
$$x_3 = a_{23}x_2$$

Eliminating x_1 from these equations, gives the new set of equations:

$$x_2 = a_{01} \, . \, a_{12}x_0 + a_{12} \, . \, a_{21}x_2 + (a_{32} + a_{31} \, . \, a_{12})x_3$$
$$x_3 = a_{23} \, . \, x_2$$

and this gives the new graph, Fig. 6(b), in which node x_1 has been removed.

This process of *node elimination* is one which may be applied to the simplification of flow-graphs provided the appropriate rules are followed. Provided there is no self-loop at node x_k, the elimination of x_k from the graph involves the following operations:

Rule (4). Any two nodes x_j, x_l which are linked via x_k (i.e., such that there is a path, a_{jk}, from x_j to x_k, and a path, a_{kl}, from x_k to x_l), now appear as linked by a single path, of gain $a_{jk} \, . \, a_{kl}$, going from x_j to x_l.

Rule (5). If x_k forms part of a feedback loop, then its elimination results in the appearance of one or more self-loops corresponding to the original feedback loop.

Thus, if a_{jk} links node x_j to x_k, while a_{kj} links x_k to x_j, then, when x_k is eliminated, this feedback loop becomes a self-loop at x_j, of loop gain $a_{jk} \cdot a_{kj}$.

In the example shown above, x_0 was linked to x_2 via node x_1 in Fig. 6(a). By rule (4), the elimination of x_1 results in a path of gain $a_{01} \cdot a_{12}$ going from x_0 direct to x_2. Again, x_3 was linked to x_2 via x_1, and so Fig. 6(b) contains a path of gain $a_{31} \cdot a_{12}$ going from x_3 to x_2. This combines with the path a_{32} by rule (1). Finally, x_1 and x_2 are linked in Fig. (6a) by a simple feedback loop, of loop gain $a_{12} \cdot a_{21}$. This appears, by rule (5), as a self-loop at x_2.

Suppose we now wish to reduce the graph one stage further by eliminating node x_2. This requires an explicit expression for x_2; i.e., from the equation,

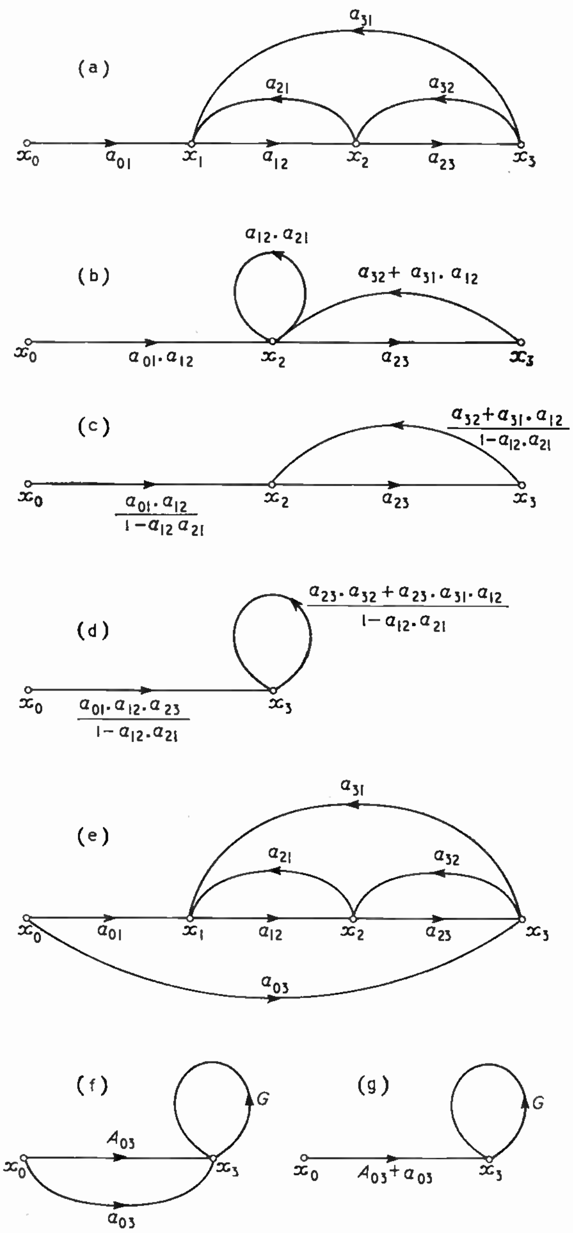


Fig. 6. *General graph (a) and successive transformations (b), (c) and (d). The addition of a new forward path to (a) leads to (e), (f) and (g)*

$$x_2 = a_{01} \cdot a_{12}\, x_0 + a_{12} \cdot a_{21}\, x_2 + (a_{32} + a_{31} \cdot a_{12})\, x_3$$

we must obtain the explicit form,

$$x_2 = \frac{a_{01} \cdot a_{12}}{1 - a_{12} \cdot a_{21}}\, x_0 + \frac{a_{32} + a_{31} \cdot a_{12}}{1 - a_{12} \cdot a_{21}}\, x_3$$

before substituting for x_2 in $x_3 = a_{23} \cdot x_2$.

In terms of the flow-graph this means we must first remove the self-loop at x_2, giving the graph of Fig. 6(c). Now we can eliminate node x_2, and obtain the reduced graph of Fig. 6(d).

If the original graph had had a self-loop of loop gain a_{11} at x_1, then the same sort of process would have been necessary before eliminating node x_1. In fact, the only change would have been in the first equation. Instead of, $x_1 = a_{01}\, x_0 + a_{21}\, x_2 + a_{31} \cdot x_3$ we should have had,

$$x_1 = \left(\frac{a_{01}}{1 - a_{11}}\right) x_0 + \left(\frac{a_{21}}{1 - a_{11}}\right) x_2 + \left(\frac{a_{31}}{1 - a_{11}}\right) x_3$$

Hence we can deduce a further general rule:

Rule (6). In eliminating a node x_k which has a self-loop, of loop gain a_{kk}, rules (4) and (5) may be used as before, provided we replace every branch a_{rk} which *enters* node x_k, by a branch $a_{rk}/(1 - a_{kk})$.

Effect of adding a Further Loop

It is worth while to compare the gains from x_0 to x_3 which would be obtained from the graphs considered above. With no self-loops in the original graph we get, from Fig. 6(d):

$$\frac{x_3}{x_0} = \frac{a_{01} \cdot a_{12} \cdot a_{23}}{1 - a_{12} \cdot a_{21}} \Bigg/ \left[1 - \frac{a_{23} \cdot a_{32} + a_{23} \cdot a_{31} \cdot a_{12}}{1 - a_{12} \cdot a_{21}}\right]$$

$$= \frac{a_{01} \cdot a_{12} \cdot a_{23}}{1 - a_{12} \cdot a_{21} - a_{23} \cdot a_{32} - a_{12} \cdot a_{23} \cdot a_{31}}$$

The numerator in this expression is the forward gain from node x_0 to node x_3; i.e., the gain we would have obtained had no return paths existed. Each of the terms in the denominator, part from unity, is the loop gain of one of the feedback loops in the graph.

Now if the graph had contained a self-loop—say at x_1—the same expression would be obtained, provided we replaced the branches entering x_1; viz:— a_{01}, a_{21} and a_{31}, by branches $\frac{a_{01}}{1-a_{11}}$, $\frac{a_{21}}{1-a_{11}}$, and $\frac{a_{31}}{1-a_{11}}$.

The gain would then become:

$$\frac{x_3}{x_0} = \frac{\frac{a_{01}}{1-a_{11}} \cdot a_{12} \cdot a_{23}}{1 - a_{12} \cdot \frac{a_{21}}{1-a_{11}} - a_{23} \cdot a_{32} - a_{12} \cdot a_{23} \cdot \frac{a_{31}}{1-a_{11}}}$$

$$= \frac{a_{01} \cdot a_{12} \cdot a_{23}}{1 - a_{11} - a_{12}.a_{21} - a_{23}.a_{32} - a_{12}.a_{23}.a_{31} + a_{11}.a_{23}.a_{32}}$$

The denominator has been changed in that the loop gain of the self-loop a_{11} now appears in common with all the other loop gains of the graph but, in addition, the product of a_{11} and the feedback loop $a_{23} \cdot a_{32}$ also appears, with a different sign prefixed. It should be noted that a_{11} and $a_{23}.a_{32}$ represent the only loops in the graph which have neither nodes nor branches in common.

Effect of adding a New Forward Path

Now suppose that the graph of Fig. 6(a) is modified by the addition of a further forward path, a_{03}, linking nodes x_0 and x_3, as in Fig. 6(e); i.e., we consider the modified set of equations:

$$\begin{aligned} x_1 &= a_{01}\,x_0 + a_{21}\,x_2 + a_{31}\,x_3 \\ x_2 &= \qquad\quad a_{12}\,x_1 + a_{32}\,x_3 \\ x_3 &= a_{03}\,x_0 + a_{23}\,x_2. \end{aligned}$$

It is clear that the flow-graph will reduce to one similar to Fig. 6(d), but with the addition of a further forward path a_{03}, as shown in Fig. 6(f). Writing,

$$G = \frac{a_{23}\,.\,a_{32} + a_{23}.\,a_{31}.\,a_{12}}{1 - a_{12}\,.\,a_{21}}\;,\; A_{03} = \frac{a_{01}\,.\,a_{12}\,.\,a_{23}}{1 - a_{12}\,.\,a_{21}}$$

we carry out the process of graph reduction as before to obtain the reduced graph, Fig. 6(f), and then using rule (1), reduce this to Fig. 6(g). Hence,

$$\frac{x_3}{x_0} = \frac{A_{03} + a_{03}}{1 - G} = \frac{a_{01}\,.\,a_{12}\,.\,a_{23} + a_{03}\,(1 - a_{12}\,.\,a_{21})}{1 - a_{12}\,.\,a_{21} - a_{23}\,.\,a_{32} - a_{23}\,.\,a_{31}\,.\,a_{12}}$$

The effect of this additional forward path has been to introduce a term into the numerator of the gain expression of the graph of Fig 6(a) viz:—$a_{03}\,(1 - a_{12}.a_{21})$; i.e., the product of the new forward path itself, and a factor $(1 - a_{12}.a_{21})$ where $a_{12}.a_{21}$ is the loop gain of the only feedback loop in the graph which does *not* touch the forward path a_{03}.

Solution by Determinants. The General Gain Formula

Considering the equations corresponding to the graph of Fig. 6(e), let us solve the system for x_3 in terms of x_0 by Cramer's rule. To do this, we re-write the equations in the form:

$$\begin{aligned} -a_{01}\,x_0 &= -x_1 + a_{21}\,x_2 + a_{31}\,x_3 \\ 0 &= a_{12}\,x_1 - x_2 + a_{32}\,x_3 \\ -a_{03}.\,x_0 &= \qquad\quad + a_{23}\,x_2 - x_3 \end{aligned}$$

$$\text{By Cramer: } \frac{x_3}{x_0} = \begin{vmatrix} -1 & a_{21} & -a_{01} \\ a_{12} & -1 & 0 \\ 0 & a_{23} & -a_{03} \end{vmatrix} \div \begin{vmatrix} -1 & a_{21} & a_{31} \\ a_{12} & -1 & a_{32} \\ 0 & a_{23} & -1 \end{vmatrix}$$

Evaluating these determinants shows that the first is numerically equal to the numerator and the second to the denominator of the gain expression obtained from the graph in Fig. 6(g). This suggests that if a method can be found of evaluating the determinants associated with a graph by combining the various branch gains and loop gains according to certain rules, it would be possible to calculate the gain of an arbitrary graph without performing any of the node-eliminations and other manipulations described above. Such a formula has, in fact, been derived by Mason and, in its most general form, is as follows:

(1). First select all possible forward paths from source to sink, and let the gains of these paths be $A_1, A_2, \ldots, A_k$ respectively. In the example we have been considering, there are two such paths, viz.: $(a_{01}\,.\,a_{12}\,.\,a_{23})$ and a_{03}.
(2). Take each path, A_i, and select all those feedback loops (including any self-loops) which do not touch the path A_i; i.e., which have neither nodes nor branches in common with A_i. In the example, there are no such loops corresponding to the path $a_{01}\,.\,a_{12}\,.\,a_{23}$, and only one loop, viz.: $a_{12}\,.\,a_{21}$, which does not touch path a_{03}.
(3). Let the loops which do not touch A_i have loop gains $T_1, T_2, T_3 \ldots$ Form the expression:

$$A_i\,(1 - T_1 - T_2 - T_3 \ldots + T_1T_2 + T_1T_3 + \ldots - T_1T_2T_3 - \ldots)$$

In the brackets, each of the loop gains $T_1, T_2, \ldots,$ appears, pre-fixed with a minus sign. The next group, T_1T_2, etc., consists of all possible products of pairs of loops, with the proviso that only those loops are to be taken which do not touch each other. These terms have a positive sign pre-fixed. Similarly we next take all possible products, $T_1T_2T_3$, of trios of mutually non-touching loops and again prefix a minus sign. This process is continued until all such sets of non-touching loops among the $T_1T_2, \ldots T_k$, are taken into account.

The numerator of the gain expression will then be the sum of all such expressions $A_i\;(1 - T_1 - T_2 \ldots)$. In the example, the numerator consists of the sum of $a_{01}\,.\,a_{12}\,.\,a_{23}$ (there being no loops which do not touch this) and the expression, $a_{03}\,(1 - a_{12}\,.\,a_{21})$.

To form the denominator of the gain expression we consider *all* the feedback loops occurring in the graph—not simply those which do not touch a particular forward path. To avoid confusion, we denote these by $G_1, G_2, \ldots,$ remembering that a certain subset of these will be the loops $T_1, T_2, \ldots, T_k$, already considered. The denominator is then the expression:

$$1 - G_1 - G_2 - \ldots + G_1G_2 + G_1G_3 + G_2G_3 + \ldots - G_1G_2G_3 - \ldots$$

where, in forming the products of loop gains, we are to observe the same proviso as before, viz.: that each product of loop gains shall only involve loops which do not touch each other. The appearance of such a product of non-touching loops was illustrated in the section on the effect of adding a new loop, when a self-loop was added to node x_1.

The statement of this gain formula is somewhat lengthy, but its application is quite simple. As a further example of its use we take the graph of Fig. 7 which involves various sets of non-touching loops and several forward paths. The forward paths from x_0 to x_4 are: $a_{01}\,.\,a_{12}\,.\,a_{24}$, $a_{01}\,.\,a_{12}\,.\,a_{23}\,.\,a_{34}$, $a_{01}\,.\,a_{13}\,.\,a_{34}$, and $a_{01}\,.\,a_{13}\,.\,a_{32}\,.\,a_{24}$. The numerator of the gain from

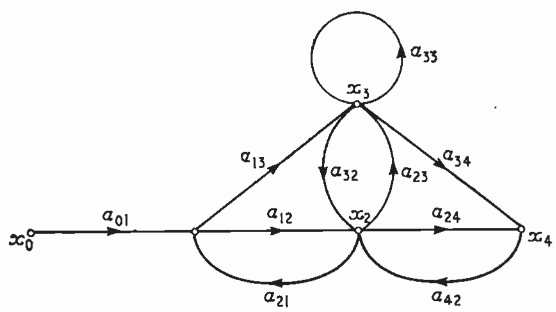


Fig. 7. *Complex graph with various non-touching loops and several forward paths*

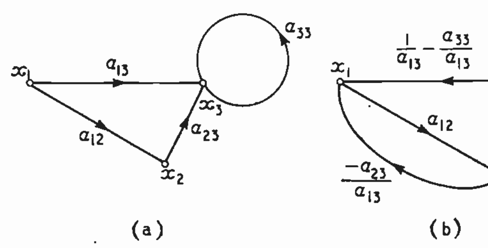


Fig. 8. *Graph (b) is the inverse of (a)*

x_0 to x_4 is then : $a_{01} . a_{12} . a_{24} (1 - a_{33}) + a_{01} . a_{12} . a_{23} . a_{34} + a_{01} . a_{13} . a_{34} + a_{01} . a_{13} . a_{32} . a_{24}$. The denominator is the expression : $1 - a_{12} . a_{21} - a_{24} . a_{42} - a_{23} . a_{32} - a_{33} + a_{33} (a_{12} . a_{21}) + a_{33} (a_{24} . a_{42}) - a_{13}\, a_{32}\, a_{21} - a_{34}\, a_{42}\, a_{23} - a_{13}\, a_{34}\, a_{42}\, a_{21}$.

It will be seen that the more complicated a graph becomes the greater will be the danger of failing to notice all the existing feedback loops and forward paths. Accordingly it is advisable, in practice, to carry out some initial reduction of the graph before applying the general formula.

Mason's derivation of the gain formula takes the form

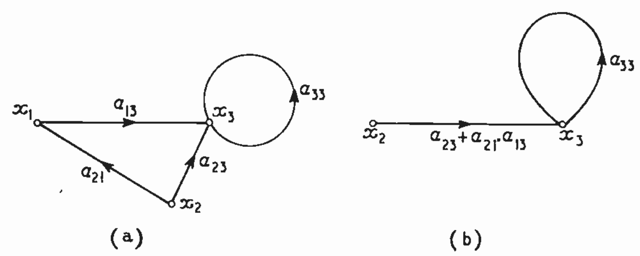


Fig. 9. *Inversion of a path leaving x_1 (a) results in the elimination of x_1 (b)*

of a topological analysis of general flow-graphs and fails to bring out the relationship with the associated determinants of the corresponding systems of equations. It is possible to prove the formula from a strictly algebraic point of view but the proof is lengthy.

Inversion

It has already been noted that the inversion of a branch in a flow-graph causes considerable alteration to the graph. Since this may result in some simplification of the gain calculation, the process is here considered in more detail. Suppose x_k is a node which appears in the original flow-graph as a pure source; i.e., a node from which branches leave but into which no branches enter. This means that x_k will only appear implicitly in the corresponding equations. In particular we shall have, for some other node x_j :

$$x_j = a_{0j}x_0 + a_{1j}x_1 + \ldots + a_{kj}x_k + \ldots + a_{nj}x_n.$$

The inversion of the branch which goes from x_k to x_j corresponds to the process of expressing x_k as a function of x_j (among other nodes)

$$x_k = \frac{1}{a_{kj}} \cdot x_j - \frac{a_{0j}}{a_{kj}} x_0 - \frac{a_{1j}}{a_{kj}} x_1 - \ldots - \frac{a_{nj}}{a_{kj}} x_n.$$

In flow-graph terms this means we carry out the following operations :

(1). The branch from x_k to x_j is reversed in direction and its gain a_{kj} is replaced by the reciprocal, $1/a_{kj}$.

(2). Every branch which enters node x_j now becomes a branch which enters node x_k, and its gain changes in sign and is divided by a_{kj}. In particular, a self-loop at x_j becomes a path going from x_j to x_k with gain $- a_{jj}/a_{kj}$. In this case, the total contribution which x_j makes to x_k is $(1 - a_{jj})/a_{kj}$. An illustration of branch inversion is shown in Fig. (8).

Failure of the Rule

Suppose we try to invert a path which leaves a node x_1 into which some other branch enters. For simplicity, consider the situation shown in the graph of Fig. 9 (a). The equations are :

$$x_1 = a_{21}\, x_2$$
$$x_3 = a_{13}\, x_1 + a_{23}\, x_2 + a_{33}\, x_3$$

Inversion of the path from x_1 to x_3 gives the equation :

$$x_1 = \left(\frac{1}{a_{13}} - \frac{a_{33}}{a_{13}} \right) x_3 - \frac{a_{23}}{a_{13}} x_2$$

Since we also have $x_1 = a_{21}x_2$, the result is the final single equation :

$$x_3 = x_2\, (a_{23} + a_{21} . a_{13}) + a_{33}\, x_3$$

and we are left with the flow-graph of Fig. 9 (b).

Thus, the only possible interpretation of inversion in such a case is the elimination of the node x_1.

There is, however, a case in which inversion can be applied when neither of the nodes are pure sources, namely when the nodes in question are connected by a feedback loop. Consider, for example, equations :

$$x_j = a_{kj}\, x_k + a_{0j}\, x_0 + a_{1j}\, x_1 + \ldots + a_{nj}\, x_n.$$
$$x_k = a_{jk}\, x_j + a_{0k}\, x_0 + a_{1k}\, x_1 + \ldots + a_{nk}\, x_n.$$

and now invert *both* paths a_{kj} and a_{jk}; i.e., invert the whole feedback loop :

$$x_j = \frac{1}{a_{jk}} x_k - \frac{a_{0k}}{a_{jk}} x_0 - \frac{a_{1k}}{a_{jk}} x_1 - \ldots - \frac{a_{nk}}{a_{jk}} x_n.$$
$$x_k = \frac{1}{a_{kj}} x_j - \frac{a_{0j}}{a_{kj}} x_0 - \frac{a_{1j}}{a_{kj}} x_1 - \ldots - \frac{a_{nj}}{a_{kj}} x_n.$$

This is a perfectly meaningful procedure and is a useful device in some cases because it can result in the elimination of other feedback loops in the graph and so

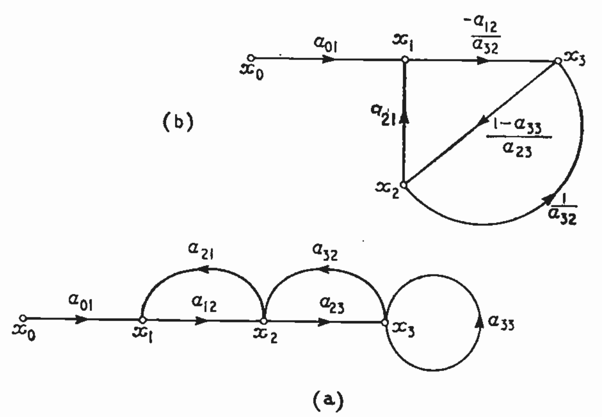


Fig. 10. *Inversion of feedback loop changes (a) to (b)*

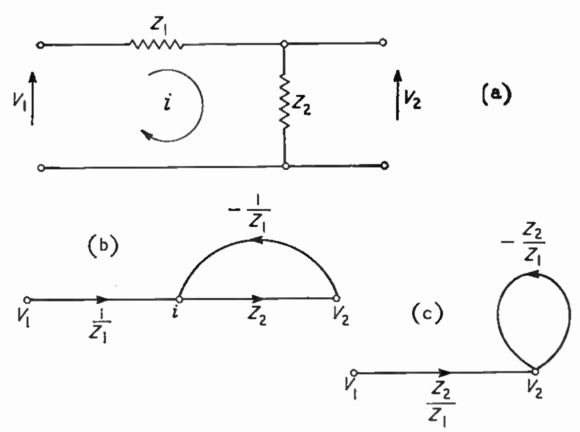


Fig. 11. *The simple circuit (a) has the graph (b) which reduces to the self-loop form (c)*

simplify gain calculations. An example of this is shown in the graphs of Fig. 10 (a) and (b), where inversion of branches a_{23} and a_{32} reduces the number of loops from three to two.

The Physical Interpretation of Flow-Graphs

It must be emphasized that a flow-graph is essentially a representation of the relations existing between currents and voltages in a network, and not of the network itself. This applies particularly to the interpretation of feedback loops. Suppose we wish to find the ratio of V_2 to V_1 in the circuit shown in Fig. 11 (a). The circuit equations can be put in the form:

$$i = \frac{1}{Z_1} V_1 - \frac{1}{Z_1} V_2 \; ; \; V_2 = Z_2 \, i$$

This gives the flow-graph, Fig. 11 (b), which reduces to the self-loop form, (c). The gain will be:

$$\frac{V_2}{V_1} = \frac{Z_2/Z_1}{1 + Z_2/Z_1} = \frac{Z_2}{Z_1 + Z_2}$$

Now this graph regards a simple potential divider as a voltage amplifier with a loop gain of $- Z_2/Z_1$. Evidently if we regard the circuit as a four-terminal black box, this does not matter in the least. However, it is obvious that care is required in interpreting a feedback loop as a representation of genuine current or voltage feedback in a physical circuit. Another example of a feedback loop occurring in a flow-graph when there is no corresponding physical feedback in the circuit, is shown when we consider any normal mesh analysis. If two meshes, say the first and second, have a common impedance, then i_2 will appear in the Kirchoff equation for the first mesh and i_1 will appear in the equation for the second mesh. This will inevitably result in a feedback loop linking nodes i_1 and i_2 in the corresponding graph.

In general, the form of the loop gains and the nature of the nodes which are connected must always be taken into account when a flow-graph is being considered as a representation of the physical behaviour of a network.

Conclusions

The use of flow-graph techniques can greatly simplify linear circuit analysis in certain problems. The algebra of such flow-graphs has been described with an accompanying interpretation in terms of more conventional methods of solving linear equations. This approach is a necessary preliminary to the treatment of Bode's feedback theory from the point of view of flow-graph analysis.

Acknowledgement

Acknowledgement is made to the Director of the Siemens Edison Swan Research Laboratory for permission to publish this paper.

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ALUMINIUM TOWER FOR RADAR AND MICROWAVE AERIALS

A 103-ft. high tower, built up from standard aluminium-alloy staging sections, has been erected at Hemel Hempstead, Herts, to demonstrate the potentialities of this system of construction for permanent or demountable towers to support aerials for radar or microwave systems.

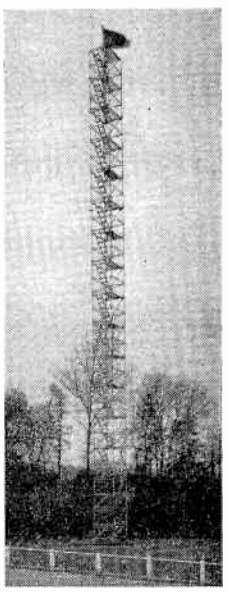
This 'Zip-Up' stairway tower is built up from a series of interchangeable folding sections which, when opened out, form rigid box frames each incorporating its own stairway. These 100-lb frames are automatically held square and secure by the diagonal stairway and by braces which are provided with patented snap-on locking hooks. Successive sections are interlocked one on top of the other by spigots which register in the tops and bottoms of the vertical corner tubes. After the 18-ft level is reached, the pre-assembled sections are hoisted up by a davit which is hooked to the top of the tower and moved up progressively.

Construction of the sections, which are manufactured by Access Equipment Ltd. of Hemel Hempstead, Herts, relies upon the close tolerance of extruded round tubing in various sizes of Noral alloy and cast aluminium-alloy couplings.

The light weight of the tower structure eliminates the necessity for elaborate reinforced-concrete foundations and enables simple footings, such as railway sleepers, to be used. The ease with which the sections can be handled makes for rapid erection even by inexperienced personnel. For example, in one recent application, the erection and guying of a 204-ft high tower was completed by a team of eight men in only four hours. The system is suitable for towers where high-speed erection is necessary for emergency communications, or where towers have to be moved from site to site.

Erection equipment is kept to a minimum and includes a guy winch, a 35-lb davit with a lifting capacity of 150 lb, a tensiometer for uniform guying, and hoisting crane for heavy aerials.

The aluminium tubing and cast couplings are supplied by the Northern Aluminium Company, Banbury.



This 103-ft high demountable tower, designed to carry aerials for radar or microwave systems, incorporates 16 standard 'Zip-Up' sections

By Computer

Vectors and Trigonometrical Tricks

We are accustomed to the idea of, say, a voltage whose magnitude at time t is $V_0 \cos \omega t$, and we regard such a voltage as having a 'frequency' $\omega/2\pi$. We are also familiar with the idea that such a voltage can be represented by the component in a particular direction of a vector whose magnitude is constant and equal to V_0, rotating with constant angular velocity ω. For engineers, the practical existence of currents of this nature is the main point, and the application of ideas borrowed from trigonometry, geometry, vector analysis, etc., is secondary.

There is, however, an underlying simplicity associated with the geometrical viewpoint and, in this article, we shall start from first geometrical principles and give precedence to geometry (mainly of an analytical kind). The significance of various trigonometrical and other formulae and techniques will thus be made clear.

Our starting point will therefore be Fig. 1, in which O is the origin of co-ordinates, Ox and Oy are perpendicular axes, and P is the point (x, y), so that OM $= x$ and PM $= y$, the angle PMO being a right angle. Let the length OP be denoted by r (always positive) and let the angle through which the axis Ox of x must be turned about O in the anticlockwise sense to lie along OP be called θ. Then we can specify the result of a displacement from O to P in two equivalent ways, namely:

(a) a displacement of magnitude r associated with a rotation through θ, and

(b) a displacement of magnitude x in the direction Ox followed by a left turn (through 90°) and a further displacement of magnitude y in the direction Oy.

The above definitions are valid whatever the magnitude of the angle θ may be. Although Fig. 1 is drawn for the case when θ exceeds a multiple of 2π by less than $\pi/2$, it applies to all cases if x_p and y_p are given appropriate signs. In general, it is very helpful to draw figures for problems of analytical geometry for the simplest case, and to assume that the laws of the algebra of negative (and, where relevant, complex) numbers will automatically extend the results obtained to other cases.

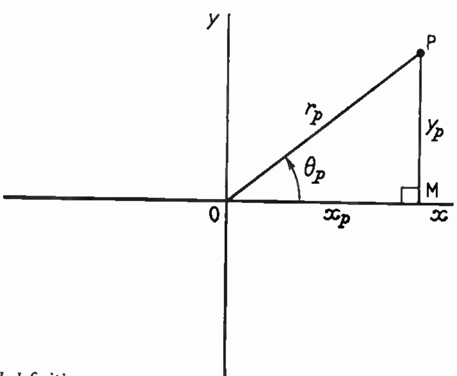
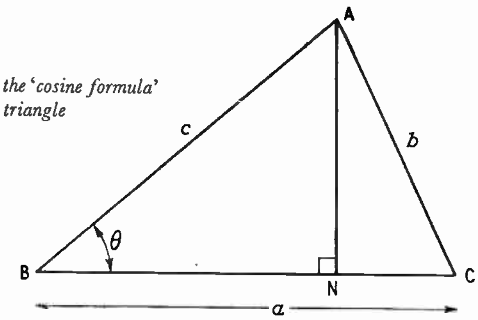


Fig. 1. *Geometrical definitions*

Fig. 2. *Deriving the 'cosine formula' for a triangle*



From the geometrical point of view, we define the symbol j as meaning 'turn left through an angle $\pi/2$ radians'. Thus a and b by themselves can be regarded as ordinary numbers but associated with lengths measured in the direction Ox. The sum $(a + b)$ and the product ab are likewise associated with lengths measured in the direction Ox, and are found by the ordinary rules of arithmetic. The product ja (or aj), however, is different; it is associated with the same length a, but measured in the direction of the y-axis instead of that of Ox. With this definition of j, we can represent the vector $\overrightarrow{\mathrm{OP}}$, or the displacement from O to P, by either the magnitude (or amplitude) r_p and the phase (or angle of rotation) θ_p, or by the symbol $(x_p + jy_p)$, representing that P can be reached from O via M. We define $\cos \theta_p$ as meaning the ratio x_p/r_p and $\sin \theta_p$ as meaning the ratio y_p/r_p in all cases.

Our next objective is to derive the formula for $\cos(\alpha - \beta)$ in a way which explains its geometrical significance, but we require one preliminary result associated with the triangle of Fig. 2, expressing b in terms of a, c and the angle θ included between a and c. It is clear from Fig. 2 as drawn that, if AN is perpendicular to BC,

$$\mathrm{AN} = c \sin \theta,\ \mathrm{BN} = c \cos \theta,\ \mathrm{CN} = a - c \cos \theta \qquad (1)$$

and these results are still correct even if $\theta > \pi/2$ (in which case N is to the left of B and therefore BN is correctly a negative quantity) or if $\angle$ACN is obtuse (in which case N is to the right of C and $c \cos \theta$ exceeds a). Since in the right-angled triangle ANC

$$\mathrm{AN}^2 + \mathrm{NC}^2 = \mathrm{AC}^2 \qquad (2)$$

it follows that

$$b^2 = c^2 \sin^2 \theta + (a - c \cos \theta)^2$$
$$= a^2 + c^2 - 2ac \cos \theta \qquad (3)$$

We now apply Equ. (3) to the triangle OAB in Fig. 3, in which $\angle$AOx is A, $\angle$BOx is B, and OA = OB = 1. The distance AB is given by drawing AX parallel to Ox and BX parallel to Oy meeting at X, so that

$$\mathrm{AX} = \cos \mathrm{B} - \cos \mathrm{A};\ \mathrm{BX} = \sin \mathrm{B} - \sin \mathrm{A}$$
$$\mathrm{AB}^2 = \mathrm{AX}^2 + \mathrm{BX}^2 = (\cos \mathrm{B} - \cos \mathrm{A})^2 + (\sin \mathrm{B} - \sin \mathrm{A})^2 \qquad (4)$$

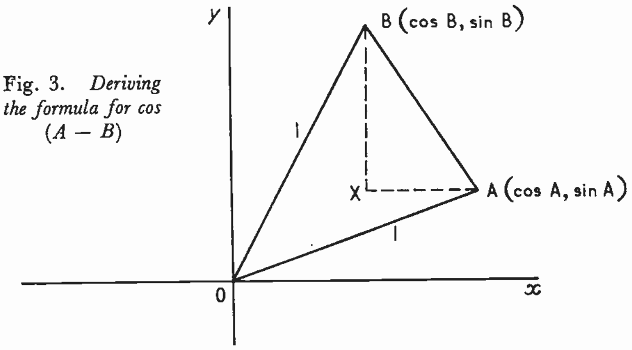


Fig. 3. *Deriving the formula for cos (A − B)*

The usual formula

$$\cos(A - B) = \cos A \cos B + \sin A \sin B \quad \ldots \quad (5)$$

follows from Equ. (3), by substituting OA = 1 for *a*, OB = 1 for *c* and AB [from Equ. (4)] for *b*; θ is (B − A), and Equ. (5) is true whatever the values of A and B may be.

Now in most textbooks on trigonometry, Equ. (5), and similar equations for sin (A ± B) are derived in a late chapter. We prefer here to derive most well-known and useful results as special cases of Equ. (5), which we have proved whatever the sizes of the angles A and B may be. We have only assumed the theorem of Pythagoras [in Equs. (2) and (4)] and the basic ideas associated with plotting graphs when the axes are rectangular, so we are in no danger of "arguing in a circle" by using Equ. (5). First, put A equal to $\pi/2$ radians (or 90°) and we find

$$\cos(\pi/2 - B) = \sin B \quad \ldots \quad (6)$$

since from Fig. 1 if $\theta_p = \pi/2$, $x_p = 0$ and $y_p = r_p$.

Again, putting A equal to zero gives

$$\cos(-B) = \cos B \quad \ldots \quad (7)$$

and in a similar way all the well-known results about the trigonometrical ratios of angles differing from a given angle by multiples of π may be derived. These results can be summarized by means of the following mnemonics

S	A	Subtract from π (180°)	Leave it alone
T	C	Subtract π (180°)	Subtract from 2π (360°)

(8)

The meaning of these mnemonics will be made clear by examples; it should be noted that the first one makes the word CAST if we start at the bottom right-hand corner and proceed anticlockwise. The mnemonics only apply to the sine, cosine and tangent; if the ratio required is, say, sec 127°, that must be written as 1/cos 127° before applying the mnemonics.

Now cos 127° is associated with an angle in the second (top left) quadrant of the mnemonics (8), where the letter S appears and the instruction 'Subtract from 180°'. The mnemonics therefore tell us that cos 127° is − cos (180° − 127°) = − cos 53°; the sign is minus because the ratio (cosine) required does not begin with the letter S falling in the quadrant. The 'A' in the first (top right) quadrant denotes that *all* trigonometrical ratios are there positive. Similarly, tan 253° = + tan (253° − 180°) = + tan 73° and sin 340° = − sin (360° − 340°) = − sin 20°. When letters are involved, we proceed as if any single letter represented a small acute angle, whether it in fact does so or not. Thus, sin (180° − A) is treated as in the second quadrant, and reduces to + sin [180° − (180° − A) = + sin A. In the case of cos (270° + B), the angle is treated as in the fourth quadrant, and reduces to + cos [360° − (270° + B)] = + cos (90° − B); this reduces to sin B as already seen in Equ. (6).

Our safeguard in any of the above is that they can always be derived by giving suitable values to A and B in Equ. (5) but the mnemonics (8) do the calculation for us instantaneously.

We have already used Fig. 2 to derive Equ. (3); a further result, obvious from this figure, is

$$c \sin B = b \sin C \quad \ldots \quad (9)$$

since both are equal to AN. In the form of Equ. (9) the result is true even if B or C is an obtuse angle; it can be divided through by sin B sin C to give

$$\frac{b}{\sin B} = \frac{c}{\sin C} = \frac{a}{\sin A} \quad \ldots \quad (10)$$

the last member of Equ. (10) being added by symmetry [or by drawing CX perpendicular to AB]. Equs. (3) and (10) are those used to 'solve triangles', that is to say, to find all of *a*, *b*, *c*, A, B, C which are not known when enough is given to determine the triangle. Equ. (3) is not always very convenient and alternatives are available, but we shall not discuss these here since solving triangles is not often necessary for electrical engineering. Any soluble triangle *can* be solved by means of these two equations.

There are a number of general results, similar to Equ. (5) and derivable from it, which permit us to manipulate trigonometrical expressions into the form most suitable for a particular problem. These results are:

$$\left.\begin{aligned} \cos(A + B) &= \cos A \cos B - \sin A \sin B \\ \sin(A + B) &= \sin A \cos B + \cos A \sin B \\ \sin(A - B) &= \sin A \cos B - \cos A \sin B \end{aligned}\right\} \quad (11)$$

The first of these is derived from Equ. (5) by putting − B instead of B, and the last of Equs. (11) is similarly derived from the second. To obtain the second, replace A by ($\pi/2$ − A) in Equ. (5).

So far, the obvious significance of Equs. (5) and (11) is that they are 'addition formulae', giving the trigonometrical ratios of compound angles A ± B in terms of those of the component angles A and B. If now we add the first of Equs. (11) to Equ. (5), we obtain

$$2 \cos A \cos B = \cos(A - B) + \cos(A + B) \quad (12)$$

and if we subtract the first of Equs. (11) from Equ. (5)

$$2 \sin A \sin B = \cos(A - B) - \cos(A + B) \quad (13)$$

By similarly manipulating the second and third of Equs (11) we derive

$$2 \sin A \cos B = \sin(A + B) + \sin(A - B) \quad (14)$$

$$2 \cos A \sin B = \sin(A + B) - \sin(A - B) \quad (15)$$

Equs. (12) to (15) enable us to break up a product of two trigonometrical ratios into a sum; this is often advantageous in numerical work, because adding is easier than multiplying. Thus for example

$$\sin 33° \sin 27° = \tfrac{1}{2}(\cos 6° - \cos 60°)$$
$$= \tfrac{1}{2}[0{\cdot}99452 - 0{\cdot}5] = 0{\cdot}24726$$

Conversely, we may sometimes find it desirable to factorize a trigonometrical expression given as the sum or difference of two terms. In this case we really need to use Equs. (12) to (15) the other way round, and many trigonometrical textbooks include separate formulae for this purpose which are equivalent to Equs. (12) to (15) with

$$A + B = x, \quad A - B = y, \quad A = \tfrac{1}{2}(x + y)$$
$$B = \tfrac{1}{2}(x - y) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

Consider for example the expression

$$X = \sin x + \sin\left(x + \frac{2\pi}{3}\right) + \sin\left(x + \frac{4\pi}{3}\right) \quad (17)$$

Applying Equ. (14) to the first and last terms, with

$$A + B = x + \frac{4\pi}{3}, \; A - B = x$$

$$\text{so that } A = x + \frac{2\pi}{3}, \; B = \frac{2\pi}{3} \quad \ldots \quad \ldots \quad (18)$$

we find

$$X = 2\sin\left(x + \frac{2\pi}{3}\right)\cos\frac{2\pi}{3} + \sin\left(x + \frac{2\pi}{3}\right)$$
$$= \sin\left(x + \frac{2\pi}{3}\right)\left\{1 + 2\cdot\cos\frac{2\pi}{3}\right\} = 0 \quad \ldots \quad (19)$$

since $\cos(2\pi/3) = -\cos(\pi/3) = -\frac{1}{2}$.

This is typical of the way in which trigonometrical expressions often contain astonishing possibilities of simplification within themselves. Very often, a process like that indicated in Equ. (18) is begun simply because the given expression, here Equ. (17), does not look as if it was in a convenient form, and any alternative is worth considering. When the process is complete, we have often not only simplified the given expression, but obtained a clue as to its significance; in Equ. (17), if the middle term had the coefficient k instead of 1, Equ. (19) would tell us that $\sin(x + 2\pi/3)$ was a factor, which is not at all obvious from the form of Equ. (17).

In Fig. 1 we noted that we could regard the position P of a point in the (x, y) plane as determined either by the 'Cartesian' co-ordinates x_p and y_p or by the 'polar' co-ordinates r_p and θ_p. The trigonometrical results we have obtained will enable us to see how these two kinds of co-ordinates fit together.

We shall find it convenient to represent the vector $\overrightarrow{OP}$ or the displacement from O to P by the symbol $x + jy$ as already mentioned. At this stage we shall assume that vectors can be added and multiplied in the way which common sense suggests, namely

$$(x_1 + jy_1) + (x_2 + jy_2) = (x_1 + x_2) + j(y_1 + y_2) \quad (20)$$

and

$$(x_1 + jy_1)(x_2 + jy_2)$$
$$= (x_1x_2 + j^2y_1y_2) + j(x_1y_2 + x_2y_1) \quad \ldots \quad (21)$$

if we should find it necessary to do such things; we will consider the significance of what we have thus achieved later. In Equ. (12) we shall replace j^2 by -1; we can think of this as the only arbitrary rule associated with the multiplication of vectors. It is not an unreasonable rule in view of the association of j with a turn to the left through $\pi/2$ radians already mentioned; two such turns in succession do amount to a reversal, which is reasonably represented by a minus sign.

If now we replace j^2 by -1 in Equ. (21) and put the two vectors to be multiplied in polar form, we find

$$(x_1 + jy_1)(x_2 + jy_2)$$
$$= r_1r_2(\cos\theta_1 + j\sin\theta_1)(\cos\theta_2 + j\sin\theta_2) \quad (22a)$$
$$= r_1r_2\{(\cos\theta_1\cos\theta_2 - \sin\theta_1\sin\theta_2)$$
$$+ j(\sin\theta_1\cos\theta_2 + \cos\theta_1\sin\theta_2)\} \quad \ldots \quad (22b)$$
$$= r_1r_2\{\cos(\theta_1 + \theta_2) + j\sin(\theta_1 + \theta_2)\} \quad (22c)$$

Equ. (22b) is derived by direct application of Equ. (21) to the right-hand side of Equ. (22a); Equ. (22c) is derived from Equ. (22b) by means of the first two of Equs. (11).

Now the form of Equ. (22c) shows us that the process of multiplication, defined arbitrarily but sensibly by means of Equ. (21), means that multiplication is in fact achieved by multiplying the magnitudes (r_1 and r_2) and adding the arguments (θ_1 and θ_2), or by inflating the vector (r_2, θ_2) in the ratio $r_2 : 1$ and subsequently rotating it through an angle θ_1 in an anticlockwise sense.

If $r_1 = r_2 = 1$, we note that multiplying the vectors involves adding their arguments. Now we first come across this idea of multiplying by means of addition when we are introduced to common logarithms, and it is this fact that makes it appropriate to represent $\cos\theta + j\sin\theta$ as $\exp.(k\theta)$ for some value of k. Further, since differentiating this expression twice reverses its sign and also multiplies its value by k^2, k^2 must be -1 like j^2.

Thus we arrive at the result that the vector $\overrightarrow{OP}$ in Fig. 1 can appropriately be represented as either $(x_p + jy_p)$ or $r_p \exp.(j\theta_p)$.

Historically, the significance of $\exp.(j\theta)$ was discovered in a slightly different but equivalent form known as De Moivre's Theorem, namely

$$(\cos\theta + j\sin\theta)^n = \cos n\theta + j\sin n\theta \quad \ldots \quad (23)$$

where n is not necessarily an integer. It is clear that Equ. (23) fits in with Equ. (22c), and we have only mentioned Equ. (23) separately because it enables us to obtain explicit formulae for $\cos n\theta$ and $\sin n\theta$ in terms of $\cos\theta$ and $\sin\theta$. Thus from Equ. (23) with $n = 2$

$$\cos 2\theta + j\sin 2\theta = (\cos\theta - j\sin\theta)^2$$
$$= (\cos^2\theta - \sin^2\theta) + 2j\sin\theta\cos\theta \quad \ldots \quad \ldots \quad (24)$$

so that

$$\cos 2\theta = \cos^2\theta - \sin^2\theta; \; \sin 2\theta = 2\sin\theta\cos\theta \quad (25)$$

Equ. (25) could also have been derived by putting $A = B = \theta$ in the first two of Equs. (11). But we could give n any value in Equ. (23) and write down an expression for $\cos n\theta$ and $\sin n\theta$ containing powers of $\cos\theta$ and $\sin\theta$ by using the Binomial Theorem (Mathematical Tools, April 1959) which would require repeated application of Equs. (11). These expressions can, if we wish, be expressed entirely in terms of $\cos\theta$; when this is done, they become what is known as Tchebycheff polynominals (or functions) in $\cos\theta$. The point we wish to emphasize here, however, is that the form of these expressions, easily obtainable from Equ. (23) but having both sines and cosines present, is often just as useful as the strict Tchebycheff form in which only $\cos\theta$ (replaced by x) is allowed.

CORRECTION

In 'Mathematical Tools', July 1959, there is an error in Equ. (5b). which should read $I_1 = -(2/\pi^2)$

and not $I_1 = -(2/\pi^2)\left[1 + \frac{1}{\pi}\right]$

Hybrid Junctions

FREQUENCY CHARACTERISTICS AS PHASE DIVIDERS AND DUPLEXERS

By R. Levy, M.A.(Cantab.)*

SUMMARY. *A review of the existing theoretical analyses of hybrid junctions is presented to show how the phase difference between the waves from the two output arms varies with frequency. It is shown that this variation is very small in the case of all well-matched 90° hybrid junctions but, for the hybrid ring, a large variation in phase difference occurs due to asymmetry of the junction. In the application of 90° hybrids to the design of balanced duplexers in the transmit condition, short-circuits are placed symmetrically in two arms (the previous 'output arms') to divert all the power into the fourth arm, and the symmetry of this arrangement gives an inherently broad-band duplexing performance. On the other hand, in the case of a hybrid ring duplexer in the transmit condition it is necessary to add an extra quarter guide-wavelength in one of the short-circuited arms. Because of asymmetry, if the extra quarter-wavelength is placed in one arm a very narrow-band v.s.w.r. characteristic is obtained but, if it is placed in the other (i.e., in the arm symmetrically placed between the input and output waveguides of the duplexer) an extremely broad-band characteristic is obtained.*

As is well known in the case of all hybrids, there are optimum positions for the short-circuits for the broadest-band match.

Typical results for an X-band hybrid ring in the transmit condition show that the bandwidth to the points where the v.s.w.r. drops to 0·8 *is* 350 *Mc/s with the extra quarter-wavelength in one arm, but is no less than* 1,650 *Mc/s with the extra quarter-wavelength in the other arm.*

90° Hybrid Junctions

One of the most useful classes of hybrid junction consists of a 3-dB directional coupler with a symmetrical coupling element. The most usual forms for this type of hybrid in waveguide are the short slot, or narrow-wall coupler[1], the top-slot or broad-wall coupler[2] and the branched-guide coupler[3]. Ideally, in all these couplers, the phase of the output waves differ by 90° and it is of interest to investigate the variation of the phase difference for a particular coupler over a frequency band. This problem has been solved by Riblet[1], but his results appear to have escaped attention as they are presented only briefly in an appendix. The following is the analysis in an expanded form.

General Theory of 90° Hybrids

The box in Fig. 1 represents a linear passive network,

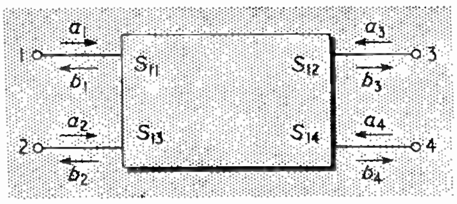


Fig. 1. *General symmetrical 90° hybrid*

the external behaviour of which is represented by the equation:

$$(b_i) = (S_{ij})\ (a_j) \qquad \ldots \quad (1)$$

where (b_i) is a column vector representing the scattered voltages at the junctions, (a_j) is a column vector representing the input voltages, and (S_{ij}) is the scattering matrix of the network.

If $i \neq j$, S_{ij} represents the transmission coefficient between arms i and j, while S_{ii} represents the reflection coefficient of arm i.

Now if the junction is a 90° hybrid it is symmetrical, and therefore:

$$\left.\begin{aligned} S_{11} &= S_{22} = S_{33} = S_{44} \\ S_{12} &= S_{21} = S_{34} = S_{43} \\ S_{13} &= S_{31} = S_{24} = S_{42} \\ S_{14} &= S_{41} = S_{23} = S_{32} \end{aligned}\right\} \qquad \ldots \quad (2)$$

so that the scattering matrix becomes:

$$\left\|\begin{matrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{11} & S_{14} & S_{13} \\ S_{13} & S_{14} & S_{11} & S_{12} \\ S_{14} & S_{13} & S_{12} & S_{11} \end{matrix}\right\| \qquad \ldots \quad (3)$$

Since the network is lossless, (S_{ij}) is unitary, i.e.,

$$(S_{ij})\ (S_{ij}{}^*) = (1) \qquad \ldots \quad (4)$$

which results in the four equations:

$$\left.\begin{aligned} |S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 + |S_{14}|^2 &= 1 \\ S_{11}S_{12}{}^* + S_{12}S_{11}{}^* + S_{13}S_{14}{}^* + S_{14}S_{13}{}^* &= 0 \\ S_{11}S_{13}{}^* + S_{12}S_{14}{}^* + S_{13}S_{11}{}^* + S_{14}S_{12}{}^* &= 0 \\ S_{11}S_{14}{}^* + S_{12}S_{13}{}^* + S_{13}S_{12}{}^* + S_{14}S_{11}{}^* &= 0 \end{aligned}\right\} \qquad \ldots \quad (5)$$

A useful relationship is derived from the last two equations by forming the product:

$$\begin{aligned} &(S_{11}S_{13}{}^* + S_{12}S_{14}{}^*)\ \ (S_{11}{}^*S_{14} + S_{12}{}^*S_{13}) \\ &= (S_{13}S_{11}{}^* + S_{14}S_{12}{}^*)\ \ (S_{12}S_{13}{}^* + S_{14}{}^*S_{11}) \qquad \ldots \quad (6) \end{aligned}$$

* *Applied Electronics Laboratories of The General Electric Company, Limited, Stanmore, England.*

which reduces to:

$$(|S_{11}|^2-|S_{12}|^2)\ (S_{13}{}^*S_{14}-S_{13}S_{14}{}^*) = (|S_{13}|^2-|S_{14}|^2)\ (S_{11}{}^*S_{12}-S_{11}S_{12}{}^*) \quad \ldots \quad (7)$$

The condition for complete isolation between arms 1 and 2 is that $S_{12} = 0$, and with this condition Equs. (5) and (7) reduce to:

$$\left.\begin{array}{l} |S_{11}|^2+|S_{13}|^2+|S_{14}|^2 = 1 \\ S_{13}S_{14}{}^*+S_{14}S_{13}{}^* = 0 \\ S_{11}S_{13}{}^*+S_{13}S_{11}{}^* = 0 \\ S_{11}S_{14}{}^*+S_{14}S_{11}{}^* = 0 \\ |S_{11}|^2(S_{13}{}^*S_{14}-S_{13}S_{14}{}^*) = 0 \end{array}\right\} \quad \ldots \quad (8)$$

These equations can be satisfied simultaneously only if $S_{11} = 0$, for the alternative $S_{11} \neq 0$ implies $S_{13} = 0$ or $S_{14} = 0$. If the former (say), then the junctions 1–4 and 2–3 are mutually isolated, a degenerate case of no interest. Hence $S_{11} = 0$, and perfect (non-degenerate) isolation implies perfect match, so that the equations (8) reduce to:

$$|S_{13}|^2+|S_{14}|^2 = 1 \quad \ldots \quad (9)$$

$$S_{13}S_{14}{}^*+S_{14}S_{13}{}^* = 0 \quad \ldots \quad (10)$$

Now in the case of a perfect hybrid, Equ. (9) gives:

$$|S_{13}| = |S_{14}| = \frac{1}{\sqrt{2}} \quad \ldots \quad (11)$$

Putting $S_{13} = \frac{1}{\sqrt{2}} \exp. (j\,\theta_3)$, $S_{14} = \frac{1}{\sqrt{2}} \exp. (j\,\theta_4)$, Equ. (10) becomes

$$\cos(\theta_3 - \theta_4) = 0 \quad \ldots \quad (12)$$

i.e., for perfect match, isolation and power division, the hybrid gives a 90° phase difference between equal waves in arms 3 and 4. In this case:

$$\begin{aligned} S_{13}{}^*S_{14}-S_{13}S_{14}{}^* &= \tfrac{1}{2}\,[\exp. -j\,(\theta_3-\theta_4) - \exp. j\,(\theta_3-\theta_4)] \\ &= j \sin(\theta_3-\theta_4) \\ &= j \quad \ldots \quad (13) \end{aligned}$$

and hence the coefficient of $(|S_{11}|^2 - |S_{12}|^2)$ in Equ. (7) cannot vanish. Putting:

$$\left.\begin{array}{l} S_{1k} = A_k \exp. (j\,\theta_k) \\ k = 1, 2, 3, 4 \end{array}\right\} \quad \ldots \quad (14)$$

Equ. (7) becomes for a slightly imperfect hybrid (i.e., $A_1, A_2 \ll 1$):

$$(A_1{}^2-A_2{}^2) = (A_3{}^2-A_4{}^2)\ A_1A_2 \sin(\theta_2-\theta_1) \quad \ldots \quad (15)$$

Hence, since $A_3 \approx A_4$, we have:

$$A_1 \approx A_2 \quad \ldots \quad (16)$$

which implies a relationship between the input v.s.w.r. and the isolation of the hybrid, as shown in Table 1.

TABLE 1

V.S.W.R.	Isolation (dB)
0·98	−40
0·95	−32
0·90	−26

The experimentally-determined performance of many hybrids shows that this relationship holds quite closely, even for 0° or 180° hybrids. Symmetry considerations (i.e., the 3-dB split) make this conclusion seem entirely reasonable.

A combination of equations (14) with the second of the set (5) results in the relation:

$$A_1A_2 \cos(\theta_1-\theta_2) + A_3A_4 \cos(\theta_3-\theta_4) = 0$$

$$\text{i.e., } |\cos(\theta_3-\theta_4)| = \frac{A_1A_2}{A_3A_4} |\cos(\theta_1-\theta_2)| \quad \ldots \quad (17)$$

From Equs. (11) and (16) the voltage isolation of a slightly imperfect hybrid is:

$$\frac{A_2}{\sqrt{2}A_4} = \frac{A_1}{\sqrt{2}A_3} \quad \ldots \quad (18)$$

and hence in the worst possible case when $\theta_1 = \theta_2$, the phase difference between the waves transmitted by arms 3 and 4 is given by:

$$\begin{aligned} |\cos(\theta_3 - \theta_4)| &= 2\ (\text{voltage isolation})^2 \\ &= 2/\text{antilog}\ (I/10) \quad \ldots \quad (19) \end{aligned}$$

where I dB is the isolation of the hybrid; e.g., for $I = 30$, $|\cos(\theta_3 - \theta_4)| = 2 \times 10^{-3}$, giving a phase difference which differs from the ideal 90° by only 0·1°. When the isolation falls to 20 dB the phase difference has increased to 1·1°. These results are plotted in Fig. 2.

Phase Properties of the Branched-Guide Hybrid

The results given above for the general case can be confirmed by an extension of the analysis given by Reed and Wheeler[3] for the particular type of 90° hybrid known as the branched-guide hybrid [described, for example, in Reference (4)]. The expressions for the input v.s.w.r., isolation, and power division between

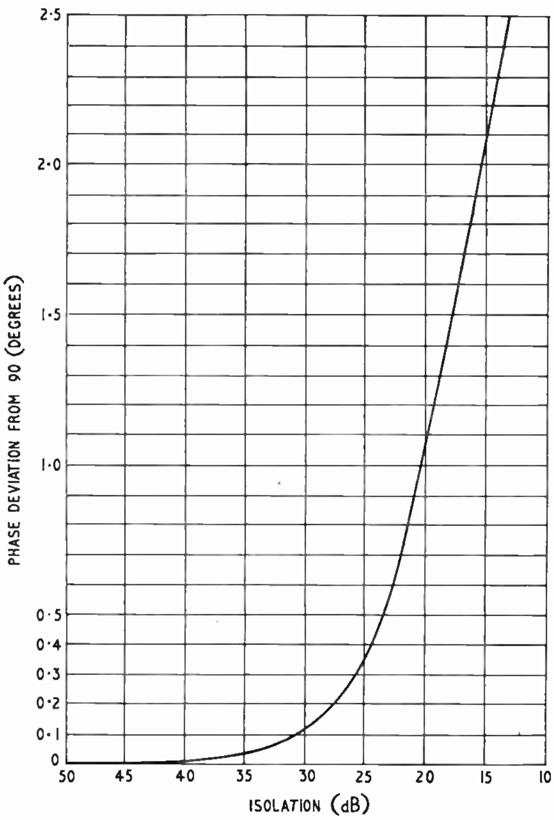


Fig. 2. *Phase variation from ideal of any 90° hybrid as a function of the isolation*

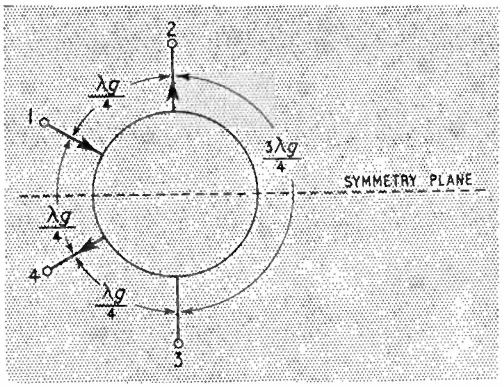


Fig. 3. *Schematic diagram of the hybrid ring*

arms are obtained as a function of a parameter t, given by the expression:

$$t = \tan\left(\frac{\pi}{4} \cdot \frac{\lambda_{go}}{\lambda_g}\right) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

where λ_g is the guide wavelength and λ_{go} is the guide wavelength at the design frequency where $t = 1$. Carrying this theory a stage further, the expression for the phase difference between the waves in the two output arms in the case of the branched-guide hybrid with four branch guides has been obtained. The formula which results is somewhat unwieldly and is not reproduced here. However, a tedious computation has shown that it is certainly within $\pm$ 0·005° of 90° for a guide wavelength deviation of $\pm$6%, where the isolation is theoretically greater than 45 dB. According to Riblet's theory this value of isolation would correspond to a phase difference $\Delta\theta$ of $\pm$ 0·0036°, which is in good agreement with the value produced by the above solution obtained from Reed and Wheeler's theory.

Properties of the Hybrid Ring

Scattering Matrix

In the case of a perfect hybrid ring, a signal input to arm 1 (Fig. 3) produces in-phase signals in arms 2 and 4, while a signal input to arm 3 divides equally in the same arms but with 180° phase difference. In the general case however, the only symmetry relations which can be written are the following:

$$\left.\begin{array}{l} S_{11} = S_{44} \\ S_{22} = S_{33} \\ S_{12} = S_{21} = S_{34} = S_{43} \\ S_{13} = S_{31} = S_{24} = S_{42} \end{array}\right| \quad \ldots \quad \ldots \quad \ldots \quad (21)$$

and the scattering matrix becomes:

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & S_{23} & S_{13} \\ S_{13} & S_{23} & S_{22} & S_{12} \\ S_{14} & S_{13} & S_{12} & S_{11} \end{bmatrix} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (22)$$

This matrix contains six parameters compared with four in the case of the 90° hybrid [Equ. (3)], so that two additional complex quantities are required to specify the behaviour of the hybrid over a frequency band. Hence, there is no reason to suppose that the phase division of the hybrid remains nearly constant over a broad band as in the case of all 90° hybrids.

Calculation of the Variation in Phase Difference of the Outputs of the Hybrid Ring

The variation of the power division with frequency for the hybrid ring has been calculated by Reed and Wheeler[3], and their method may be extended to calculate the phase division. For an input to arm 1 (Fig. 3) the complex ratio of the output waves in arms 2 and 4 at planes in each of these arms equidistant from the centre of the hybrid is given by:

$$\frac{b_2}{b_4} = \frac{\left\{\dfrac{2}{(A+D)+j(B+C)} + \dfrac{2}{(A'+D')+j(B'+C')}\right\}(1+t^2)}{\left\{\dfrac{(A-D)+j(B-C)}{(A+D)+j(B+C)} - \dfrac{(A'-D')+j(B'-C')}{(A'+D')+j(B'+C')}\right\}} \quad \ldots \quad (23)$$

where:

$$\begin{array}{ll} A = \dfrac{1 - 10t^2 + 5t^4}{(1 - 3t^2)}, & A' = \dfrac{t^4 - 10t^2 + 5}{(t^2 - 3)} \\ B = 2\sqrt{2t}, & B' = B \\ C = \sqrt{2t}(3 - t^2), & C' = \sqrt{2t}(3t - 1/t) \\ D = 1 - 3t^2, & D' = t^2 - 3 \end{array} \quad \ldots \quad (24)$$

and t is the frequency dependent parameter introduced in Equ. (20). For small deviations from the design frequency of the hybrid it is sufficiently accurate to expand the parameters listed in (24) in a Taylor series to the second term; i.e.,

$$\begin{array}{ll} A = 2 - 6\Delta t, & A' = 2 + 10\Delta t \\ B = 2\sqrt{2} + 2\sqrt{2}\Delta t, & B' = \mathrm{B} \\ C = 2\sqrt{2}, & C' = 2\sqrt{2} + 4\sqrt{2}\Delta t \\ D = -2 - 6\Delta t, & D' = -2 + 2\Delta t \end{array} \quad (25)$$

Substituting these values in Equ. (23) it may be shown that the phase of b_2/b_4 is given by:

$$\Delta\theta = \tan^{-1}(-\Delta t/\sqrt{2}) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (26)$$

Thus, for $\Delta t = \pm$ 0·05, $\Delta\theta = \mp$ 2°, and for $\Delta t = \pm$ 0·10, $\Delta\theta = \mp$ 4°.

An accurate computation gives results which agree closely with these values. Equ. (20) shows that a positive Δt, (i.e., $t > 1$) implies a decrease in λ_g; i.e, an increase in frequency from the design frequency where $t = 1$. Writing:

$$\left.\begin{array}{l} b_2 = |b_2| \exp.\,(-j\,\theta_2) \\ b_4 = |b_4| \exp.\,(-j\,\theta_4) \end{array}\right\} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (27)$$

or:

$$\frac{b_2}{b_4} = \left|\frac{b_2}{b_4}\right| \exp.\,-j\,(\theta_2 - \theta_4) = \left|\frac{b_2}{b_4}\right| \exp.\,(+j\,\Delta\theta) \quad \ldots \quad (28)$$

then a negative value of $\Delta\theta$ implies $\theta_2 > \theta_4$, or a phase increase in arm 2 for increase in frequency. Equ. (26) shows that the opposite is the case for frequency decrease, i.e., the phase of the wave in arm 2 is then less than that in arm 4 at planes in these arms equidistant from the centre of the hybrid. Table 2 shows some theoretical characteristics for a hybrid ring in waveguide 16 at a mid-band frequency of 10,000 Mc/s.

Use of Hybrids in Balanced Duplexers

Comparison of 90° and 180° Hybrids for use in Duplexers

A duplexer is a switching device which connects either a transmitter to an aerial in one position of the switch, or the transmitter to a load and the aerial to a receiver in the other. Fig. 4 shows balanced duplexers using either 90° or 180° hybrids. In the transmit condition the switches are short-circuits in the waveguide, and in the receive condition of the duplexer the switches are open-circuits.

TABLE 2

t	Wavelength Deviation $\Delta\lambda_g/\lambda_{g0}$	Frequency Deviation $\Delta f/f_0$	V.S.W.R.	Isolation (dB)	Power Division (dB)	Phase Division Degrees	Increase or Decrease of Phase in Arm 2
+0·10	− 6%	+3·4%	0·93	29·3	0·14	4	Increase
+0·05	−3%	+1·7%	—	35·2	0·025	2	Increase
0	0	0	1·0	∞	0	0	—
−0·05	+3%	−1·7%	—	35·2	0·025	2	Decrease
−0·10	+6%	−3·4%	0·93	29·3	0·14	4	Decrease

The 90° hybrid is now more frequently used in balanced duplexers than the 180° hybrid since it gives a more compact layout. In the transmit condition the short-circuits are placed at equal distances from the hybrid and lie in the same plane. Hence they may be replaced by a single shorting plate, or in the case of a duplexer for a pulsed radar system a common TR cell may be used. A photograph of a typical X-band

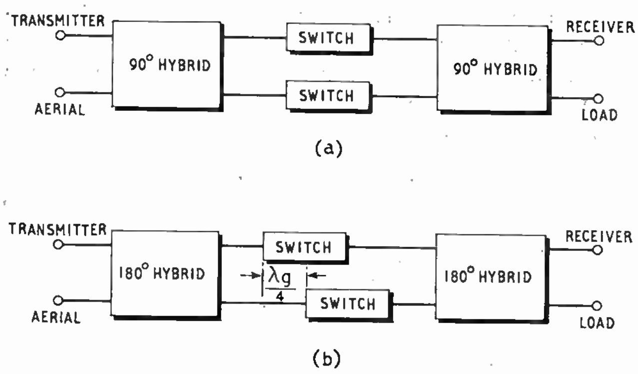


Fig. 4. *(a) Balanced duplexer using 90° hybrids; (b) balanced duplexer using 180° hybrids*

duplexer using two top-slot 90° hybrids in cascade is shown in Fig. 5. This duplexer uses a VX9204 plug-in pre TR cell.

In the case of the 180° hybrids it is necessary to stagger the position of the switches by a quarter guide-wavelength at the design frequency. Two types of 180° hybrid are in general use, the hybrid-T and the hybrid ring. In the former, the frequency sensitivity of the extra quarter wave means that in the transmit condition the duplexer is inherently narrow-band unless a compensating element is placed in one of the arms[5]. The device also suffers from the disadvantages of being difficult to manufacture and of being a 'three-dimensional' structure; i.e., with arms lying along a set of three axes mutually at right angles. The main advantage of the hybrid-T is that it is a symmetrical device so that the amplitude and phase division are both frequency independent.

On the other hand, the hybrid ring may be optimized to have a very broad-band performance as a transmit duplexer, as shown later, and all the arms lie in the same plane. However, it is difficult to design a compact duplexer using hybrid rings because the output arms containing the switches subtend an angle of 120°, so that these arms must each contain a 120° bend.

V.S.W.R. of a 90° Hybrid as a Duplexer in the Transmit Condition

The input v.s.w.r. of the transmit duplexer is determined by the combination of two waves, one reflected from the hybrid itself and one due to the departure of the wave division from ideal in both amplitude and phase. In the case of the 90° hybrid the relative phase difference between the waves in the output arms remains almost constant at 90° across a broad band and only the amplitude difference need be considered. With this assumption, C. W. Jones has shown[6] that if the ratio of the powers in the output arms of the hybrid is given by R dB, where:

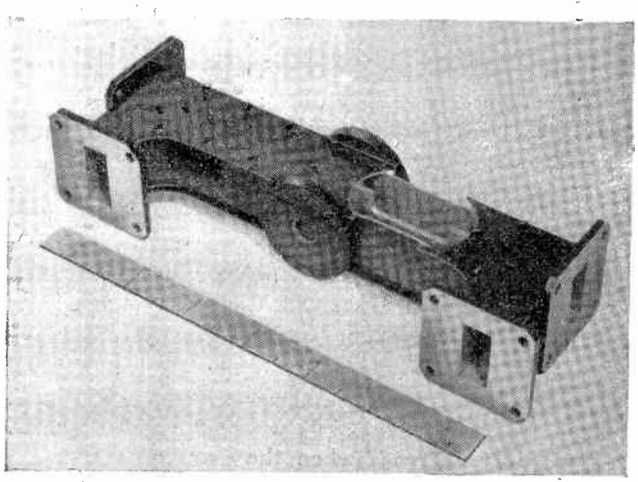
$$R = 10 \log_{10} \frac{1+\alpha}{1-\alpha} \qquad \ldots \qquad (29)$$

then assuming a hybrid which is inherently perfectly matched, the v.s.w.r. of the duplexer in the transmit condition is:

$$S_T = \frac{1-\alpha}{1+\alpha} \qquad \ldots \qquad (30)$$

Thus, for a power ratio of 1 dB (i.e., $R = 1$) $\alpha = 0{\cdot}115$, giving $S_T = 0{\cdot}79$. However, the v.s.w.r. of the hybrid alone may be of the same order, and advantage may be taken of this fact to obtain a considerable degree of

Fig. 5. *X-Band duplexer using top-slot 90° hybrids and a plug-in VX.9204 pre T.R. cell*



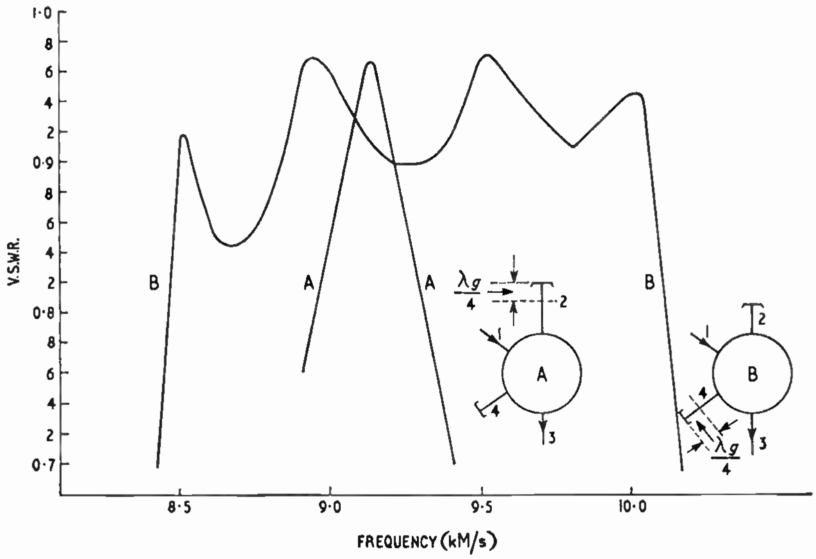


Fig. 6. *V.S.W.R. of hybrid ring in duplexer transmit condition*

cancellation between the two waves over a wide frequency band by correct phasing of the short-circuits with respect to the hybrid[7]. A v.s.w.r. of better than 0·85 over a frequency band of at least 15% may be obtained in both receive and transmit conditions of the duplexer using any pair of 90° hybrids of the type described in the references[1,2,3].

V.S.W.R. of the Hybrid Ring Duplexer in the Transmit Condition

In the transmit condition of a duplexer using hybrid rings, short-circuits are placed in arms 2 and 4 of the hybrid (shown in Fig. 3) with a quarter-wavelength difference in spacing of the position of the shorts in these arms with respect to the centre of the hybrid. This arrangement is asymmetrical in that the extra quarter-wavelength may be placed either in arm 2 or in arm 4, and the resulting arrangements are not identical. In the case of a perfect hybrid the wave incident at arm 1 may be thought to divide equally and in phase between arms 2 and 4, be reflected from the short-circuits, and re-enter the hybrid 180° out of phase, one wave having travelled one half-wavelength more than the other so that the waves recombine in arm 3. However, when the extra quarter-wavelength is placed in arm 2 the wave travelling via the short in this arm must have travelled one whole wavelength more than the wave which is reflected from the short in arm 4. Inevitably, this arrangement will be very narrow band owing to the frequency sensitivity of the extra distance travelled by the wave reflected from arm 2. For a broad-band performance it is necessary to place the extra quarter-wavelength in arm 4 so that the waves travelling via the two short-circuits will not only have the same phase to recombine in arm 3 but will also have travelled the same physical length.

The results of experiments made on an X-band hybrid ring as a transmit duplexer are shown in Fig. 6. The bandwidth to the points where the input v.s.w.r. falls to 0·8 is 350 Mc/s when the extra quarter-wavelength is placed in arm 2, but it is no less than 1,650 Mc/s when it is placed in arm 4. In the latter case, care was taken to place the short-circuits in the optimum position relative to the centre of the hybrid in order to obtain cancellation of the wave reflected from the hybrid[7].

Conclusions

In the choice of hybrids for use in microwave circuits ease of construction and compactness of layout are often prime considerations. 90° hybrids not only satisfy this condition but also retain their 90° phase-shift property over a wide band, and give an excellent performance when used in duplexers. Of 180° hybrids, the various types of hybrid-T (or "magic-T") are difficult to manufacture, result in awkward three-dimensional layouts, and are narrow band when used in the transmit duplexer condition. The hybrid ring is also difficult to use in complicated microwave circuits and, though it may give a wide-band duplexer performance when care is taken to position the short-circuits correctly, the hybrid does not retain its 0° or 180° property over a broad band.

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OBITUARY

It is with regret that we learn of the death of two pioneer engineers:

William Theodore Ditcham, A.M.I.E.E., Senior M.I.R.E., born in London, 1881, joined Marconi's Wireless Telegraph Company Ltd., in 1915.

The subsequent years to 1915 saw him engaged in the installation and demonstration of wireless equipments of the day and, with Captain H. J. Round, Mr. Ditcham developed World War I wireless direction finders. In 1919, Round and Ditcham established two-way contact with Nova Scotia and, in 1920, inaugurated the world's first radio-telephony news service. From 1925 to 1944, Mr. Ditcham was in charge of the development of Marconi transmitters and, during 1944, he was made assistant to the Engineer-in-Chief, a position he held until his retirement in 1949.

Noel Meyer Rust joined Marconi's in 1913 and subsequently worked under C. S. Franklin and, in the late 1920s, he became largely responsible for the development of the Stille system of recording on magnetized wire. In the 1930s he carried out a great deal of original research into the design of broadband feeders and aerials; among the many tangible results of this was his contribution to the aerial system of the B.B.C.'s first television station at Alexandra Palace.

During World War II, Mr. Rust worked principally for the Admiralty and did much valuable work on the design of microwave aerials.

Mr. Rust retired in 1955 after 42 years in Marconi's. He was 68 this year.

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.

Breakdown of Silicon Power Rectifiers

SIR,—When investigating repeated breakdown failures of silicon junction rectifiers used in a circuit designed to provide 700 volts 400 mA d.c. from a 2,400-c/s supply an effect was noticed which appears to be important whenever such rectifiers are used in series to rectify a high-voltage high-frequency supply.

The failures occurred despite strict adherence to the manufacturers' recommendations both to equalize the inverse voltage across each rectifier by shunting it with a resistance small compared with the reverse resistance, and to damp out transient voltages produced by transformer switching surges. Investigations using a differential amplifier fed from a capacitive potentiometer to examine the voltage across each rectifier showed that although the peak-inverse voltage across each series stack was only about half the manufacturers' maximum ratings, the voltage distribution across individual rectifiers was very unequal, having a spread of about 10 to 1, so that the inverse voltage rating of individual rectifiers was being exceeded. The explanation of this effect is as follows:

The rectifiers will, in general, exhibit carrier-storage effects; therefore, when reverse voltage is applied to a rectifier following heavy forward conduction a large reverse current flows for a time depending on the magnitude of this reverse current. The time taken to restore normal insulation for a given reverse current varies considerably from rectifier to rectifier; therefore, in any one stack one rectifier recovers its high reverse resistance first and this interrupts the flow of reverse current through the stack and, as the reverse current is now very small, the remaining rectifiers usually stay in the low-resistance state for the remainder of the cycle and hence nearly all the inverse voltage appears across the one rectifier which fails as a result.

These effects may be eliminated by shunting each rectifier with a small capacitor. In the reverse condition, when one rectifier of a stack assumes a high resistance, the current through the stack is not quite cut off as a small reverse current is diverted through the shunting capacitor, thereby bringing all rectifiers quickly into the correct high-resistance state.

To avoid damage to the rectifiers it is essential to eliminate these minority-carrier storage effects early in the reverse cycle before the inverse voltage reaches a dangerous level.

In general, this is better achieved by capacitor shunting than by resistance shunting since, with the former, the reverse current by-passing a high-resistance rectifier is proportional to the rate of change of inverse voltage and is, hence, at a maximum at the commencement of the inverse-voltage cycle and, in the latter, maximum current is not attained until the inverse voltage is also at a maximum.

To equalize the inverse voltages across individual rectifiers to within 10 per cent the required value of shunting capacitors is usually in the range 1,000 to 5,000 pF.

M. A. WESTON.

Signals Research & Development Establishment,
Christchurch, Hants.
5th August 1959.

Improving Aerial Directivity for Pulsed Signals

SIR,—The interesting article by Mr. N. F. Barber on the subject of directional arrays, in the June issue, prompts me to draw attention to another way in which directional properties may be obtained.

It is well known that a linear array of receiving elements, each of which has omnidirectional sensitivity, may be made to possess a directional response in a given plane. This depends on the existence of an ordered phase relationship between the outputs of the individual elements and various devices are used to enable this relationship to be recognized, in the presence of random background 'noise', and used to determine the angular bearing of the signal source relative to the receiving array. For example, the outputs may simply be added together either directly (plain array) or after systematic changes have been made in their relative amplitudes and phases (tapered arrays, superdirective arrays)[1,2]. Alternatively, they may be added in two or more groups and the results then multiplied together (multiplicative arrays)[3].

When the signal takes the form of a pulsed carrier (e.g., in echo-ranging systems, radio beacons, etc.), the pulse envelopes of the outputs from the receiver elements will be characterized by an ordered time distribution, the envelope delay between the outputs of any two channels depending on the angle of approach of the wave fronts to the plane of the array.

This additional means of recognizing a signal and of determining the bearing of its source normally seems to be disregarded completely, the difference in pulse arrival times appearing merely as a nuisance because it leads to distortion of the total output signal. The implication however is that existing receiving arrays, based only on the amplitude and phase characteristics of the carrier, do not make full use of the information presented to them by the arrival of pulsed waves in the medium.

A fundamental investigation of this question is being carried out at Birmingham University and it is hoped that this may lead to the development of improved directional receiving arrays.

It is interesting to note that, in contrast with the phase comparison method, the directional information is contained unambiguously in the pulse outputs of *any pair* of elements in an equi-spaced array, thus suggesting that the full length of the array need not be utilized. It will probably be found, however, that increasing the number of elements used for pulse delay recognition will improve the chances of detecting coherent signals in the presence of random noise; i.e., will improve the signal/noise ratio of the receiving system.

Electrical Engineering Department, V. G. WELSBY.
University of Birmingham.
26th August 1959.

REFERENCES

1 P. M. Woodward and J. D. Lawson, "The Theoretical Precision with which an Arbitrary Radiation Pattern may be obtained from a Source of Finite Size", *J. Instn elect. Engrs*, Part III, 1948, Vol. 95, p. 363.
2 D. G. Tucker, "Some Aspects of the Design of Strip Arrays", *Acustica*, 1956, Vol. 6, p. 403.
3 V. G. Welsby and D. G. Tucker, "Multiplicative Receiving Arrays", *J. Brit. I.R.E.*, June 1959, Vol. 19, p. 369.

Fields

SIR,—The recent Editorial on "Fields" in *Electronic & Radio Engineer*[1] raises some interesting points. It is certainly true that the field concept has come to the forefront of late, but it may be questioned whether it belongs there logically. Students may now be told, e.g., that one must not say that "charge A exerts a force on charge B" but rather that "in the vicinity of charge A is a field which exerts a force on charge B". Now if one defines the field as the force, this statement is a mere tautology; if he suggests that the field has some physical existence independent of the charge B, it must be answered that there are no experimental grounds whatsoever for such a belief. In fact, if one were interested only in forces between point charges, the "electric field" would be a completely useless concept, and likewise the "electrostatic potential". When complicated distributions of charge are involved, however, these concepts can be considered as useful summaries of results—they would seem to be nothing more—in the necessary summations or integrations.

It does not seem common to make any actual mistakes by taking the "electric field" too literally. The same cannot be said for the "magnetic field". In many text books there is still a problem in which two charges (e.g., electrons) move along parallel paths at a velocity v, the line joining them being perpendicular to the direction of motion. It is concluded, then, on the basis of the action of magnetic fields, that the force between them is $(1 - v^2/c^2)$ times what it would be if they were at rest, c being the velocity of light. Actually, all the evidence points to the conclusion that only relative motion affects the force, so that in this case, the relative velocity being zero, the force would be just the usual electrostatic force. (It is true that relativity theory, by juggling times and distances, can do something about this. But why create difficulties in the first place?) In fact, an enormous amount of nonsense has arisen from taking the "magnetic field" concept too seriously; to anyone interested in seeing how much of this may be overcome, without abandoning the field concept, Bewley's book[2] is highly recommended.

Actually there have been formulations of electrodynamics which did not employ the field concept at all. Those of the previous century, and especially the earlier part thereof, naturally suffered from a lack of complete knowledge of the problem. Ritz's work[3], and more recently that of Brown[4] and of Moon and Spencer[5], offers electrodynamics in which everything is expressed in terms of the forces between charges, as functions of their relative position, velocity, and acceleration, and the field concept is not needed. Much material pertinent to these matters may also be found in O'Rahilly's book[6].

Naturally, the works mentioned above have been concerned primarily with theoretical electrodynamics. For the needs of electrical engineering, the field concept, even though nothing but a fiction, may be a most useful fiction. It might be questioned, however, whether either engineering or science can be ultimately the better for a fiction which, in places, contradicts the facts. Certainly some serious consideration of these matters is indicated; it may be that a little of the effort which is now going into the invention of elementary particles might well be spent on the invention of an elementary electrodynamics, free from logical inconsistencies and from conflicts with experience.

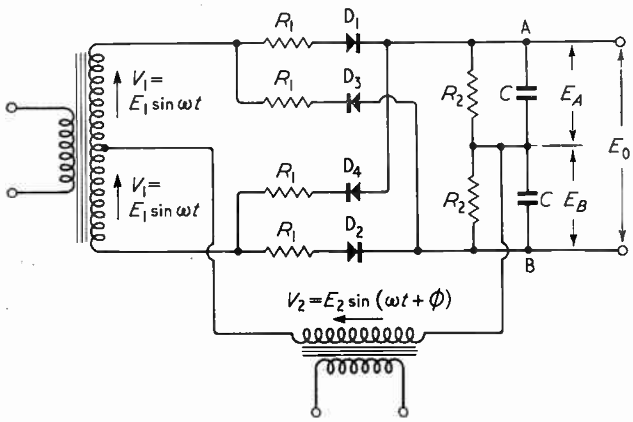
Queen's University, H. L. ARMSTRONG.
Kingston,
Ontario, Canada.
24th June 1959.

REFERENCES

1 *Electronic and Radio Engineer*, Vol. 36, p. 199, 1959.
2 L. V. Bewley, "Flux Linkages and Electromagnetic Induction", Macmillan, 1952.
3 W. Ritz, *Ann. de Chimie et de Phys*, Vol. 13, p. 145, 1908.
4 G. B. Brown, *Proc. Phys. Soc.*, B, Vol. LXVIII, p. 672, 1955.
5 P. Moon and D. E. Spencer, *J. Franklin Inst.*, Vol. 257, pp. 203, 305 and 369; 1954.
6 A. O'Rahilly, "Electromagnetics", Longmans, Green & Co., London, 1938.

Balanced Push-Pull Phase Detector

SIR,—A theory for the balanced push-pull diode phase detector in the figure was given recently by one of the authors[1]. There it was assumed that $R_2/R_1 \gg 1$. We now give a formula for the operation of the detector which does not make this assumption, though the



CR_1, CR_2 values are still considered to be large compared with the period of the applied sine waves.

The average currents flowing into node A through the diodes D_1 and D_4 are respectively[1]

$$\frac{1}{\pi R_1}(\alpha E_1 \sin\theta_1 - E_A\theta_1) \text{ and } -\frac{1}{\pi R_1}(\beta E_1 \sin\theta_4 + E_A\theta_4)$$

where $2\theta_1$ and $2\theta_4$ are the angles of conduction of D_1 (and D_3) and D_4 (and D_2) respectively,

$$0 < \theta_1 \leqslant \pi/2 \leqslant \theta_4 \leqslant \pi; \quad \alpha \equiv (1 + x^2 + 2x\cos\phi)^{\frac{1}{2}},$$
$$\beta \equiv (1 + x^2 - 2x\cos\phi)^{\frac{1}{2}}; \; x \equiv E_2/E_1$$

The algebraic sum of these currents must equal the current flowing through R_2. That is,

$$\frac{1}{\pi R_1}\left\{(\alpha E_1 \sin\theta_1 - E_A\theta_1) - (\beta E_1 \sin\theta_4 + E_A\theta_4)\right\} = \frac{E_A}{R_2} \quad .. \quad (1)$$

Now, $E_A = \alpha E_1 \cos\theta_1 = -\beta E_1 \cos\theta_4$

Substituting these values of E_A in Equation (1), we have

$$\frac{n}{\pi}(\tan\theta_1 - \theta_1) - \frac{1}{2} = \frac{1}{2} - \frac{n}{\pi}(\tan\theta_4 - \theta_4) \quad .. \quad (2)$$

where

$$n \equiv R_2/R_1 \quad .. \quad (3)$$

Equ. (2) reduces to Equ. (10) of the previous paper[1] if $n \gg 1$. Further calculations can be made on the basis of equations (11) and (12) of the previous paper, which are still valid. Again we note that the relation between θ_1 and θ_4 does not depend on the particular values of x and ϕ responsible for them. This formula can be discussed with respect to x and ϕ as before. To indicate the effect of finite values of n, we only show in Table 1 how the critical value of θ_1 (and ϕ) at which the diode D_4 just starts conducting over the entire cycle ($\theta_4 = 180°$) varies with n for the particular case, $x = 1$.

TABLE 1

n	θ_{1crit}	ϕ_{crit}
∞	77° 27′	24° 30′
100	77° 33′	24° 18′
30	77° 45′	23° 57′
10	78° 17′	22° 54′
3	79° 51′	19° 58′
1	82° 37′	14° 33′
0·3	86° 13′	7° 34′
0·1	88° 25′	3° 10′

We note that as n decreases, the critical value of ϕ decreases; that is, the flat portion of the output versus phase curve decreases. This fact seems to be important as a design criterion.

It may be remarked that a load R connected across the output terminals may be taken as loading R_2 by a shunt resistance $R/2$, thus altering the effective value of n.

The authors wish to thank Prof. R. S. Krishnan and Dr. G. Suryan for their interest and encouragement. One of the authors (S.K.) is grateful to the Council of Scientific and Industrial Research for a Research Assistantship and the other (R.C.) to the Department of Atomic Energy for a Research Fellowship.

Department of Physics, S. KRISHNAN.
Indian Institute of Science, R. CHIDAMBARAM.
Bangalore-12 (S. India).
11th June 1959.

REFERENCE

1 S. Krishnan, "Diode Phase Detectors", *Electronic and Radio Engineer*, February 1959, Vol. 36, p. 45.

Subjective Impairment of Television Pictures

SIR,—I would like to correct an error which was overlooked in my article in the May issue of *Electronic & Radio Engineer.*

It occurs towards the end of the article in the section headed "Conclusions". The third paragraph on the right-hand side of page 178 should read:

"The differences between flat and triangular noise levels for the same degree of picture impairment are given by the table of Fig. 2. For 405-line pictures the difference is 4 dB and for 625-line pictures it is 6 dB, and it is interesting to note that the C.M.T.T. revised noise-weighting network for television agrees with these values to an amount which is well within one opinion limen."

The values quoted of 6 and 8 dB refer, in fact, to the weighting network and not to the values taken from Fig. 2.

Designs Department, L. E. WEAVER.
The British Broadcasting Corporation,
London, W.1.
5th June 1959.

New Books

Fluctuation Phenomena in Semi-Conductors

By A. Van der Ziel. Pp. 168 + viii. Butterworth's Scientific Publications, 4 & 5 Bell Yard, London, W.C.2. Price 35s.

As the author states in his introduction, the subject is of interest both because it should cast light on the internal mechanisms of semi-conductors and because an understanding of it is essential to the design of semiconductor devices and circuits for minimum noise. A very brief chapter on the characterization of noisiness in two- and four-terminal networks is followed by a ten-page review of the statistical concepts and mathematical techniques which the author wishes to use; the latter includes the Wiener-Khintchine theorem, Carson's theorem relating the power spectrum of a random process to the Fourier transform of the constituent events and Langevin's equation which allows one to use a differential equation to represent the average behaviour of a stochastic system.

The remainder of the book presents mathematical analyses of various theories and models of the several types of noise which occur in semiconductors. Among these are generation-recombination noise (or shot noise), which is easily understood, and inverse-frequency noise which the author takes to be a form of modulation noise. A criticism of the book is that the reader is so rarely shown quantitative experimental evidence either to establish the relative magnitudes of the different effects or to support the theoretical models. Even in the section of Chapter 9 (Shot noise in transistors) headed "Experimental verification of the theory" we get only qualitative descriptions such as ". . . found that the experimental values of g_{s1} agreed very well with the theoretical expectations" and ". . . found good agreement between the experimental values of g_{s1} and I_{eq} at relatively low frequencies, but at high frequencies . . .". This experimental evidence may indeed be satisfactory, but the reader should resent being asked to take it on trust in the absence of quantitative definitions of "good agreement" and "relatively low".

The author assumes that the reader is already familiar with general semiconductor nomenclature and theory; and the serious worker in the semiconductor field will find this a useful reference book on the various models for fluctuation phenomena in semiconductors.

D.A.B.

Conductance Design of Active Circuits

By Keats B. Pullen Jr. Pp. 330 + xiii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 80s.

In a sense, this book is a companion to the author's "Conductance Curve Design Manual", reviewed in the September 1958 issue. In that book he gave data on a considerable number of valve types; in this, he explains how to use such data.

In essence the book deals with ways of computing the proper operating conditions for a valve or transistor, calculating stage gain and the correct component values.

The author claims that " 'Conductance Design of Active Circuits' is the first textbook published on the use of conductance curves in the design of tube and transistor circuits". It may well be that this is the first book to deal only with this subject; the bulk of the material is to be found in many textbooks of wider coverage, however. This does not mean that the fuller treatment given here is without value; the book will certainly be useful to the beginner in design.

The point of novelty, both in this book and in the previous one, appears to be the inclusion of curves of $g_p (= 1/r_a)$ and g_m as well as the usual anode-volts–anode-current curves. This is not really extra information, for it is inherently present in the usual characteristics curves. It is useful to have it separately presented and it is more accurate, as long as the g_p and g_m curves are derived by direct measurement.

W.T.C.

Radio Circuits: A Step-by-Step Survey. 4th Edition

By W. E. Miller, M.A.(Cantab.), M.Brit.I.R.E. Revised by E. A. W. Spreadbury, M.Brit.I.R.E., Associate Editor of *Wireless & Electrical Trader*. Published by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Pp. 172. Price 15s. (postage 10d.).

This book explains in simple language all the varieties of circuits that are found in radio receivers of the kind that are used for sound broadcasting. The fourth edition has, in the light of modern developments, been revised and it now covers all types of receiver, including those using transistors, car radio and f.m. receivers.

From Microphone to Ear. 2nd Edition

By G. Slot. A Philips' Technical Library publication covering, in simple language, modern sound recording and reproduction techniques. Available from Cleaver-Hume Press Ltd., 31 Wright's Lane, Kensington, London, W.8. Pp. 258. Price 21s.

Acta Polytechnica Publications:

"Kurze Ubergänge für H_{01}-Welle", by M. G. Andreasen. Electrical Engineering Series No. 3. Pp. 22. Price Sw.Kr. 7.00.

"Stetige Ubergänge für H_{01} Welle mit besonderer Berücksichtigung des konischen Ubergangs", by M. G. Andreasen. Electrical Engineering Series No. 4. Pp. 25. Price Sw.Kr. 7.00.

"The Theory of the Indirectly Heated Thermistors", by Nils Bjork. (In English) Electrical Engineering Series No. 5. Pp. 45. Price Sw.Kr. 7.00.

"Polarization-Transforming Plane Reflector for Microwaves", by J. Aagesen. (In English.) Electrical Engineering Series Vol. 8, No. 7. Pp. 27. Price Sw.Kr. 5.00.

"Table of $W = Z/(1 + Z)$ for Complex Numbers", by Karl Holberg and Jens R. Jensen. Mathematics and Computing Machinery Series MA 3. Pp. 144. Price Sw.Kr. 14.00.

"Modern Instruments and Methods for Acoustics Studies of Speech", by Gunnar Fant. (In English.) Physics including Nucleonics Series No. 1. Pp. 81. Price Sw.Kr. 7.00.

These five publications can be obtained from Acta Polytechnica Scandinavica Publishing Office, Box 5073, Stockholm 5, Sweden.

Modern Transistor Circuits

By J. M. Carroll. Pp. 268 + xii. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 66s.

". . . a comprehensive collection of modern transistor circuits, classified and arranged for easy reference. Almost 200 circuits are presented with complete design information and electronic component values".

Mullard Circuits for Audio Amplifiers

Pp. 136 + vi. Mullard Ltd., Torrington Place, London, W.C.1. Price 8s. 6d.

"This book covers twelve of the most popular Mullard circuits, including the latest designs for stereophonic amplification and, besides giving all the necessary circuit information, suggests practical chassis layouts for each amplifier. Half-tone and line illustrations are used throughout to give the constructor the greatest possible guidance".

The Cathode-Ray Tube and its Applications. 3rd Edn.

By G. Parr, M.I.E.E. and O. H. Davie, M.I.E.E. Pp. 433 + xii. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 50s.

"In this revised edition, the original theme of the use of the tube in oscillography has been pursued and brought up to date without expanding it to include the numerous applications which the ordinary user of the tube would seldom meet. The book remains as planned—a guide to the operation and use of the tube as one of the most versatile measuring instruments that has ever been devised."

Magnetic Sound Recording

By D. A. Snel. A Philips' Technical Library Publication. "This book explains the how and why of magnetic sound-recording and also focuses more attention on the variety of possible uses for recorders". Available from Cleaver-Hume Press Ltd., 31 Wright's Lane, London, W.8. Pp. 217. Price 25s.

Modern Electronic Components

By G. W. A. Dummer, M.B.E., M.I.E.E., Sen. Mem. I.R.E. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 55s.

"In this book Mr. Dummer has sought to present the first comprehensive survey of the characteristics of the more common components, together with information on their behaviour under the arduous environmental conditions to which they are now frequently subjected. Chapters on component specification, transistor-circuit components, reliability, future developments, etc., are also included. . . ."

Industrial Electronics Handbook

Edited by W. D. Cockrell. Pp. 1408 + xv. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price £8 14s. 6d.

The contents of this new publication include: Fundamentals, Control Elements, Power Supplies, Control Circuits, Circuit Applications, Instruments and Computers, Equipment, Mechanical Design, Users' Requirements, Letters Patent in the United States and Technical Information Sources.

Radio Engineering Handbook. 5th Edition

Edited by K. Henney. Pp. 1800 + v. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price £9 14s.

This revised and enlarged edition includes new chapters on microwave tubes, semiconductor diodes, transistors, nonlinear circuits, avigation electronics, direct-current and low-frequency measurements and alternating-current measurements.

The Calculation of the Median Sky Wave Field Strength in Tropical Regions (Radio Research: Special Report No. 27)

By W. R. Piggott, O.B.E., B.Sc. Pp. 38. Published for the Department of Scientific and Industrial Research by H.M. Stationery Office, Kingsway, London, W.C.2. Price 2s. 6d. (by post 2s. 10d.).

"This paper is an attempt to improve the prediction of sky wave h.f. field strengths in tropical regions".

Low-Frequency Amplifiers

Edited by A. Shure, Ph.D., Ed.D. Pp. 79 + vi. John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price $1.80.

The Atlantic Cable

By B. Dibner. Pp. 96. Burndy Library, Norwalk, Connecticut, U.S.A. Price: in cloth $3.50; in paper $2.50.

Principles of Electronics

By M. R. Gavin, M.B.E., D.Sc., F.Inst.P., M.I.E.E. and J. E. Houldin, B.Eng., Ph.D., F.Inst.P., A.M.I.E.E. Pp. 348 + xii. The English Universities Press Ltd., 102 Newgate Street, London, E.C.1. Price 30s.

"This new book, which is based on the experience of the authors in teaching and industrial research, attempts to give a general introduction to the subject of electronics suitable for a first degree or diploma course in physics or electrical engineering."

Cours Elémentaire de Mathématiques Supérieures (2nd Edn): Vol. 5.—Equations Differentielles et Applications.

By Prof. J. Quinet. Pp. 228 + xii. Dunod Editeur, 92 rue Bonaparte, Paris (6e). Price 1,100 F.

Nachrichtentechische Fachberichte:

Band 11—"Nachrichtentechnisches Schriftum", by Prof. Dr. H. Meinke and Dr. Ing A. Rihaczek. Pp. 60. Price DM 12.80.
Band 12—"Funktechnik" (Funkgeräte, Antennen, Ortung Wellenausbreitung), by Prof. Dr. H. Meinke. Pp. 116. Price DM 17.50.
Band 13—"Erzeugung von Schwingungen mit wesentlichen nichtlinearen negativen Widerständen", by R. Urtel. Pp. 38. Price DM 6.60.

These three books are published by Friedr. Vieweg & Sohn, (20b) Braunschweig, Germany, Burgplatz 1.

Fundameutals of Radio Telemetry

By Marvin Tepper. Pp. 116 + viii. Published by John F. Rider Publisher Inc., U.S.A. Distributed by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 24s.

"The author of 'Fundamentals of Telemetry' has started his book from the beginning. No assumption is made that the reader has been exposed to this field. Outside of a basic knowledge of electronics no other background is necessary to understand this text."

Semiconductor Abstracts

By Battelle Memorial Institute, U.S.A. Pp. 456. Abstracts of literature on semiconducting and luminescent materials and their applications. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 96s.

Testing of Screened Enclosures

By J. Miedzinski, B.Sc., A.M.I.E.E. Pp. 27. Technical Report M/T132. Published by the British Electrical and Allied Industries Research Association, Thornycroft Manor, Dorking Road, Leatherhead, Surrey. Price 24s. (postage 8d.).

Research for Industry 1958

Pp. 135 + iv. A report on work done by the Industrial Research Associations. Published for the Department of Scientific & Industrial Research by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 7s. 6d.

Basics of Missile Guidance and Space Techniques. Vols. 1 and 2.

By Marvin Hobbs. Pp. 144 + vi (Vol. 1), pp. 146 + iv (Vol. 2). John F. Rider Publisher Inc., 116 West 14th Street, New York 11, N.Y., U.S.A. Price Vols. 1 and 2, $3.90 each.

The Conversion of Ionospheric Virtual Height-Frequency Curves to Electron Density-Height Profiles

Pp. 48 + v. Published for the Department of Scientific & Industrial Research by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 3s. 6d.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for June* 1959

Date 1959 June	MSF 60 kc/s 1500 G.M.T. Parts in 10^{10}	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^9
1	− 180	NM
2	− 178	− 10
3	− 177	− 10
4	NM	− 11
5	NM	− 10
6	NM	− 10
7	NM	− 10
8	NM	− 10
9	− 177	− 9
10	− 178	− 9
11	− 177	− 8
12	NM	− 8
13	− 178	− 8
14	− 178	− 7
15	− 175	− 7
16	− 176	− 7
17	− 176	− 6
18	− 178	− 7
19	− 178	− 7
20	− 178	− 6
21	− 180	− 5
22	− 180	− 5
23	− 181	− 6
24	− 184	− 6
25	− 181	− 7
26	− 181	− 6
27	− 182	− 6
28	− 179	− 6
29	− 181	NM
†30	− 181	− 7

* *Nominal frequency is defined to be that frequency corresponding to a value of* 9 192 631 770 *c/s for the N.P.L. caesium resonator.*

† *At* 2400 *G.M.T. on this day the phase of the seconds pulses was advanced by* 20 *milliseconds. NM = Not Measured.*

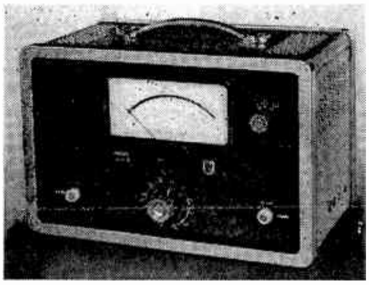
New Products

Broadband-Millivoltmeter

Research & Control Instruments Ltd., sole distributors for Philips measuring instruments in the U.K., has announced the introduction of a new instrument type GM6012.

It is a valve voltmeter which has a sensitivity 1 mV to 300 V f.s.d. (12 ranges) and a maximum bandwidth 2 c/s to 1 Mc/s. The input impedance is 4 MΩ in parallel with 20 pF on the 1-mV to 3-V ranges and 10 MΩ in parallel with 10 pF on the 10 to 300-V ranges.

For calibration, the instrument contains a



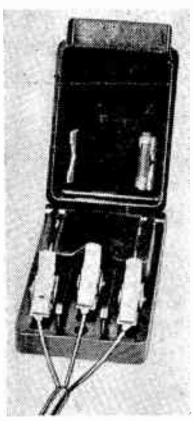
stabilized RC generator with automatically-controlled output voltages of 30 mV and 10 V.

Facilities are provided to enable the amplifier, which has an approximate gain of 50 times, to be used separately.

Research & Control Instruments Ltd.,
207 *Kings Cross Road, London, W.C.*1.

Safebloc Mains Coupler

The Rendar Safebloc is designed to provide a safe quick means of connecting



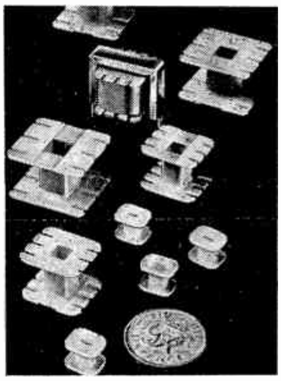
2-core and 3-core bare-ended flexible leads to the mains.

It incorporates in the base three nickel-plated clips, colour-coded red, green and black, and clearly marked L, N and E. With the lid closed, all live items are completely inaccessible; as the lid is raised, the supply is disconnected from all exposed metallic parts. The live line is fused with a 5-amp cartridge fuse as standard. Size: 5 in. long × 2⅝ in. wide and 2 in. high.

Rendar Instruments Ltd.,
Victoria Road, Burgess Hill, Sussex.

Nylon Coil Bobbins

George Goodman Ltd. has produced a range of low-priced nylon coil bobbins (for miniature transformers) in sizes which



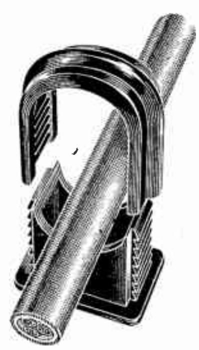
are suitable for assembling square stacks of MEA187 (RCL450), MEA262 (RCL531) and MEA218 laminations.

George Goodman Ltd.,
Robin Hood Lane, Birmingham 28, *Warwicks.*

Cable Clamps

Polystyrene cable clamps, known as 'Quickclips', are being marketed by Alma Components Ltd. They are available in several sizes to take cable diameters ranging from $\frac{3}{16}$ in. × ⅜ in. up to 1¼ in. × ⅞ in., and in three different colours (black, clear and white).

The base of the clamp (which is tapped to take a 1-B.A. screw) is fixed to the wall or chassis. The cables are then placed in position and the clip pressed home (with finger pressure). To release the cables, the



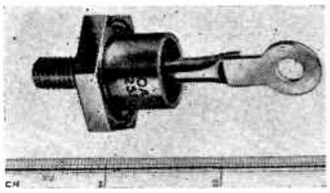
clip is removed from the base by easing it a little at a time from alternate sides.

Alma Components Ltd.,
551 *Holloway Road, London, N.*19.

Silicon Power Rectifiers

A range of three 15-A silicon power rectifiers has been introduced by Mullard.

The range comprises the OA250, OA251 and OA252, with maximum peak inverse



voltages of 50 V, 100 V and 200 V respectively, and a peak recurrent forward-current rating for all three types of 75 A at maximum p.i.v. Four OA252 rectifiers in a bridge circuit will deliver 30 A at 100 V.

Mullard Ltd.,
*Torrington Place, London, W.C.*1.

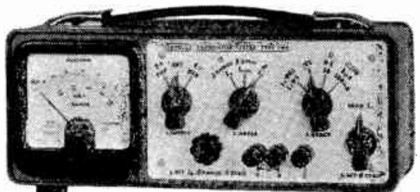
Transistor Testers TM4 and TM5

Two portable instruments designed to measure the large and small d.c. current gains of p-n-p and n-p-n transistors, the values of resistances, and the leakage currents of diodes, transistors and low-voltage electrolytic capacitors, have been announced by Levell Electronics.

The TM5 caters for tests on transistors used in radio receivers and the TM4 for tests on transistors at low power levels.

Four base-current ranges and four associated collector-current ranges are provided for tests on germanium or silicon transistors in the common-emitter connection. The base current is variable from one-fifth of full scale to full-scale value on each range and the meter can be switched to read the base current, the corresponding collector current, or the collector leakage current with the base open-circuit. Variable reverse current can be applied to the meter to balance out the leakage on the collector-current ranges so that the meter reading is proportional to the d.c. current gain factor of the transistor.

The collector-current supply is obtained



from a 3-V battery and it is arranged that the base current can be set to full-scale reading only when this battery has a potential greater than 2 V.

The TM4 measures: current gain from 100 to 500, collector current from 0·1 to 100 mA, base current from 1 to 1,000 μA, leakage current from 0·1 to 1mA and resistance from 0 to 2 MΩ.

The TM5 collector, base and leakage current ranges are exactly five times greater than the same ranges on the TM4, while the current-gain range is the same and the resistance range is from 0 to 500 kΩ.

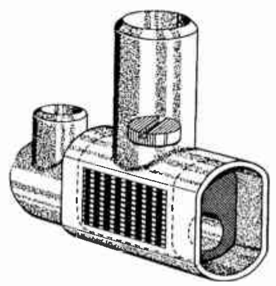
The power supply of four U11 (1½ V) torch batteries is self contained.

Levell Electronics Ltd.,
High Street, Edgware, Middx.

Nylon Connector Strips

Elkay Electrical are now marketing 1-way and 12-way nylon connectors. The 12-way strip is flexible and is moulded in such a manner that one or more of the connectors can be cut off as required. The 1-way connector is a single-entry type with an inspection hole to ascertain that the conductors are firmly gripped by the screw.

The continuous current rating of the 1-way type 1009 is 15 A, with a maximum



flash over-to-earth voltage of 4 kV and an insulation resistance of $>10^6$ MΩ. The size is ⅝ in. × $\frac{5}{16}$ in. × $\frac{11}{16}$ in. with a cable-hole diameter of $\frac{5}{32}$ in.

Elkay Electrical Manufacturing Co. Ltd.,
42 Woburn Place, London, W.C.1.

4,000-Mc/s Travelling-Wave Tubes

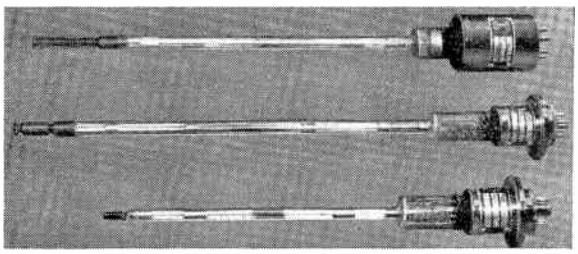
Three new tubes, designed to form a complete amplifier set for applications in three-tube microwave repeater equipments to C.C.I.T.T. standards, have been announced by the English Electric Valve Co. Ltd.

They are: N1031 a low noise tube, N1032 an intermediate amplifier tube and N1033 a power output tube. All three tubes operate in waveguide mounts over the frequency band 3,800 to 4,200 Mc/s and are pre-focused during manufacture to facilitate their installation, only minor matching adjustments being necessary to attain correct broadband operation.

Details from the makers' specification include:

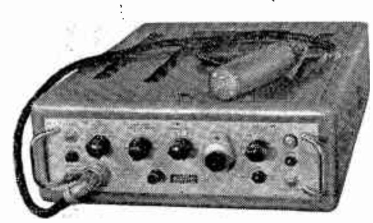
	N1031	N1032	N1033
Power output (saturated)	2 mW min.	250 mW min.	6 W min.
Gain	25 dB min.	36 dB min.	36 dB min.
Noise factor	9 dB max.	20 dB max.	30 dB max.
Cathode current ..	300 μA max.	4 mA max.	30 mA max.
Helix voltage	600 V max.	1·7 kV max.	2·5 kV max.
Collector voltage ..	1 kV max.	1·7 kV max.	1·4 kV
Magnetic field	500 gauss	550 gauss	550 gauss

English Electric Valve Co. Ltd.,
Chelmsford, Essex.



The Nanoscope

This is an auxiliary unit which, when connected with an ordinary laboratory oscilloscope, permits the observation of very



fast repetitive waveforms of the order of a few nanoseconds (10^{-9} sec) duration. It has been developed by the Atomic Energy Research Establishment primarily for use with scintillation counters.

The Nanoscope operates by sampling the fast waveform with very narrow pulses at a number of points. This results in the production of a series of pulses, stretched so as to render them visible on a normal oscilloscope of restricted bandwidth, which are proportional in amplitude to the corresponding ordinates of the sampled waveform. These ordinate pulses are brightened at their peaks, thus tracing the fast waveform as a series of dots on the oscilloscope.

In order that the sampled pulses may be centred on the screen, provision is made to delay the sampling pulse until the time-base sweeps have started and have reached the appropriate point for the display; coarse and fine delay controls are provided for this. The fine delay control is a helical potentiometer with 1,000 divisions, the whole of its range being equal in delay to twice the time-base sweep.

The Nanoscope has a sensitivity of 5 mV per trace width, an input noise of 5 mV, a recurrence frequency from 100 c/s to 10 kc/s, a rise time $< 3 \times 10^{-9}$ sec and an internal time-base ranging from 5 mμsec to 5 μsec.

An oscilloscope for use with the Nanoscope must have a d.c.-connected Y amplifier (minimum bandwidth d.c.–50 kc/s, sensitivity 1 V/cm). It must also have a terminal connected with the time-base. arranged to give a positive peak amplitude of 10 or 20 V. The sweep should be variable between 10 msec and 1 sec. The cathode of the oscilloscope tube must also be accessible to provide a 'brightening' connection.

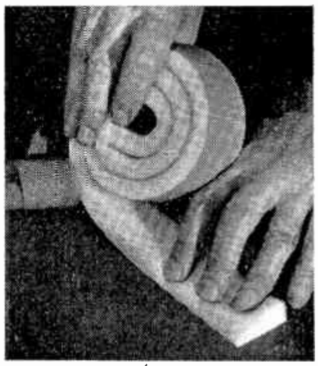
Lion Electronic Developments Ltd.,
Hanworth Trading Estate, Feltham, Middx.

Self-Adhesive Urethane Foam

Sealdraught Ltd. are now marketing self-adhesive urethane foam under the trade name Tesamoll.

The foam is bonded to a p.v.c. support which, in turn, carries the adhesive which is protected by a corrugated p.v.c. backing.

It may be used for applications such as



acoustical damping, thermal insulation, elimination of vibration and sealing.

Tesamoll can be supplied in widths from ¼ in. to 18½ in. and thicknesses from ⅛ in. to ½ in.

The basic foam, and not the adhesive backing, will withstand a temperature of 140° F at high humidities and 212° F at low humidities. The manufacturers state that the tensile strength is 9–13 lb/sq.in., the thermal conductivity is 0·2–0·25 BTU per sq. ft/hr/°F/in. and the dielectric strength is 15 kV/cm (at 50 c/s, 91° F).

Sealdraught Ltd.,
Chandos House, Buckingham Gate,
London, S.W.1.

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 journal titles conform generally with the style of the World List of Scientific Periodicals. An Author and Subject Index to the abstracts is published annually; it includes a selected list of journals abstracted, the abbreviations of their titles and their publishers' addresses. Copies of articles or journals referred to are not available from Electronic & Radio Engineer. *Application must be made to the individual publishers concerned.*

ACOUSTICS AND AUDIO FREQUENCIES

534.213 : 539.2 2444

Theory of Ultrasonic Absorption in Metals: the Collision-Drag Effect.—T. Holstein. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 479–496.)

534.232 : 534.88 2445

High-Fidelity Underwater Sound Transducers.—C. C. Sims. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 866–871.) "The problems involved in designing an underwater sound transducer for the audio-frequency range are discussed, and a new transducer for the range from about 40 c/s to about 20 kc/s is described."

534.232 : 537.228.1 : 546.431.824-31 2446

Description of the Resonances of Short Solid Barium Titanate Cylinders.—J. S. Arnold & J. G. Martner. (*J. acoust. Soc. Amer.*, Feb. 1959, Vol. 31, No. 2, pp. 217–226.) The distributions of axial and radial vibration on the plane surfaces of $BaTiO_3$ disks have been determined for a number of modes.

534.61 2447

Detecting Sound Fields.—C. B. Sacerdote. (*J. acoust. Soc. Amer.*, Feb. 1959, Vol. 31, No. 2, pp. 133–136.) A feedback system is described in which the sound pressure at any one point is automatically maintained constant. A controlling signal, derived from a microphone with uniform frequency response corrects for non-uniformity in the loudspeaker response. The method is applied to the calibration of microphones and in the study of diffraction and reflection effects.

534.61-8 : 621.395.616 2448

Condenser Microphones with Plastic Diaphragms for Airborne Ultrasonics: Part 1.—K. Matsuzawa. (*J. phys. Soc. Japan*, Dec. 1958, Vol. 13, No. 12, pp. 1533–1543.) Report of an investigation of microphones of the type described by Kuhl et al. (985 of 1955), dealing in particular with the required roughness of the back plate and the prestressing of the diaphragm.

534.614-8 : 621.3.018.75 2449

Modifications to Standard Pulse Techniques for Ultrasonic Velocity Measurements.—A. Myers, L. Mackinnon & F. E. Hoare. (*J. acoust. Soc. Amer.*, Feb. 1959, Vol. 31, No. 2, pp. 161–162.) Modifications to the circuit proposed by Cedrone & Curran (1234 of 1955) allow measurements to be made on short solid specimens having low attenuation.

534.782 2450

Electronics and the Phonetician.—H. J. F. Crabbe. (*Wireless World*, June 1959, Vol. 65, No. 6, pp. 289–294.) Description of electronic equipment for speech analysis and synthesis.

534.782 2451

The Design and Operation of the Mechanical Speech Recognizer at University College London.—P. Denes. (*J. Brit. Instn Radio Engrs*, April 1959, Vol. 19, No. 4, pp. 219–229. Discussion, pp. 230–234.) Description of a device which converts the sound waves produced by a speaker into a series of typewritten digits.

534.782 2452

Theoretical Aspects of Mechanical Speech Recognition.—D. B. Fry. (*J. Brit. Instn Radio Engrs*, April 1959, Vol. 19, No. 4, pp. 211–218. Discussion, pp. 230–234.)

534.839 2453

Noise Measurement. — E. Lübcke. (*Frequenz*, July 1958, Vol. 12, No. 7, pp. 209–213.) Various methods of objective noise measurement are discussed with particular emphasis on the need for taking account of the increase in annoyance with increasing frequency. A suitable noise meter with analyser is described.

534.851 : 389.6 2454

The International Standardization of Disk Recording.—P. H. Werner. (*Tech. Mitt. PTT*, 1st Sept. 1958, Vol. 36, No. 9, pp. 355–360. In French & German.) Summary of the recommendations adopted by the International Electrotechnical Commission for the standardization of commercial and professional disk recordings, including the standards proposed for stereophonic recordings. For magnetic-tape standards see 1010 of 1958.

621.395.61 2455

Acoustic-Front Damping in Dynamic Microphones.—W. T. Fiala. (*Audio*, March 1959, Vol. 43, No. 3, pp. 28 .. 86.) Acoustic resistance is provided by a narrow circular slot formed between the pole plate and a solid cap. With this construction the controlling resistance improves the high-frequency response and forms a sound entrance which needs no further protection.

621.395.623.7 **2456**

Acoustic Interaction in Vented Loudspeaker Enclosures.—E. de Boer. (*J. acoust. Soc. Amer.*, Feb. 1959, Vol. 31, No. 2, pp. 246–247.) An expression for sound pressure at a distance is derived from the combined sound fields of diaphragm and aperture. Near-field interaction can be represented in the conventional electrical analogue circuit by a mutual inductance. Applications are discussed. See also 3384 of 1957 (Lyon).

AERIALS AND TRANSMISSION LINES

621.315.212.001.4 **2457**

Capabilities of Coaxial Cable: Parts 1 & 2.—E. T. Pfund, Jr, W. F. Croft, B. Suverkrop & P. S. Klasky. (*Electronic Ind.*, Nov. & Dec. 1958, Vol. 17, Nos. 11 & 12, pp. 55–57 & 75–77.) The results are given of tests on six different 50-Ω solid-sheathed coaxial cables. Each cable was subjected to extremes of temperature, to vibration tests and to high-altitude corona discharge. Measurements were made of attenuation, capacitance and dielectric strength.

621.372.2 **2458**

General Considerations on the Theory of Multiple Lines.—W. Oehrl, G. Seeger & H. G. Stäblein. (*Arch. elekt. Übertragung*, June 1958, Vol. 12, No. 6, pp. 245–250.) By orthogonal transformation a set of n coupled transmission lines is converted into n equivalent lines so that the equations of the two-wire transmission line become applicable.

621.372.2 **2459**

Theory of the Helical Line of Finite Wire Thickness with reference to the Rotation of the Plane of Polarization of Waveguide Waves.—G. Piefke. (*Arch. elekt. Übertragung*, July 1958, Vol. 12, No. 7, pp. 309–316.) A general equation is obtained for calculating the propagation constants of all possible modes on a helical line with finite wire thickness and any outer medium. Propagation of linearly polarized waves inside a metal tube with an internal helical groove is investigated to show that a rotation of the plane of polarization occurs.

621.372.2 : 621.372.51.012.11 **2460**

A New Circle Diagram for Transformations in Transmission-Line Technique.—G. W. Epprecht. (*Arch. elekt. Übertragung*, June 1958, Vol. 12, No. 6, pp. 289–293.) Quadripole impedance transformation using a modified Carter diagram is described.

621.372.8 **2461**

The Application of the Asymptotic Integration of the Wave Equation to the Solution of some Waveguide and Resonator Problems.—A. Gutman. (*Dokl. Ak. Nauk S.S.S.R.*, 21st April 1959, Vol. 125, No. 6, pp. 1252–1255.) Three types of waveguides are considered: (*a*) an infinite waveguide with consecutively increasing and decreasing circular cross-section; (*b*) a semi-infinite waveguide with flared end; (*c*) a system of two helical waveguides.

621.372.8 : 621.372.413 **2462**

Resonant Properties of Nonreciprocal Ring Circuits.—F. J. Tischer. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 66–71. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 804.)

621.372.8.029.65 **2463**

Rectangular and Circular Millimetre Waveguides.—P. D. Coleman & R. C. Becker. (*Electronics*, 1st May 1959, Vol. 32, No. 18, pp. 50–51.) Tables of the dimensions and electrical characteristics of waveguides for use in the frequency range 26–350 kMc/s.

621.372.829 **2464**

Space Resonance in a Helical Waveguide Located in a Magnetic-Dielectric Medium.—V. P. Shestopalov & B. V. Kondratiev. (*Dokl. Ak. Nauk S.S.S.R.*, 1st April 1959, Vol. 125, No. 4, pp. 794–797.) A brief mathematical analysis. The retardation of waves by a helix and a dielectric is shown graphically. Examination of the curves indicates that the magnetic-dielectric medium not only increases the retardation but narrows the band within which only fast-moving waves can be propagated.

621.372.832.6 **2465**

Microwave Multiplexing Circuits.—R. E. Stone. (*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 62–65.) A theoretical and practical discussion of the hybrid ring as a multiple coupling unit.

621.372.852.1 : 621.318.134 **2466**

A Microwave Ferrite Frequency Separator.—H. Rapaport. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 53–58. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 804.)

621.372.852.15 : 621.318.134 **2467**

Ferrite-Loaded, Circularly Polarized Microwave Cavity Filters.—W. L. Whirry & C. E. Nelson. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 59–65. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 804.)

621.372.852.32 **2468**

A Medium-Power Ferrimagnetic Microwave Limiter.—E. N. Skomal & M. A. Medina. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 1000–1001.) The limiter can be adjusted to attenuate an input of 30–1 000 W peak to an output of 100 mW at frequencies near 9 360 Mc/s.

621.396.67 : 623.827 **2469**

Submarine Communication Antenna Systems.—R. W. Turner. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 735–739.) The special problems involved in the design of aerials for submarines are described.

621.396.677.3 **2470**

A Simple Graphical Method for Preparing Power Distribution Diagrams for Horizontal Dipole Arrays.—P. J. Joglekar. (*J. Inst. Telecommun. Engrs, India*, Dec. 1958, Vol. 5, No. 1, pp. 49–54.) The horizontal-directivity factors are plotted for arrays one, two and four dipoles wide.

621.396.677.3 : 523.164 **2471**

Design of 'Optimum' Arrays for Direction-Finding.—N. F. Barber. (*Electronic Radio Engr*, June 1959, Vol. 36, No. 6, pp. 222–232.) A survey of the principles underlying the design of aerials for use in radio astronomy. See 3338 of 1958.

621.396.677.7 : 621.318.134 **2472**

Nonmechanical Beam Steering by Scattering from Ferrites.—M. S. Wheeler. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 38–42. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 804.)

621.396.677.7 : 621.318.134 **2473**

An Electronic Scan using a Ferrite Aperture Luneberg Lens System.—D. B. Medved. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 101–103. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 805.)

621.396.677.833.029.64 **2474**

Directivity Pattern of 3-cm Parabolic Reflector, Feed of which is Displaced from the Focus.—S. Swarup & G. P. Srivastava. (*J. Inst. Telecommun. Engrs, India*, Dec. 1958, Vol. 5, No. 1, pp. 44–48.)

621.396.677.859 **2475**

Rigid Radome Design Considerations.—P. Davis & A. Cohen. (*Electronics*, 17th April 1959, Vol. 32, No. 16, pp. 66 . . 69.) Data are given on thin-shell and space-frame methods of construction.

AUTOMATIC COMPUTERS

681.142 **2476**

Random Number Generator using Subharmonic Oscillators.—F. Sterzer. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 241–243.) The outputs of two 2-kMc/s oscillators driven from a common 4-kMc/s source are combined in a hybrid ring. The generator can produce random binary digits at a maximum rate of 3×10^7 digits/sec.

681.142 **2477**

Pulsed Analogue Computer for Simulation of Aircraft.—A. W. Herzog. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 847–851.) A description is given of the logical design of a computer for solving the nonlinear differential equations encountered in the simulation of aircraft performance under various flight conditions.

681.142 : 061.3 **2478**

Summarized Proceedings of a Conference on Solid-State Memory and Switching Devices—London, September 1958.—(*Brit. J. appl. Phys.*, April 1959, Vol. 10, No. 4, pp. 153–158.) Some of the newer techniques are discussed with reference to ferrite, ferroelectric and superconducting devices and magnetic thin films.

681.142 : [621.314.7+621.318.042 **2479**

Transistors and Cores in Counting Circuits.—F. Rozner & P. Pengelly. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 272–274.) It is shown how square-loop magnetic cores and transistors can be combined to form noncritical reliable counting circuits capable of handling input frequencies up to 750 kc/s.

681.142 : 621.316.8 **2480**

Unifying Design Principle for the Resistance Network Analogue.—F. C. Gair. (*Brit. J. appl. Phys.*, April 1959, Vol. 10, No. 4, pp. 166–172.) A 'cell principle' technique due to Macneal (1331 of 1954) is extended and generalized. As an illustration, Poisson's equation is integrated over the volume of a representative small cell in which form it is easier to appreciate the analogy with the resistance network and to arrive at the relevant design parameters.

681.142 : 621.316.8 **2481**

Use of an Electronic Analogue Computer with Resistance Network Analogues.—J. P. Korthals Altes. (*Brit. J. appl. Phys.*, April 1959, Vol. 10, No. 4, pp. 176–180.) A method for solving partial differential equations of the elliptical type is described. The method is iterative and adjustments are made automatically by means of electronic storage elements and a switching mechanism. Some results are given.

681.142 : 621.318.134 **2482**

Magnetic Matrix Stores.—W. A. Cole. (*Wireless World*, June 1959, Vol. 65, No. 6, pp. 279–283.) Review of computer storage systems based on ferrites with rectangular hysteresis loops.

681.142 : 681.42.002.2 **2483**

Lens Designing by Electronic Digital Computer: Part 1.—C. G. Wynne. (*Proc. phys. Soc.*, 1st May 1959, Vol. 73, No. 473, pp. 777–787.) Different methods of designing optical lenses by digital computer are considered in the light of existing aberration theory.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.7 **2484**

The D.O.F.L. Microelectronics Program.—T. A. Prugh, J. R. Nall & N. J. Doctor. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 882–894.) Some of the microminiaturization techniques used at the Diamond Ordnance Fuze Laboratories are described.

621.3.049.7 **2485**

The Micro-module: a Logical Approach to Microminiaturization.—S. F. Danko, W. L. Doxey & J. P. McNaul. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 894–903.) A review is given of the progress of miniaturization; the advantages of the micro-module system, the assembly of microelement components into units with a specified electronic function, are outlined.

621.3.049.7 **2486**

Micro-module Design Progress.—P. G. Jacobs. (*Elect. Mfg*, March 1959, Vol. 63, No. 3, pp. 78–85.) Details are given of the application of a modular technique in the construction of a complete prototype communication receiver and certain digital circuits for multiplex equipment.

621.3.049.75 **2487**

Test Patterns for Printed-Circuit Materials.—T. D. Schlabach & E. E. Wright. (*Bell Lab. Rec.*, March 1959, Vol. 37, No. 3, pp. 93–95.)

621.3.049.75 : 621.396.65 **2488**

Printed Circuits Applied to Broad-Band Microwave Links.—R. Rowland. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 256–261.) Detailed review of the characteristics and advantages of typical printed circuits.

621.316.825 **2489**

Thermistors for Linear Temperature Readings.—A. B. Soble. (*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 66–67.) Two circuits are described for deriving an error voltage which varies linearly with the difference between actual and desired temperature.

621.316.86 **2490**

The Performance of Pyrolytic Carbon Resistors.—R. H. W. Burkett. (*Brit. Commun. Electronics*, April 1959, Vol. 6, No. 4, pp. 264–268.) The major factors affecting performance are carbon film thickness, moisture permeability and thickness of protective coating, operating temperature, and the quality of the ceramic substrate. The effect of these factors is small if the carbon film has a thickness of at least 100 Å.

621.316.86.019.3 **2491**

Realistic Temperature-Power Derating Requirements for Carbon Composition Resistors.—J. L. Easterday & H. Braner. (*Elect. Mfg*, March 1959, Vol. 63, No. 3, pp. 141–145, 182.) Two derating methods are described, based on statistical analyses of load/life data.

621.318.57 : 621.318.042 **2492**

Rectangular-Hysteresis-Loop Magnetic Cores as Switching Elements.—J. F. Kaposi. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 278–283.) Theoretical bases and experimental results are given for core switching with voltage and current pulses.

621.319.4.012 **2493**

Dielectric Losses with Periodic Rectangular Voltage Pulses.—B. Gross. (*Frequenz*, July 1958, Vol. 12, No. 7, pp. 230–231.) Application of a different method to the solution of a capacitor loss problem dealt with by Eisenlohr (1352 of 1957).

621.372.01 **2494**

Elements of Electronic Circuits: Part 3—Amplitude Limiting.—J. M. Peters. (*Wireless World*, June 1959, Vol. 65. No. 6, pp. 276–278.) Part 2: 2138 of July.

621.372.5 **2495**

A General Network Theorem.—G. Wunsch. (*NachrTech.*, May 1958, Vol. 8, No. 5, pp. 205–208.) The realization of the transmission factor of a four-terminal network with lumped circuit elements by a passive symmetrical quadripole and a transformer is discussed.

621.372.5 **2496**

Network Characteristics.—J. T. Allanson. (*Electronic Radio Engr*, June 1959, Vol. 36, No. 6, pp. 233–237.) The 'common network transformation' theorem of Griffiths & Mole (see 705 of 1957) is shown to lack generality.

621.372.5 : 512.831 **2497**

Matrix Analysis of Vacuum-Tube Circuits.—M. N. Srikantaswamy & K. K. Nair. (*J. Inst. Telecommun. Engrs, India*, Dec. 1958, Vol. 5, No. 1, pp. 23–32.) The equivalent four-terminal network for a feedback amplifier is derived, and the criterion for oscillation is deduced.

621.372.54 **2498**

Tchebycheff Approximation for Loss-Free Image-Parameter Filters according to Cauer.—H. Zemanek & K. Walk. (*Nachrichtentech. Z.*, June 1958, Vol. 11, No. 6, pp. 307–314.)

621.373 **2499**

Passage of a Nonlinear Oscillatory System through Resonance.—B. V. Chirikov. (*Dokl. Ak. Nauk S.S.S.R.*, 11th April 1959, Vol. 125, No. 5, pp. 1015–1018.) A mathematical analysis of a nonlinear oscillator with one degree of freedom which can be described by a Hamiltonian.

621.373.1.018.756 **2500**

Improved Low-Repetition-Rate Millimicrosecond Pulse Generator.—C. A. Burrus. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 295–296.) A discharge-line device using a vibrating-reed type of mercury switch in a suitable coaxial mounting provides pulse repetition rates of up to 500 c/s.

621.373.43 **2501**

Half-Cycle Resonant Delay Circuit.—B. Howland. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 993–994.) A new circuit generating a rectangular pulse of duration equal to the natural half cycle of an *LC* circuit. The principles are applicable to the generation of mμs pulses.

621.373.431.1 **2502**

Series Diode increases Multivibrator Sensitivity.—M. M. Vojinovic. (*Electronics*, 24th April 1959, Vol. 32, No. 17, pp. 90–91.) "Triggering sensitivity of monostable multivibrators may be greatly increased by using a semiconductor diode as a series nonlinear element in the feedback loop. Good stability can be achieved."

621.373.431.1 : 621.314.7 **2503**

Multivibrator Circuit using *p-n-p* and *n-p-n* Junction Transistors.—D. T. Jovanovic. (*Electronic Engng*, May 1959, Vol. 31, No. 375, p. 301.) Details of an improved circuit giving low output impedance and small power consumption.

621.373.431.2 : 621.314.7 **2504**

Analysis and Design of a Transistor Blocking Oscillator including Inherent Nonlinearities.—J. A. Narud & M. R. Aaron. (*Bell Syst. tech. J.*, May 1959, Vol. 38, No. 3, pp. 785–852.) The nonlinear differential equations governing circuit performance are derived. Analogue and digital computer solutions yield pulse responses in agreement with experimental results.

621.374.3 **2505**

Ten-Mc/s Pulse-Amplitude Discriminator.—J. Mey. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 282–284.) "A pulse-height discriminator has been developed which is capable of operating at repetition rates up to 10 Mc/s. It accepts positive input pulses with a threshold adjustable from 1 to 11 V. The output signal is of constant shape and amplitude. The circuit is described and test results are given."

621.374.32 **2506**

A New Type of Ring Counter.—P. J. Westoby. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 295–298.) A system is described using separate Eccles-Jordan circuits, in which any number of stages may be employed. Advantages over other systems are indicated.

621.374.32 : 621.318.57 **2507**

How Ring Counters Work.—E. Bukstein. (*Radio TV News*, March 1959, Vol. 61, No. 3, pp. 44–45 . . 160.) Basic thyratron and thermionic valve circuits are described and reference is made to the use of the circuit as an electronic rotary switch.

621.374.4 **2508**

Dividing Wide Frequency Bands.—E. L. Laine. (*Electronic Ind.*, Dec. 1958, Vol. 17, No. 12, pp. 62–63.) A regenerative circuit is described which will divide a sinusoidal input frequency by two over a wide frequency range.

621.374.43 : 517.94 **2509**

Some Properties of Mathieu and Related Functions exemplified by the Regenerative Modulation Process.—H. Jungfer. (*Frequenz*, June & July 1958, Vol. 12, Nos. 6 & 7, pp. 169–178 & 223–227.) A frequency divider using regenerative modulation can be used as an analogue method for the solution of Mathieu and Meissner differential equations. An experimental frequency divider is described and a number of waveforms obtained with it are reproduced.

621.374.5 : 534.22-8 : 531.76 **2510**

The Measurement of the Time Delay of Ultrasonic Delay Lines.—Sears. (See 2698.)

621.374.5 : 621.372.412 : 621.396.96 **2511**

Quartz Delay Lines for Radar Systems.—A. F. J. Swift. (*Brit. Commun. Electronics*, Feb. 1959, Vol. 6, No. 2, pp. 116–119.) Two lines are briefly discussed, the electromechanical type in which a silica or quartz bar is used in conjunction with one or two quartz crystal transducers, and the rectangular or polygon plate type applicable to ultrasonic frequencies. Delay times from 2 μs to 3 ms are obtainable at low power levels.

621.375.1 **2512**

Band-Pass Amplifiers, their Synthesis and Gain-Bandwidth Factor.—F. S. Atiya. (*Arch. elekt. Übertragung*, June & July 1958, Vol. 12, Nos. 6 & 7, pp. 251–264 & 317–325. In English.) Seven types of band-pass amplifier are investigated and compared, and their design formulae are given.

621.375.126 (083.57) **2513**

Using Cascading Charts.—H. Urkowitz. (*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 80–81.) Charts are given for determining the number of transitionally coupled double-tuned amplifier stages needed to provide a required gain and bandwidth. Two cases are considered: (*a*) equal damping of primary and secondary windings; (*b*) one-sided damping.

621.375.2.029.3 **2514**

Audio Amplifier Design cuts Plate Dissipation.—R. B. Dome. (*Electronics*, 10th April 1959, Vol. 32, No. 15, pp. 72–74.) An auxiliary signal at ultrasonic frequency applied to the grid reduces anode dissipation at low signal levels.

621.375.223.029.33 : 621.397.62 **2515**

Overloading Effects with Cathode Compensation.—(*Electronic Radio Engr*, June 1959, Vol. 36, No. 6, pp. 208–210.) Further details on cathode compensation of a video stage with *RC* anode load [see 1814 of June (Kitchin)] including compensation for an *RC* load in the anode of the preceding stage.

621.375.3 **2516**

Application of Magnetic Amplifiers to Engineering Problems.—A. D. Cawdery & H. T. Carden. (*Brit. Commun. Electronics*, March 1959, Vol. 6, No. 3, pp. 180–184.) The various types of magnetic-amplifier connection are discussed and a typical application of each is presented.

621.375.4 **2517**

Design of an Emitter-Current-Controlled Common-Emitter Transistor I.F. Amplifier Stage.—M. V. Joshi. (*J. Inst. Telecommun. Engrs, India*, Dec. 1958, Vol. 5, No. 1, pp. 17–22.) Input and output impedances, and their influence on the bandpass characteristics are studied. The stage is suitable for broadcast receivers.

621.375.4.018.75 **2518**

Transistorized Pulse Amplifier.—J. N. Barry & D. M. Leakey. (*Electronic Radio Engr*, June 1959, Vol. 36, No. 6, pp. 200–207.) Instead of pulse transformers, drift transistors with high cut-off frequencies (10–30 Mc/s) are used in diode logic circuits. With 10-V 0–6-μs input pulses the total rise and fall times are less than 0·05 μs, the delay between input and output pulses varying from 0·03 to 0·075 μs according to load conditions.

621.375.4.024 : 681.142 **2519**

D.C. Operational Amplifier with Transistor Chopper.—W. Hochwald & F. H. Gerhard. (*Electronics*, 24th April 1959, Vol. 32, No. 17, pp. 94–96.) Description of a drift-compensated d.c. amplifier with which an accuracy of 0·005% can be achieved in airborne analogue computers. See *Proc. nat. Electronics Conf.*, *Chicago*, 1958, Vol. 14, pp. 798–810.

621.375.9 : 538.569.4 **2520**

Weak-Field Nuclear-Magnetic-Resonance Maser.—Benoit, Grivet & Ottavi. (See 2548.)

621.375.9 : 538.569.4 **2521**

Study of a Weak-Field Maser-Type Self-Oscillator.—H. Benoit, P. Grivet & H. Ottavi. (*C. R. Acad. Sci.*, *Paris*, 12th Jan. 1959, Vol. 248, No. 2, pp. 220–223.) Characteristics of the maser described in 2548 below are discussed.

621.375.9 : 538.569.4.029.6 **2522**

Proposal for a 'Staircase' Maser.—A. E. Siegman & R. J. Morris. (*Phys. Rev. Lett.*, 1st April 1959, Vol. 2, No. 7, pp. 302–303.) A proposal for obtaining amplification at approximately twice the pump frequency, in which adiabatic fast passage is used to invert successively the 1–2 and 2–3 levels of a three-level system.

621.375.9 : 538.569.4.029.63 **2523**

Tunable L-Band Ruby Maser.—F. R. Arams & S. Okwit. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, pp. 992–993.) The product of voltage-gain and bandwidth was measured over the frequency range 850–2 000 Mc/s in a maser at temperatures of 1·5 and 4·2 °K.

621.375.9.029.55 : 621.3.011.23 **2524**

Low-Frequency Prototype Travelling-Wave Reactance Amplifier.—P. P. Lombardo & E. W. Sard. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 995–996.) The circuit of an amplifier with special junction-diode nonlinear capacitors in a semidistributed transmission line is described. Performance data at a signal frequency of 4·5 Mc/s show midband gains of 6·1–8·4 dB with overall effective input noise temperatures of about 120 °K.

621.376.23 : 621.385.2.029.6 **2525**

Vacuum-Diode Microwave Detection.—Dye, Hessler, Knight, Miesch & Papp. (See 2794.)

621.376.332 : 621.372.2 **2526**

A New Wide-Band Discriminator.—N. B. Chakraborti. (*Indian J. Phys.*, Dec.

1958, Vol. 32, No. 12, pp. 537–546.) A transmission-line arrangement for frequency discrimination is described. Output is linear for frequency deviations up to 50 %. Two practical circuits are given, with centre frequency 400 kc/s, and 6·4 Mc/s respectively.

GENERAL PHYSICS

537.222.2 **2527**

Surface Charges produced on Insulators by Short- and Long-Time Ionization.—S. I. Reynolds. (*Nature, Lond.*, 7th March 1959, Vol. 183, No. 4662, pp. 671–672.)

537.311.33 : 539.2 **2528**

Note on the Motion of Electrons and Holes in Perturbed Lattices.—R. R. Haering. (*Canad. J. Phys.*, Jan. 1959, Vol. 37, No. 1, pp. 47–52.) "A new set of orthogonal, localized functions is discussed. The functions bear the same relation to the functions of Luttinger and Kohn [2316 of 1955] as the Wannier functions bear to the Bloch functions. The new functions are used to give an alternative derivation of the effective mass equation."

537.32 **2529**

Thermoelectricity at Very Low Temperatures.—D. K. C. MacDonald. (*Science*, 10th April 1959, Vol. 129, No. 3354, pp. 943–949.) Review of theory and discussion of recent experimental work at temperatures below 20 °K.

537.52 **2530**

A Possible Mechanism of the Potential Variation of the Joshi Effect.—M. Venugopalan. (*J. phys. Soc. Japan*, Dec. 1958, Vol. 13, No. 12, pp. 1544–1546.)

537.52 : 621.374.3 **2531**

A Method of Studying the Electrical Recovery of a Gas after a Pulse Discharge.—R. J. Armstrong. (*J. Electronics Control*, Feb. 1959, Vol. 6, No. 2, pp. 162–164.) A method devised for a triggered discharge, as in a thyratron.

537.56 **2532**

On the Temperature of Electrons in a Plasma in a Variable Electric Field.—A. V. Gurevich. (*Zh. eksp. teor. Fiz.*, Aug. 1958, Vol. 35, No. 2(8), pp. 392–400.) The investigation shows that the electron gas can exist in two stable states with different temperatures; transition from one state to another occurs at certain critical values of the field and is followed by a considerable change in the electron temperature. A particular type of hysteresis is noted in the dependence of the electron temperature on the field amplitude and frequency. Expressions for the complex conductivity of the plasma are also derived.

537.56 **2533**

Measurement of the Attachment of Slow Electrons in Oxygen.—L. M. Chanin, A. V. Phelps & M. A. Biondi. (*Phys. Rev. Lett.*, 15th April 1959, Vol. 2, No. 8, pp. 344–346.) Drift-tube measurements are extended into the thermal energy range; a three-body attachment process is indicated.

537.56 : 538.56 **2534**

Theory of Excited Plasma Waves.—M. Sumi. (*J. phys. Soc. Japan*, Dec. 1958, Vol. 13, No. 12, pp. 1476–1485.) The excitation of oscillations in a uniform plasma by an injected electron beam is considered, and the frequency, phase velocity and time-rate of growth of the excited waves are derived as functions of the wave number.

537.56 : 538.56 **2535**

Nonlinear Electron Oscillations in a Cold Plasma.—J. M. Dawson. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 383–387.)

537.56 : 538.63 **2536**

Production of Two Temperatures in an Ionized Gas in a Magnetic Field.—E. Larish & I. Shekhtman. (*Zh. eksp. teor. Fiz.*, Aug. 1958, Vol. 35, No. 2(8), pp. 514–515.) If the electrons radiate a considerable part of their energy in a time short compared to the relaxation time of the gas, the electronic temperature will differ considerably from the ionic temperature.

538.3 : 537.122 **2537**

A Consistency Condition for Electron Wave Functions.—C. J. Eliezer. (*Proc. Camb. phil. Soc.*, April 1958, Vol. 54, Part 2, pp. 247–250.)

538.566 : 535.42] + 534.26 **2538**

Geometrical Theory of Diffraction in Inhomogeneous Media.—B. D. Seckler & J. B. Keller. (*J. acoust. Soc. Amer.*, Feb. 1959, Vol. 31, No. 2, pp. 192–205.) The theory developed by an extension of Fermat's principle and considering smooth convex bodies, is applied to the calculation of the field of diffracted rays.

538.566 : 535.42] + 534.26 **2539**

Asymptotic Theory of Diffraction in Inhomogeneous Media.—B. D. Seckler & J. B. Keller. (*J. acoust. Soc. Amer.*, Feb. 1959, Vol. 31, No. 2, pp. 206–216.) Problems discussed in 2538 above are treated here as boundary-value problems. They are solved exactly and the solutions are expanded asymptotically for high frequencies.

538.566 : 535.43 **2540**

A Note on the Scattering of a Plane Electromagnetic Wave by a Small, Thin-Walled, Cylindrical Dielectric Tube.—J. Y. Wong. (*Canad. J. Phys.*, Jan. 1959, Vol. 37, No. 1, pp. 77–78.)

538.566 : 537.56 **2541**

Nonreciprocal Electromagnetic Wave Propagation in Ionized Gaseous Media.—L. Goldstein. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 19–29. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 804.)

538.569.4 **2542**

Investigation of Relaxation in Magnetic Resonance by the Method of 'Forced Transient Precession'.—I. Solomon. (*C. R. Acad. Sci., Paris*, 5th Jan. 1959, Vol. 248, No. 1, pp. 92–94.) A method is described for measuring directly the relaxation time in liquids.

538.569.4 **2543**

Radiation Damping in Nuclear Magnetic Resonance.—A. Szöke & S. Meiboom. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 585–586.) Observations on a nuclear two-level maser have demonstrated the effect of radiation damping obtained by 'flipping' the magnetization through about 180°.

538.569.4 : 535.33 **2544**

Wide-R.F.-Level R.F. Unit for an NMR Spectrometer.—K. N. Kapur & J. W. McGrath. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 272–274.) This unit makes use of a crystal-controlled oscillator for frequency stability, and positive feedback to increase the effective Q of the sample r.f. coil. The circuit is well suited to saturation studies and relaxation-time measurements in nuclear-magnetic-resonance investigations.

538.569.4 : 538.221 **2545**

Effective Ferrimagnetic Resonance Parameters with Gilbert-Type Relaxation Terms.—R. K. Wangsness. (*Phys. Rev.*, 1st Feb. 1959, Vol. 113, No. 3, pp. 771–772.)

538.569.4 : 538.222 **2546**

Observation of the Overhauser Effect in a Gas in the Presence of a Solid Paramagnetic Material.—J. P. Borel & P. Cornaz. (*C. R. Acad. Sci., Paris*, 1st Dec. 1958, Vol. 247, No. 22, pp. 1988–1990.) Note of nuclear-magnetic-resonance measurements made on propane gas in diphenyl picryl hydrazyl.

538.569.4 : 538.61 : 538.222 **2547**

The Influence of the Saturation of Paramagnetic Resonance Absorption on the Faraday Effect.—H. Wesemeyer & J. M. Daniels. (*Z. Phys.*, 16th Oct. 1958, Vol. 152, No. 5, pp. 591–598.) Experimental investigations on a crystal of Nd $(C_2H_5SO_4)_3.9\ H_2O$ and discussion of results.

538.569.4 : 621.375.9 **2548**

Weak-Field Nuclear-Magnetic-Resonance Maser.—H. Benoit, D. Grivet & H. Ottavi. (*C.R. Acad. Sci., Paris*, 1st Dec. 1958, Vol. 247, No. 22, pp. 1985–1988.) Description of an instrument for nuclear-magnetic-resonance investigations based on proton spin resonance in circulating benzene.

539.2 : 538.222 **2549**

Method of Treating Zeeman Splittings of Paramagnetic Ions in Crystalline Fields.—G. F. Koster & H. Statz. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 445–454.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

522.1 **2550**

The Royal Greenwich Observatory.—R. v. d. R. Woolley. (*Nature, Lond.*, 7th March 1959, Vol. 183, No. 4662, pp. 640–643.) Reference is made to the national time service, associated equipment and land-line links with other standards.

523.16 **2551**

Low-Energy Corpuscular Radiation at High Latitudes.—K. D. Cole. (*Nature, Lond.*, 14th March 1959, Vol. 183, No. 4663, p. 738.) An explanation based on an acceleration mechanism.

523.16 : 550.389.2 : 629.19 **2552**

Origin of the Radiation near the Earth Discovered by means of Satellites.—T. Gold. (*Nature, Lond.*, 7th Feb. 1959, Vol. 183, No. 4658, pp. 355–358.) See also 2217 of July (Kellogg).

523.16 : 550.389.2 : 629.19 **2553**

Radiation around the Earth to a Radial Distance of 107 400 km.—J. A. Van Allen & L. A. Frank. (*Nature, Lond.*, 14th Feb. 1959, Vol. 183, No. 4659, pp. 430–434.) Radiation measurements made during the flight of Pioneer III on 6th December 1958 are discussed, and a brief description is given of the two Geiger-Müller detector tubes used and the associated telemetry equipment. See also 2217 of July (Kellogg).

523.164.3 **2554**

Anomalous Night-Time Reception of a Major Solar Radio Burst.—A. G. Smith, T. D. Carr & W. H. Perkins. (*Nature, Lond.*, 28th Feb. 1959, Vol. 183, No. 4661, pp. 597–598.) A report of a noise burst received on five different arrays operating on frequencies between 18 and 27·6 Mc/s, at 0235 U.T. on 8th March 1958 in Florida. Possible modes of propagation are discussed.

523.164.32 **2555**

Time Relationship of Metre-Wave and 3-cm-Wave Bursts.—M. R. Kundu. (*Nature, Lond.*, 11th April 1959, Vol. 183, No. 4667, pp. 1047–1048.) Metre-wave bursts have an average delay of 3 min, the delay being less for bursts of small apparent diameter.

523.42 : 621.396.96 **2556**

Radar Echoes from Venus.—R. Price, P. E. Green, Jr, J. J. Goblick, Jr, R. H. Kingston, L. G. Kraft, Jr, G. H. Pettengill, R. Silver & W. B. Smith. (*Science*, 20th March 1959, Vol. 129, No. 3351, pp. 751–753.) Report on tests made on 10th and 12th February 1958 using specially coded pulse trains at 440 Mc/s, and a maser-type amplifier for reception.

523.75 : 523.165 **2557**

Cosmic-Ray Increases associated with Solar Flares.—L. C. Towle & J. A. Lockwood. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 641–647.) Data from a neutron monitor for the years 1956–1957 have been studied.

523.75 : 550.385.4 **2558**

Characteristics of Solar Outbursts to Excite Geomagnetic Storms.—K. Sinno. (*J. Radio Res. Labs, Japan*, Jan. 1959, Vol. 6, No. 23, pp. 17–20.) For major outbursts a pronounced 'first part' before the time of maximum flare is related to short-wave fade-out, while a large 'second part' will cause a geomagnetic storm. It is concluded that the second part of the outburst provides evidence of a corpuscular cloud.

523.78 : 551.524 **2559**

Fluctuations of Temperature accompanying the Solar Eclipse.—K. Hirao, K. Akita & I. Shiro. (*J. Radio Res. Labs, Japan*, Jan. 1959, Vol. 6, No. 23, pp. 47–55.) During the eclipse there was a depression of the r.m.s. value of the fluctuations which was accompanied by a temperature inversion with a steep gradient as the eclipse passed the maximum blackout period. The r.m.s. value of the temperature fluctuations between two points vertically separated was found to increase with diminishing inversion and to continue to increase as the gradients changed from positive to negative values.

550.385 : 523.75 **2560**

Some Remarks on the Interaction of Solar Plasma and the Geomagnetic Field.—J. W. Warwick. (*J. geophys. Res.*, April 1959, Vol. 64, No. 4, pp. 389–396.) An explanation of the small scale of magnetic disturbances on the earth's surface is that currents flow in plasma sheets and move along the curved magnetic lines of force. These sheets are distant less than one earth's radius from the surface and would produce perturbation fields consistent with the size and direction necessary to explain storm-time geomagnetic variations.

550.385 : 523.78 **2561**

Preliminary Report on the Effect of the Solar Eclipse on April 19, 1958, on the Geomagnetic Field.—T. Rikitake, S. Uyeda, T. Yukutake, I. Tanaoka & E. Nakagawa. (*Rep. Ionosphere Res. Japan*, June 1958, Vol. 12, No. 2, pp. 174–181.) The changes in the *D* and *H* components are consistent with the decrease in E-Layer conductivity; *Z*-component changes are influenced by earth currents.

550.385 : 523.78 **2562**

Effect of the Solar Eclipse, 19th April 1958, on the Geomagnetic Field and Earth Currents.—Y. Yamaguchi, N. Banno, H. Ōshima & T. Araki. (*Rep. Ionosphere Res. Japan*, June 1958, Vol. 12, No. 2, pp. 182–187.)

550.389.2 : 629.19 **2563**

The Sun's Artificial Planet.—(*Brit. Commun. Electronics*, March 1959, Vol. 6, No. 3, pp. 200–203.) General scientific information, based on a Russian report of the space rocket launched on 2nd January 1959.

550.389.2 : 629.19 **2564**

Two Atmospheric Effects in the Orbital Acceleration of Artificial Satellites.—L. G. Jacchia. (*Nature, Lond.*, 21st Feb. 1959, Vol. 183, No. 4660, pp. 526–527.) A discussion of oscillations in orbital acceleration and the variations of orbital acceleration with angular distance ψ of the perigee from the sub-solar point.

550.389.2 : 629.19 **2565**

Atmospheric Tides and Earth Satellite Observations.—D. G. Parkyn: G. V. Groves. (*Nature, Lond.*, 11th April 1959, Vol. 183, No. 4667, pp. 1045–1047.) Harmonic tidal waves are suggested as the cause of the major anomalies in orbital observations; this explanation differs from that of Groves (1543 of May), who comments in reply.

550.389.2 : 629.19 **2566**

The I.G.Y. Optical Satellite Tracking Program as a Source of Geodetic Information.—F. L. Whipple & J. A. Hynek. (*Ann. Géophys.*, July/Sept. 1958, Vol. 14, No. 3, pp. 326–328. In English.) A note on the visual and photographic tracking program and the subsequent computations.

550.389.2 : 629.19 **2567**

A New Value for the Earth's Flattening, derived from Measurements of Satellite Orbits.—D. G. King-Hele & R. H. Merson. (*Nature, Lond.*, 28th March 1959, Vol. 183, No. 4665, pp. 881–882.) Data obtained from observations of 1957 β and 1958 $\beta 2$ have been combined, and give a value for the earth's flattening of $1/(298{\cdot}20 \pm 0{\cdot}3)$. See also 792 of March (Merson & King-Hele).

550.389.2 : 629.19 **2568**

Determination of Satellite Orbits from Radio Tracking Data.—I. Harris, R. Jastrow & W. F. Cahill. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 851–854.) "A computer program has been developed which permits an approximate determination of a satellite orbit from a minimal amount of tracking data."

550.389.2 : 629.19 **2569**

Observations of Sputnik III.—P. F. Checcacci & C. Carreri. (*Ricerca sci.*, Jan. 1959, Vol. 29, No. 1, pp. 72–73.) Radio observations of satellite 1958 $\delta 2$ for the period 19th May–19th July 1958 are tabulated. See also 1192 of April.

550.389.2 : 629.19 **2570**

Recordings of Transmissions from the Satellite 1958 $\delta 2$ at the Antenna Laboratory, The Ohio State University.—T. G. Hame & E. M. Kennaugh. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 991–992.) Irregularities in 20-Mc/s signals are attributed to polarization effects caused by variations in magnetic field near the observing point. Some periodic fading may be caused by a defect in the satellite aerial.

550.389.2 : 629.19 : 551.510.53 **2571**

Irregularities in the Density of the Upper Atmosphere: Results from

Satellites.—D. G. King-Hele & D. M. C. Walker. (*Nature, Lond.*, 21st Feb. 1959, Vol. 183, No. 4660, pp. 527–529.) Results of observations of earth-satellite orbits indicate that irregularities in air density, especially at heights between 180 and 225 km, tend to recur at intervals of about 28 days.

550.389.2 : 629.19 : 551.510.535 **2572**
Deduction of Ionospheric Electron Content from the Faraday Fading of Signals from Artificial Earth Satellites.—W. T. Blackband, B. Burgess, I. L. Jones & G. J. Lawson. (*Nature, Lond.*, 25th April 1959, Vol. 183, No. 4669, pp. 1172–1174.) A method is described for determining horizontal variations in the ionosphere both above and below maximum ionization. Results are compared with values of electron content deduced from ionograms.

550.389.2 : 629.19 : 621.396.81.087.4 **2573**
Recording Radio Signals from Earth Satellites.—G. H. Munro & L. H. Heisler. (*Nature, Lond.*, 21st March 1959, Vol. 183, No. 4664, pp. 809–810.) A method is described for presenting simultaneously on moving film the information necessary for Doppler frequency-shift measurements, for frequency sampling, and for the study of signal intensity variations.

551.510.52 : 621.396.11 **2574**
Models of the Atmospheric Radio Refractive Index.—Bean & Thayer. (See 2725.)

551.510.53 **2575**
Geophysical Effects associated with High-Altitude Explosions.—P. J. Kellogg, E. P. Ney & J. R. Winckler. (*Nature, London.*, 7th Feb. 1959, Vol. 183, No. 4658, pp. 358–361.) Effects likely to be associated with high-altitude explosions are discussed and related to observations of a nuclear and a chemical explosion.

551.510.53 : 523.75 **2576**
Observations of Intense Ionization of Long Duration below 50 km Altitude after some Strong Solar Flares.—B. Hultqvist & J. Ortner. (*Nature, Lond.*, 25th April 1959, Vol. 183, No. 4669, pp. 1179–1180.)

551.510.535 **2577**
Turbulence in the Upper Atmosphere.—N. Matuura & T. Nagata. (*Rep. Ionosphere Res. Japan*, June 1958, Vol. 12, No. 2, pp. 147–159.) Ionospheric irregularities are studied in terms of turbulence theory. The calculated scales of the irregularities and their random velocities are compared with experimental observations.

551.510.535 **2578**
Sporadic E and the F_2 Layer.—T. W. Bennington. (*Wireless World*, June 1959, Vol. 65, No. 6, pp. 262–263.) A comparison of curves of E_s percentage time with curves giving seasonal fluctuations of F_2-layer ionization for four stations in both hemispheres indicates that sporadic E may be caused by a downward drift of F_2 ionization.

551.510.535 **2579**
A Theoretical Consideration of the Electron and Ion Density Distributions in the Lower Portion of the F Region.—T. Yonezawa, H. Takahashi & Y. Arima. (*J. Radio Res. Labs, Japan*, Jan. 1959, Vol. 6, No. 23, pp. 21–46.) Assuming that the photoionization processes in the ionosphere are: $O^+ + O_2 \rightarrow O + O_2^+$; $O_2^+ + e \rightarrow O' + O''$, and $O^+ + N_2 \rightarrow N + NO^+$; $NO^+ + e \rightarrow N' + O'$, the physics of the lower F region can be explained fairly well. The appearance and disappearance of the F_1 layer with solar zenith angle and solar activity can be understood. No values of reaction coefficients can be found which are consistent with both the ion density distribution in the ionosphere and laboratory estimates of the coefficients.

551.510.535 **2580**
Occurrence of Giant Travelling Ionospheric Disturbances at Night.—L. H. Heisler. (*Nature, Lond.*, 7th Feb. 1959, Vol. 183, No. 4658, pp. 383–384.) Disturbances previously thought to be confined to winter daylight hours (3453 of 1958) have been observed on night ionosonde records. Characteristics of the phenomenon are discussed.

551.510.535 **2581**
Structure and Movement of Large Inhomogeneities in the Ionospheric F_2 Layer.—V. D. Gusev, L. A. Drachev, S. F. Mirkotan, Yu. V. Berezin, M. P. Kiyanovskiĭ, M. B. Vinogradova & T. A. Gaĭlit. (*Dokl. Ak. Nauk S.S.S.R.*, 11th Dec. 1958, Vol. 123, No. 5, pp. 817–820.) Simultaneous measurements have been made using three receivers 30–40 km apart, of the phase and angle of arrival of waves reflected from the F_2 layer. The size and shape of inhomogeneities and drift velocities have been calculated.

551.510.535 : 523.16 **2582**
A New Ionospheric Phenomenon.—H. J. A. Chivers & H. W. Wells. (*Nature, Lond.*, 25th April 1959, Vol. 183, No. 4669, p. 1178.) Report of an isolated event recorded at Jodrell Bank at about 1400 U.T. on 25th March 1959 by five separate receivers, each monitoring various sectors of the sky on slightly different frequencies near 80 Mc/s. A large increase of the noise level was recorded by three receivers, a small increase by another and a decrease by another.

551.510.535 : [523.3 + 523.755 **2583**
Radio Reflexions from the Moon and Solar Corona.—O. Burkard. (*Nature, Lond.*, 25th April 1959, Vol. 183, No. 4669, p. 1180.) The formula for the rotation of the plane of polarization given by Daniels & Bauer (3455 of 1958) is revised for the case when, with low solar activity, the electron density in the ionosphere has decreased and the electrons in the space around the earth cannot be neglected.

551.510.535 : 550.38 **2584**
A Study of Noon F_2 Ionization in relation to Geomagnetic Coordinates.—J. N. Bhar & P. Dhar Bhowmik. (*Indian J. Phys.*, Jan. 1959, Vol. 33, No. 1, pp. 1–17.) Earlier work of Bhar (2129 of 1957) has been extended to cover data collected between 1952 and 1955, an analysis of which indicates that for constant χ, the noon F_2 ionization varies with magnetic dip and geomagnetic latitude. The shape of the curve is related to the sunspot cycle and shows two maxima spaced asymmetrically on each side of the geomagnetic equator.

551.510.535 : 550.385 **2585**
Ionospheric Heating by Hydromagnetic Waves.—A. J. Dessler. (*J. geophys. Res.*, April 1959, Vol. 64, No. 4, pp. 397–401.) The dissipation of hydromagnetic waves in the ionosphere causes heating; this heating is greatest between heights of 150 and 200 km. It appears that the heating only becomes important during magnetic storms; it may explain the lifting of the F region and the decrease in critical frequency during these periods. Hydromagnetic heating cannot, however, explain the variations in the horizontal component of the earth's field during the main phase of magnetic storms.

551.510.535 : 550.385 **2586**
Effect of Magnetic Activity on Drifts of the F_2 Region.—B. R. Rao, E. B. Rao & Y. V. R. Murty. (*Nature, Lond.*, 7th March, 1959, Vol. 183, No. 4662, pp. 667–668.) An analysis of observations made at Waltair shows an approximately linear decrease of mean drift speed with increase of magnetic activity (K) in the range 0–6.

551.510.535 : 621.396.11 **2587**
Sporadic E at V.H.F. in the U.S.A.—R. M. Davis, Jr, E. K. Smith & C. D. Ellyett. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 762–769.) The occurrence of sporadic-E propagation in the years 1952–1955 on the Cedar Rapids–Sterling path (1 240 km) at frequencies near 28 and 50 Mc/s has been analysed. The dependence of the received power on time of day, season of year, sunspot cycle and frequency has been studied. A relation between oblique-signal intensity and vertical-incidence data has been found.

551.510.535 : 621.396.11 **2588**
On the S.W.F. Phenomenon (Dellinger Effect) and f_{min} in the World-Wide Distribution.—I. Kasuya, Y. Hakura & H. Hojo. (*J. Radio Res. Labs, Japan*, Jan. 1959, Vol. 6, No. 23, pp. 1–15.) Assuming a Chapman distribution of ionization for the D region it is shown that f_{min} is a function of solar zenith angle and magnetic latitude while the increase of f_{min} during a sudden ionospheric disturbance depends on solar zenith angle only. Observational results are analysed and the effective factor by which solar radiation is increased during a s.i.d. is estimated.

551.510.535 : 621.396.11 : 551.594.6 **2589**
Ionospheric Reflection Coefficients at V.L.F. from Sferics Measurements.—A. G. Jean, Jr, L. J. Lange & J. R. Wait. (*Geofis. pura appl.*, Sept.–Dec. 1957, Vol. 38, pp. 147–153. In English.) The

magnitude of the ionospheric reflection coefficient as a function of frequency is calculated from individual spectral analysis of the ground wave and the once-reflected sky wave. Results are compared with those obtained using an ionospheric model. The success of the method depends upon the accurate location of the source and the correct selection of strokes which contain only appreciable vertical currents in the discharge column.

551.510.535 : 621.396.11.029.45 **2590**

Diurnal Change of Ionospheric Heights Deduced from Phase Velocity Measurements at V.L.F.—Wait. (See 2735.)

551.510.535 : 621.396.11.029.62 **2591**

IGY Observations of F-Layer Scatter in the Far East.—Bateman, Finney, Smith, Tveten & Watts. (See 2733.)

551.594.5 **2592**

Auroral Activity at Low Latitudes.—D. Barbier. (*Ann. Géophys.*, July/Sept. 1958, Vol. 14, No. 3, pp. 334–355.) Spectroscopic observations made in southern France and at Tamanrasset during 1956–1957 are described.

551.594.5 : 550.385 **2593**

Movement of Auroral Echoes and the Magnetic-Disturbance Current System.—R. S. Unwin. (*Nature, Lond.*, 11th April 1959, Vol. 183, No. 4667, pp. 1044–1045.) A brief review of the findings of other investigators and a report of observations made at Invercargill, New Zealand, on 55 Mc/s. A preliminary analysis indicates that the ionization giving rise to v.h.f. radar echoes occurs in the most intense portion of the disturbance current system, and that the observed movements of electrons can account for the magnitude of the disturbance vector.

551.594.6 : 621.396.11 : 551.510.535 **2594**

'Whistler Mode' Echoes Remote from the Conjugate Point.—Dowden & Goldstone. (See 2731.)

550.389.2 : 629.19 **2595**

Scientific Uses of Earth Satellites. [Book Review]—J. A. Van Allen (Ed.). Publishers: University of Michigan Press, Michigan, and Chapman & Hall, London, 1958, 316 pp., 75s. (*Nature, Lond.*, 4th April, 1959, Vol. 183, No. 4666, p. 963.)

LOCATION AND AIDS TO NAVIGATION

534.88 **2596**

A Theory of Active Sonar Detection.—J. L. Stewart & E. C. Westerfield. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 872–881.) An investigation is carried out to determine the type of signal waveform or coding that will give the best echo/noise and echo/reverberation ratios.

534.88 : 621.398 **2597**

Survey of Underwater Missile Tracking Instrumentation.—D. T. Barry & J. M. Formwalt. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 970–977.)

621.396.93 : 621.396.11.029.45 **2598**

V.L.F. Propagation Measurements for the Radux-Omega Navigation System.—Casselman, Heritage & Tibbals. (See 2736.)

621.396.933 **2599**

A Lightweight and Self-Contained Airborne Navigational System.—R. K. Brown, N. F. Moody, P. M. Thompson, R. J. Bibby, C. A. Franklin, J. H. Garton & J. Mitchell. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 778–807.) Automatic computing techniques are used to give steering and positional information from input data obtained from an inertial north indicator, a Doppler navigational radar and an airspeed indicator.

621.396.933.2 **2600**

Direction Finder with Automatic Read-Out.—J. F. Hatch & D. W. G. Byatt. (*Electronics*, 17th April 1959, Vol. 32, No. 16, pp. 52–54.) The accuracy of an h.f. direction-finder can be improved by averaging bearings; circuits are described which give automatic read-out and averaging.

621.396.934 **2601**

Miss-Distance Indicator scores Missile Accuracy.—J. A. Adams. (*Electronics*, 17th April 1959, Vol. 32, No. 16, pp. 42–45.) Transponder and aerial systems in missile and target form a space-coupled oscillatory system. The oscillation is modulated at a frequency which depends on their separation distance.

621.396.96 **2602**

Problems connected with Echo Filtering in Radar.—V. Tiberio. (*Ricerca sci.*, Dec. 1958, Vol. 28, No. 12, pp. 2464–2481.) Various systems of filtering combined with carrier stabilization are described and the resulting increase in range, even in the presence of jamming, is discussed. For comment by U. Pizzarelli, with details of calculations relating bandwidth and sweep velocity, see *ibid.*, pp. 2585–2590.

621.396.96 : 551.5 **2603**

Meteorological 'Angel' Echoes.—D. Atlas. (*J. Met.*, Feb. 1959, Vol. 16, No. 1, pp. 6–11.) An analysis of meteorological conditions associated with radar echoes suggests four sources of reflection: thermal columns below cumulus clouds; cumulus cloud boundaries; layers associated with sharp vertical gradients or minima in moisture; and, under conditions of anomalous propagation, boundary surfaces between differentially moistened surface air over adjacent cold and warm water.

621.396.96 : 551.5 **2604**

Study of Mesosystems associated with Stationary Radar Echoes.—T. Fujita. (*J. Met.*, Feb. 1959, Vol. 16, No. 1, pp. 38–52.)

621.396.96 : 621.397.24 **2605**

Scan Converter aids Phone-Line Radar Relay.—H. W. Gates & A. G. Gatfield. (*Electronics*, 17th April 1959, Vol. 32, No. 16, pp. 48–51.) A scan-converter storage tube is substituted for the radar tube; its reading gun scans its screen at a slow rate and the output is used to modulate a carrier which can be transmitted over a telephone line.

621.396.962.3(083.57) **2606**

Predicting Accurate Radar Ranges.—L. Young. (*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 58–61.) Formulae and a chart are given for determining the maximum range of a pulse radar equipment scanning in azimuth.

621.396.962.33 **2607**

Frequency Stability Requirements on Coherent Radar Oscillators.—M. M. Brady. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, pp. 1001–1002.) A discussion of stability criteria for Doppler radar systems.

621.396.962.33 **2608**

The Performance of Doppler Navigation Systems.—T. G. Thorne & J. A. Billings. (*Brit. Commun. Electronics*, March 1959, Vol. 6, No. 3, pp. 176–179.) A review is given of the performance of military Doppler equipment under various conditions and reference is made to the important design features that control the accuracy and reliability of Doppler systems.

621.396.968 **2609**

Phenomena of Scintillation Noise in Radar Tracking Systems.—J. H. Dunn, D. D. Howard & A. M. King. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 855–863.) The effects of different components of target noise on the performance of tracking systems have been studied theoretically and experimentally; suggestions are made for minimizing these effects.

621.396.969.11 **2610**

The C.A.A. Doppler Omnirange.—S. R. Anderson & R. B. Flint. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 808–821.) Theoretical and experimental investigations show that the Doppler VOR is compatible with the v.h.f. omnirange and gives a 7 : 1 improvement in respect of site errors. See also 1897 of June (Hansel).

621.396.969.11 : 621.396.822 **2611**

Noise-Modulated Distance Measuring Systems.—B. M. Horton. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 821–828.) An analysis is given for distance-measuring systems involving correlation between the random noise modulation on the transmitted and reflected signals. Such systems are not subject to the ambiguities present in periodically modulated signals, nor to the 'fixed error', and are suitable for measuring distances down to a few feet.

621.396.969.13 **2612**

Precision Radar Height Finder.—(*Engineering, Lond.*, 31st Oct. 1958, Vol. 186, No. 4834, p. 560.) Description of the Marconi Type-S 244 height finder and ancillary equipment. The vertical angle of the aerial is measured relative to the ground, and the absolute height accuracy obtainable is better than ±1 700 ft at 150 miles.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.583 : 621.385.032.14 **2613**

Barium Getter Films.—J. J. B. Fransen & H. J. R. Perdijk. (*Philips tech. Rev.*, 30th April 1958, Vol. 19, No. 10, pp. 290–300.) Experiments to investigate the action and properties of getter films are described.

535.215 : 537.311.33 **2614**

A Lateral Photovoltaic Effect in *p-n* Junctions.—P. Gosar. (*C. R. Acad. Sci., Paris*, 1st Dec. 1958, Vol. 247, No. 22, pp. 1975–1977.) A theoretical study of a one-dimensional model in which nonuniformity of illumination is due to shadows formed by soldered contacts of unequal size. See also 2639 of 1957 (Wallmark).

535.215 : 539.23 : 546.817.221 **2615**

Measurement of the Reflection Factor, the Optical Absorption and the Spectral Response of Photoconductive Layers. Case of Oxidized Lead Sulphide.—V. Schwetzoff. (*C. R. Acad. Sci., Paris*, 10th Dec. 1958, Vol. 247, No. 23, pp. 2117–2119.)

535.215 : [546.482.21 + 546.482.31 **2616**

Dependence of Induced Conductivity on Electron Energy in Thin Films of Cadmium Sulphide and Cadmium Selenide Bombarded by Slow Electrons.—Li-Chzhi-Tszyan'. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 77–81.) The method and apparatus are described. Measurements have been made on films of thickness ≈10^{-4} cm under bombardment by electrons of energy 0–15 eV. An irregular step-like function is observed in the variation of the induced conductivity with the energy of bombarding electrons. The energy distance between two consecutive steps is 2·5 eV for CdS and 1·8 eV for CdSe. See also *ibid.*, pp. 82–88.

535.215 : 546.482.21 **2617**

On the Irradiation of Photoconductive Single Crystals of Cadmium Sulphide by Protons at 1·4 MeV.—R. Barjon, C. Brachet, M. Lambert, M. Martineau & J. Schmouker. (*C. R. Acad. Sci., Paris*, 5th Jan. 1959, Vol. 248, No. 1, pp. 83–86.) Four phenomena are noted: green luminescence, induced conductivity, a permanent modification of the photoconductivity and modification of the spectral response. See also 1594 of May (Martineau).

535.215 : 546.482.21 **2618**

Research on Fluorescent Emission Lines and on Luminous Absorption Lines in Pure Cadmium Sulphide, Cooled to 4·2°K.—M. Bancie-Grillot, E. F. Gross, E. Grillot & B. S. Razbirine. (*C. R. Acad. Sci., Paris*, 5th Jan. 1959, Vol. 248, No. 1, pp. 86–89.)

535.215 : 546.87 : 539.23 **2619**

The Temperature Dependence of the Spectral Photoelectric Electron Emission of Semiconducting Bismuth Films.—R. Suhrmann, G. Wedler & E. A. Dierk. (*Z. Phys.*, 27th Oct. 1958, Vol. 153, No. 1, pp. 96–105.)

535.342-15 : 546.87 **2620**

De Haas-van Alphen-Type Oscillations in the Infrared Transmission of Bismuth.—W. S. Boyle & K. F. Rodgers. (*Phys. Rev. Lett.*, 15th April 1959, Vol. 2, No. 8, pp. 338–339.)

535.37 : 546.482.21 **2621**

Some Properties of Green and Red-Green Luminescing CdS.—Y. T. Sihvonen, D. R. Boyd & C. D. Woelke. (*Phys. Rev.*, 15th Feb. 1959, Vol. 113, No. 4, pp. 965–968.)

535.37 : 546.482.21 : 538.6 **2622**

Influence of a Magnetic Field on the Lines of Blue Fluorescence or Luminous Absorption of Certain Crystals of Pure Cadmium Sulphide Cooled to 4·2 °K.—E. F. Gross, E. Grillot, B. P. Zakhartchenia & M. Bancie-Grillot. (*C. R. Acad. Sci., Paris*, 12th Jan. 1959, Vol. 248, No. 2, pp. 213–216.) A Zeeman splitting has been observed on three lines. There was no diamagnetic displacement.

535.376 **2623**

Phosphor with Fluorescence Larger than the Energy Gap.—S. P. Keller & G. D. Pettit. (*Phys. Rev.*, 1st Feb. 1959, Vol. 113, No. 3, pp. 785–786.) An emission which is energetically greater than the energy gap is observed at 770 °K with a Pr^{+3}-activated SrS phosphor.

535.376 **2624**

On Cathodothermoluminescence.—A. Bril, H. A. Klasens & T. J. Westerhof. (*Physica*, Oct. 1958, Vol. 24, No. 10, pp. 821–827.) When phosphors are heated from −180° C to 20°C in a demountable tube strong maxima and minima of the light output are observed. If sealed and thoroughly degassed tubes are used the effect disappears. Gas adsorption on the phosphors is a probable explanation of the effect.

535.376 : 546.472.21 **2625**

Preparation of Zinc Sulphide Single Crystals.—T. Matsumura, H. Fujisaki & Y. Tanabe. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A.*, Dec. 1958, Vol. 10. No. 6, pp. 459–471.) Details of preparation by gas reaction and sublimation methods.

537.226 **2626**

Dielectric Properties of some Polycrystalline Stannates and Cerates.—B. Piercy. (*Trans. Faraday Soc.*, Jan. 1959, Vol. 55, Part 1, pp. 39–51.) The substances have a perovskite structure but do not exhibit ferroelectric or antiferroelectric properties. A negative temperature coefficient of capacitance was observed in $BaSnO_3$ below room temperature.

537.226 : 546.431.824-31 **2627**

Temperature Dependence of the Breakdown Field of Barium Titanate.—P. H. Fang & W. S. Brower. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 456–458.) The breakdown field in general decreases with increasing temperature but shows minima of the critical temperatures of the phase transformations.

537.226 : 546.431.824-31 **2628**

Thermal Conductivity of $BaTiO_3$ Ceramics.—I. Yoshida, S. Nomura & S. Sawada. (*J. phys. Soc. Japan*, Dec. 1958, Vol. 13, No. 12, pp. 1550–1551.)

537.226 : 546.431.824-31 : 536.2.082.74 **2629**

Coefficient of Thermal Expansion of Barium Titanate.—J. R. G. Keyston, J. D. Macpherson & E. W. Guptill. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 246–248.) The coefficient of expansion can be found by measurement of the resonance frequency, over a temperature range of −200 to 100°C, of one or more modes of a microwave cavity constructed of $BaTiO_3$.

537.227/.228 **2630**

Charge Release of Several Ceramic Ferroelectrics under Various Temperature and Stress Conditions.—L. W. Doremus. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 921–924.) The results are given of measurements on the polarization change, piezoelectric coefficient and pyroelectric charge release for several polycrystalline compounds.

537.227 **2631**

On the Effect of Polarization on $Pb_3NiNb_2O_9$-$Pb_3MgNb_2O_9$ Solid Solutions.—G. A. Smolenskiĭ, A. I. Agranovskaya & S. N. Popov. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 167–168.) The dependence of permittivity and loss angle on temperature in the range −200° to + 250°C is shown graphically.

537.227 **2632**

New Ferroelectric with Composition of the Type $A_2{}^{2+}(B_I{}^{3+}B_{II}{}^{5+})O_6$: Part 1.—G. A. Smolenskiĭ, V. A. Isupov & A. I. Agranovskaya. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 170–171.) The temperature dependence of the dielectric constant and loss angle for Pb_2ScNbO_6 and Pb_2ScTaO_6 from −150° to +200°C is shown graphically.

537.227 : 538.569.4 **2633**

Electron Paramagnetic Resonance of Manganese IV in $SrTiO_3$.—K. A. Müller. (*Phys. Rev. Lett.*, 15th April 1959, Vol. 2, No. 8, pp. 341–343.)

537.227 : 546.431.824-31 **2634**

Direct Observation of Antiparallel Domains during Polarization Reversal in Single-Crystal Barium Titanate.—R. C. Miller & A. Savage. (*Phys. Rev. Lett.*, 1st April 1959, Vol. 2, No. 7, pp. 294–296.) The domains have been observed by looking along the ferroelectric axis through the electrodes (either transparent aqueous lithium chloride or semitransparent metal films). The sample was mounted between crossed polaroids, and white light was used.

537.227 : 546.431.824-31 : 621.318.57 **2635**

A Possible Model for the Switching of Barium Titanate Crystals.—J. C. Burfoot. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 641–649.) "It is suggested that an adequate explanation of the switching current in barium titanate single crystals can be given by a domain model which includes an effective wall mass and a 'viscous' opposition, but no depolarizing field and no specific representation of impurities or lattice imperfections."

537.228.1 : 534.133 : 538.566.029.64 **2636**

Attenuation of Hypersonic Waves in Quartz.—G. E. Bömmel & K. Dransfeld. (*Phys. Rev. Lett.*, 1st April 1959, Vol. 2, No. 7, pp. 298–299.) Report of measurements of the acoustic absorption in crystalline quartz at frequencies between 1 and 4 kMc/s from room temperature to 4·2°K.

537.311.33 **2637**

Characteristics of Stationary Electronic Processes in Semiconductors.—G. M. Guro. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 3–12.) Examination of the bipolar diffusion in which the diffused length does not depend on the concentration of traps. It is shown that the equilibrium charge density produced by nonuniform generation is determined not only by electron and hole mobility but also by the capture of minority carriers by traps and recombination centres. The charge density can be of either sign, and is zero for a particular fractional occupancy of the trap. See 2443 of 1958.

537.311.33 **2638**

The Application of the Thermodynamics of Irreversible Processes to Conduction Phenomena in Semiconductors.—W. Czaja. (*Helv. phys. Acta*, 10th March 1959, Vol. 32, No. 1, pp. 1–23. In German.) Isothermal and non-isothermal effects are investigated and general equations for the behaviour of an isothermal *p-n* junction are derived. The thermal conductivity of a homogeneous semiconductor under various conditions is calculated. See also 2693 of 1958 (van Vliet).

537.311.33 **2639**

Theory of the Field Effect.—I. V. Boiko. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 13–15.) Mathematical analysis based on the results of the general theory of the field effect for the limiting case when the mean free path of the carriers tends to zero.

537.311.33 **2640**

Field Effect at High Frequency.—F. Berz. (*J. Electronics Control*, Feb. 1959, Vol. 6, No. 2, pp. 97–112.) Theory is developed using the same basic assumptions as Garrett (153 of 1958) with more complete conditions at the back surface. Results are illustrated by a numerical example.

537.311.33 **2641**

The Behaviour of some Impurities in III-V Compounds.—J. T. Edmond. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 622–627.) Investigation of the behaviour, as acceptors or donors, of Mg, Zn, Cd, Si, Ge, Sn, Pb, S, Se and Te in InSb, InAs, GaSb and GaAs.

537.311.33 **2642**

Theoretical Transport Coefficients for Polar Semiconductors.—R. T. Delves. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 572–576.) Calculation of mobility, thermoelectric power and Hall coefficient for carriers scattered by optical-mode lattice vibrations.

537.311.33 : 535.215 **2643**

Influence of Surface Recombination on Photoconductivity of Semiconductors.—G. L. Bir. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 67–76.) Results of an investigation show that the magnitude of photoconductivity can be determined from the amount of surface recombination and also from the properties of the surface barrier. This explains the great sensitivity of photoconductive cells to surface treatment. In the case of an anisotropic layer the electric field decreases the surface recombination velocity and increases the photoconductivity.

537.311.33 : 538.214 **2644**

Theory of Impurity Paramagnetism in Semiconductors at Low Temperature.—J. Seiden. (*C. R. Acad. Sci., Paris*, 22nd Dec. 1958, Vol. 247, No. 25, pp. 2313–2315.) Interaction between non-ionized impurities as a function of concentration is discussed and the susceptibility is calculated approximately.

537.311.33 : 538.22 **2645**

Magnetic Properties of $A^{III}B^{V}$ Compounds.—G. A. Busch & R. Kern. (*Helv. phys. Acta*, 10th March 1959, Vol. 32, No. 1, pp. 24–57. In German.) The magnetic susceptibility of Si, GaP, GaAs, GaSb, InP, $InP_{0\cdot2}As_{0\cdot8}$, InAs and InSb was measured in the temperature range from 60°K to below the respective melting point. A model is proposed to account approximately for the abnormal characteristics obtained for InSb, but quantitative agreement with experimental results is not achieved.

537.311.33 + 538.221] : 539.2 **2646**

Behaviour of Semiconductor and Magnetic Materials in Radiation Environment.—A. Boltax. (*Elect. Mfg*, March 1959, Vol. 63, No. 3, pp. 90–95.) Results of irradiation tests on Ge, Si, Cu_2O, and seven typical magnetic-core materials are summarized in graphs and charts. Techniques for minimizing radiation effects are noted.

537.311.33 : 539.2 **2647**

Cyclotron Resonance in Semiconductors with Complex Equipotential Surfaces.—Yu. A. Firsov. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 44–61.)

537.311.33 : 546.23 **2648**

Investigation of the Influence of Annealing Time on the Thermoelectric Power of Polycrystalline Selenium.—D. Vidal & G. Blet. (*C. R. Acad. Sci., Paris*, 10th Dec. 1958, Vol. 247, No. 23, pp. 2109–2110.)

537.311.33 : [546.28 + 546.289 **2649**

Influence of Deformation on the Energy Spectrum and Electrical Properties of *p*-Type Germanium and *p*-Type Silicon.—G. E. Pikus & G. L. Bir. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 154–156.) An expression is derived for the energy spectrum of a deformed crystal.

537.311.33 : 546.28 **2650**

Stabilization of Silicon Surfaces by Thermally Grown Oxides.—M. M. Atalla, E. Tannenbaum & E. J. Scheibner. (*Bell Syst. tech. J.*, May 1959, Vol. 38, No. 3, pp. 749–783.) A detailed study has been made of the stable surfaces obtained with the system Si-SiO_2. Information is given on the thermal oxidation process and properties of the oxide, the electronic properties of the resulting interface, the practical application of the process and resulting device characteristics.

537.311.33 : 546.28 **2651**

Diffusion of Phosphorus into Silicon under Conditions of Controlled Vapour Pressure.—M. J. Coupland. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 577–584.) Detailed report of an experimental study of the relation between surface concentration and ambient vapour pressure. See 3518 of 1958.

537.311.33 : 546.28 **2652**

Breakdown in Silicon *p-n* Junctions.—J. Shields. (*J. Electronics Control*, Feb. 1959, Vol. 6, No. 2, pp. 130–148.) Some useful empirical relations for the variation of effective ionization coefficients associated with the avalanche breakdown phenomena in *p-n* junctions are derived from measurements on Si alloy junctions.

537.311.33 : 546.289 **2653**

Large-Scale Preparation of Ultra-pure Germanium.—J. M. Wilson. (*Research, Lond.*, Feb. 1959, Vol. 12, No. 2, pp. 47–53.) Chemical and physical purification processes are described.

537.311.33 : 546.289 **2654**

Plastic Creep of Germanium Single Crystals.—H. G. Van Bueren. (*Physica*, Oct. 1958, Vol. 24, No. 10, pp. 831–837.)

537.311.33 : 546.289 **2655**

Deformation Potential in Germanium from Optical Absorption Lines for Exciton Formation.—W. H. Kleiner & L. M. Roth. (*Phys. Rev. Lett.*, 15th April 1959, Vol. 2, No. 8, pp. 334–336.) A second absorption peak observed by Zwerdling et al. (*Phys. Rev.*: to be published) is explained by the presence of strain.

537.311.33 : 546.289 **2656**

Two Electrical Phenomena on Liquid Germanium Surfaces.—A. I. Bennett. (*Phys. Rev.*, 1st Feb. 1959, Vol. 113, No. 3, pp. 773–774.) The separate phenomena described are (*a*) repulsion of surface scum by electric discharge from the molten surface to a solid Ge electrode; (*b*) induction of nucleation by an electric field applied to the surface of supercooled Ge.

537.311.33 : 546.289 **2657**

Conductivity of Grain Boundaries in Grown Germanium Bicrystals.—B.

Reed, O. A. Weinreich & H. F. Mataré. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 454–456.) Grain-boundary conduction has been studied as a function of doping. Resistivities of about 3 000–11 000 Ω/square were found for various samples and the behaviour shows only a small temperature dependence from 2°K to 300°K. This suggests that grain-boundary behaviour is not due to the segregation of impurities at the boundary.

537.311.33 : 546.289 **2658**

Effect of Minority Impurities on Impurity Conduction in *p*-Type Germanium.—H. Fritzsche & K. Lark-Horovitz. (*Phys. Rev.*, 15th Feb. 1959, Vol. 113, No. 4, pp. 999–1001.) Two different processes for impurity conduction at low and high impurity concentrations could be distinguished by observing the change of resistivity resulting from a change in the degree of compensation in the temperature range of impurity conduction (as for *n*-type Ge).

537.311.33 : 546.289 **2659**

Optical Constants of Germanium in the Region 1 to 10 eV.—H. R. Philipp & E. A. Taft. (*Phys. Rev.*, 15th Feb. 1959, Vol. 113, No. 4, pp. 1002–1005.)

537.311.33 : 546.289 : 535.215 : 538.63 **2660**

The Influence of Fast Holes on the Photoelectromagnetic Effect in Germanium.—A. K. Walton & T. S. Moss. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 692–694.) Results of measurements of the short-circuit photoelectromagnetic current are consistent with the existence of 2% of fast holes which have six times the normal mobility.

537.311.33 : 546.289 : 538.569.4 **2661**

Electron Spin Resonance in Nickel-Doped Germanium.—G. W. Ludwig & H. H. Woodbury. (*Phys. Rev.*, 15th Feb. 1959, Vol. 113, No. 4, pp. 1014–1018.) Electron spin resonance absorption, proportional in intensity to the Ni^- concentration, has been detected at 14 kMc/s.

537.311.33 : 546.289 : 538.63 **2662**

Analysis of Phonon-Drag Thermomagnetic Effects in *n*-Type Germanium: Part 2.—C. Herring, T. H. Geballe & J. E. Kunzler. (*Bell Syst. tech. J.*, May 1959, Vol. 38, No. 3, pp. 657–747.) Observations reported in Part 1 (3885 of 1958) have been accounted for quantitatively by theory, and yield information on phonon-phonon scattering, transport and deformation-potential theories and surface effects.

537.311.33 : 546.289 : 538.632 **2663**

Measurement of Dependence of the Hall Effect in *n*-Type Ge on Pressure up to 10 000 kg/cm².—A. I. Likhter & T. S. D'yakonova. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 95–103.)

537.311.33 : 546.289 : 539.16 **2664**

Energy Levels in Irradiated Germanium.—E. I. Blount. (*Phys. Rev.*, 15th Feb. 1959, Vol. 113, No. 4, pp. 995–998.)

537.311.33 : 546.289 : 621.314.7 **2665**

A Self-Limiting Electrolytic Etching Method for the Manufacture of Thin Base Layers of *n*-Type Germanium.—E. Fröschle. (*Telefunken-Röhre*, Sept. 1958, No. 35, pp. 63–76.) In the method described a negatively biased rectifying contact is placed on the rear of the wafer to be etched, and etching will cease when the barrier layer has been reached. Mirror-like surfaces of diameter 50–250μ and thickness 2·5–25 μ have been obtained.

537.311.33 : 546.682.19 **2666**

Some Effects of Copper as an Impurity in Indium Arsenide.—C. Hilsum. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 685–686.) Room-temperature annealing, and doping with Cu, are found to give similar effects on heat-treated InAs.

537.311.33 : 546.73.86 **2667**

Alloying of the Semiconductor Compound $CoSb_3$.—L. D. Dudkin & N. Kh. Abrikosov. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 142–151.) The effect of 13 different elements on the thermoelectric and thermal properties of $CoSb_3$ is investigated. Impurity atoms of Ni and Te act as donors, Sn as an acceptor; Fe lowers the thermal conductivity of the lattice.

537.311.33 : 546.786-31 **2668**

Electrical Conduction in Crystals and Ceramics of WO_3.—S. Sawada & G. C. Danielson. (*Phys. Rev.*, 1st Feb. 1959, Vol. 113, No. 3, pp. 803–805.) Experimental measurements of resistivity up to 1 000°K are described and discussed. Hall measurements near room temperature gave reasonable results for carrier density and mobility.

537.311.33: [546.812.231 + 546.289.231 **2669**

Anomalous Behaviour in the Hall Coefficients of the Semiconducting Compounds SnSe and GeSe.—S. Asanabo & A. Okazaki. (*Proc. phys. Soc.*, 1st May 1959, Vol. 73, No. 473, pp. 824–827.) During thermal cycles, anomalous variations were observed in the resistivity and Hall coefficient but not in their ratio. Possible explanations are suggested.

537.311.33 : 546.873.241 **2670**

Some Adiabatic and Isothermal Effects in Bismuth Telluride.—W. Williams. (*Proc. phys. Soc.*, 1st May 1959, Vol. 73, No. 473, pp. 739–744.) Isothermal and quasi-adiabatic Hall and Nernst coefficients were measured on *n*- and *p*-type Bi_2Te_3 at temperatures from 100°K to 450°K. With the magnetic field in two orthogonal directions, the difference between the two Hall coefficients showed the expected temperature variation, but the Nernst coefficients did not.

537.311.33 : 621.317.3 **2671**

Determination of the Surface Conductivity of Semiconducting Crystals by the 'Wedge' Method.—R. N. Rubinshteĭn & V. I. Fistul'. (*Dokl. Ak. Nauk S.S.S.R.*, 21st March 1959, Vol. 125, No. 3, pp. 542–545.) The specific resistance of the crystal is measured by two probes which are moved along the surface of a Ge sample cut in the shape of a wedge of angle $\leqslant 7°$. By measuring the gradient of the potential along the specimen, the surface and volume conductivity are calculated by means of formulae derived. Experimental results are given for inhomogeneous and homogeneous Ge samples before and after etching with H_2O_2.

537.312 : 537.531 **2672**

Conductivity induced by Radiation in Polycrystalline Cadmium Sulphide and Polyethylene.—C. G. Clayton, B. C. Haywood & J. F. Fowler. (*Nature, Lond.*, 18th April 1959, Vol. 183, No. 4668, pp. 1112–1113.)

538.22 **2673**

Low-Temperature Electrical and Magnetic Behaviour of Dilute Alloys: Mn in Cu and Co in Cu.—I. S. Jacobs & R. W. Schmitt. (*Phys. Rev.*, 15th Jan. 1959, Vol. 113, No. 2, pp. 459–463.)

538.22 : 538.569.4 **2674**

Magnetic Absorption in the Spin System of Some Paramagnetic Salts at about 1 325 Mc/s.—H. Hadders, P. R. Locher & C. J. Gorter. (*Physica*, Oct. 1958, Vol. 24, No. 10, pp. 839–847.) A description primarily of the apparatus and resonant cavity method of measurement of absorption and dispersion, and of relating absorption to susceptibility. Results are given of measurements made at 20·4°K of absorption as a function of a parallel and of a perpendicular static magnetic field. It is suggested that in parallel fields the absorption connected with the nondiagonal elements of the magnetic moment plays a dominating role. See also 1834 of 1957 (Smits et al.).

538.22 : 538.632 **2675**

Change of Sign of the Hall Constant when Atoms in an Alloy assume an Ordered Arrangement.—A. P. Komar, N. V. Volkenshteĭn & G. V. Fedorov. (*Dokl. Ak. Nauk S.S.S.R.*, 21st March 1959, Vol. 125, No. 3, pp. 530–531.) Brief investigation of the variation of the Hall coefficient in Ni_3Mn alloy in the temperature range 296°–4·2°K.

538.221 **2676**

Measurement of the Boundary Layer Width between Domains in Ferromagnetics.—L. V. Kirenskiĭ & V. V. Veter. (*Dokl. Ak. Nauk S.S.S.R.*, 21st March 1959, Vol. 125, No. 3, pp. 526–529.) A description of a method for the measurement of boundary layers of ferromagnetics containing 3% Si. The results show that the width of the layer in single crystals is not fixed. Values of 0·89 μ and 0·64 μ were obtained on two different samples.

538.221 **2677**

On the Influence of the Demagnetizing Field on Domain Structure.—J. Kociński. (*Acta phys. polon.*, 1958, Vol. 17, No. 5, pp. 283–294.) A discrepancy between experimental results and Néel's theory of domain structure for a crystal rod of Fe can be partly explained by the influence of the demagnetizing field acting near the end of the rod.

538.221 **2678**
Temperature Dependence of Spontaneous Magnetization in Small Ferromagnetic Particles.—E. Kneller. (*Z. Phys.*, 16th Oct. 1958, Vol. 152, No. 5, pp. 574–585.) An investigation of the spontaneous magnetization of a super-paramagnetic alloy with ferromagnetic particles of average diameter 23 Å, in the temperature range $0\cdot38$–$0\cdot93\,\theta$, where θ is the Curie point. Results are discussed with reference to previous measurements.

538.221 **2679**
On the Ferromagnetic Phase in Manganese - Aluminium System. — H. Kōno. (*J. phys. Soc. Japan*, Dec. 1958, Vol. 13, No. 12, pp. 1444–1451.) A metallographical investigation of the Mn-Al system in the composition range 47 %–60 % Mn is described. The conditions necessary for formation, structure and stability, and the thermal and magnetic properties of a metastable tetragonal ferromagnetic phase are discussed.

538.221 : 538.632 **2680**
On the Hall Effect at the Curie Point.—N. S. Akulov & A. V. Cheremushkina. (*Zh. eksp. teor. Fiz.*, Aug. 1958, Vol. 35, No. 2(8), pp. 518–519.) Brief description of an investigation carried out on Fe-Al alloys between −200 and +500 °C. Results are shown graphically.

538.221 : 538.632 **2681**
Anomalously High Hall Effect in the Chromium - Tellurium Ferromagnetic Alloy.—I. K. Kikoin, E. M. Buryak & Yu. A. Muromkin. (*Dokl. Ak. Nauk S.S.S.R.*, 11th April 1959, Vol. 125, No. 5, pp. 1011–1014.) Examination of galvanomagnetic effects in ferromagnetic alloys with non-ferromagnetic constituents. A 50 % Cr-Te alloy showed a very large Hall coefficient.

538.221 : 539.23 **2682**
Thin Magnetic Films.—R. S. Webley. (*Nature, Lond.*, 4th April 1959, Vol. 183, No. 4666, pp. 999–1000.) A description of apparatus for monitoring the low-frequency hysteresis loop both during deposition of the film and afterwards. Experiments indicate that the variation in coercivity during formation is influenced very strongly by secondary effects such as the nature of the substratum.

538.221 : 539.23 **2683**
Dependence of Geometric Magnetic Anisotropy in Thin Iron Films.—T. G. Knorr & R. W. Hoffman. (*Phys. Rev.*, 15th Feb. 1959, Vol. 113, No. 4, pp. 1039–1046.) Results indicate that the magnetic anisotropy results from a fibre axis structure which develops during deposition.

538.221 : 621.317.411.029.63 **2684**
Internal Ferromagnetic Resonance in Nickel.—J. C. Anderson & B. Donovan. (*Proc. phys. Soc.*, 1st April 1959, Vol. 73, No. 472, pp. 593–599.) The internal resonance revealed by measurements of the complex permeability of polycrystalline and colloidal Ni has been investigated over the temperature range 5°–100 °C.

538.221 : 621.318.124 **2685**
An Anomaly in the Characteristics of the Magnetic Properties of Ba Ferrite as a Function of Final Sintering Temperature.—G. Heimke. (*Naturwissenschaften*, June 1958, Vol. 45, No. 11, pp. 260–261.) Anomalies in the density, magnetic saturation and remanence were observed in specimens containing $CaO.SiO_2$ sintered at 1150° and 1180 °C.

538.221 : 621.318.134 **2686**
Reversing Ferrite Temperature Coefficients.—A. B. Przedpelski. (*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 74–76.) The results are given of an investigation of the influence of a d.c. magnetic field on the temperature coefficient of permeability of ferrite and yttrium garnet materials. As flux density is increased, the temperature coefficient in each case is found to change polarity and increase negatively until saturation of the material is approached.

538.221 : 621.318.134 **2687**
Crystal-Oriented Ferroxplana.—A. J. Stuijts & H. P. J. Wijn. (*Philips tech. Rev.*, 10th Feb. 1958, Vol. 19, No. 7/8, pp. 209–217.) Methods for aligning crystals are described and the subsequent magnetic properties are discussed. The permeability of some specimens is increased by a factor of 2·5–3 and the limiting frequency is about 0·8 times that of unaligned specimens. See also 1202 *q* of 1958 (Jonker et al.) and 1802 of 1958 (Braun).

538.221 : 621.318.134 **2688**
On the Solubility of MgO in Magnesium Ferrite.—L. C. F. Blackman. (*Trans. Faraday Soc.*, March 1959, Vol. 55, Part 3, pp. 391–398.)

538.221 : 621.318.134 **2689**
Resonance Measurements on Nickel-Cobalt Ferrites as a Function of Temperature and on Nickel Ferrite-Aluminates.—J. E. Pippin & C. L. Hogan. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 77–82. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, p. 804.)

538.221 : 621.318.134 **2690**
Ferrimagnetic Resonance in Some Polycrystalline Rare-Earth Garnets.—G. P. Rodrigue, J. E. Pippin, W. P. Wolf & C. L. Hogan. (*Trans. Inst. Radio Engrs*, Jan. 1958, Vol. MTT-6, No. 1, pp. 83–91. Abstract, *Proc. Inst. Radio Engrs*, April 1958, Vol. 46, No. 4, pp. 804–805.)

538.222 : 538.569.4 **2691**
Millimetre-Wave Paramagnetic Resonance Spectrum of ^{6}S-State Impurity (Fe^{+++}) in $MgWO_4$.—M. Peter. (*Phys. Rev.*, 1st Feb. 1959, Vol. 113, No. 3, pp. 801–803.) Experimental determination of the spectrum and its description by a spin Hamiltonian containing only second-order terms.

538.632 **2692**
Hall-Effect Devices.—W. J. Grubbs. (*Bell Syst. tech. J.*, May 1959, Vol. 38, No. 3, pp. 853–876.) A general survey of operation of 18 devices.

539.23 : 546.86 **2693**
Electrical and Magnetic Properties of Thin Films of Antimony.—A. Colombani, C. Vautier & P. Huet. (*C. R. Acad. Sci., Paris*, 24th Nov. 1958, Vol. 247, No. 21, pp. 1838–1841.) Results are given of resistivity, Hall effect and magnetoresistance measurements on vapour-deposited films of thickness 100–2 500 Å.

621.315.61 : 678.84 **2694**
Selection Guide for Silicone Dielectrics.—C. G. Currin. (*Electronics*, 10th April 1959, Vol. 32, No. 15, pp. 64–65.) Dielectric and mechanical properties and the average life of silicone insulating materials are tabulated.

MATHEMATICS

517.63 : 517.522 **2695**
The Application of the Laplace Transformation for the Summation of Weakly Convergent Series.—O. Heymann. (*Arch. elekt. Übertragung*, July 1958, Vol. 12, No. 7, pp. 326–330.)

517.94 : 621.372.8 **2696**
Asymptotic Behaviour of the Characteristic Functions of the Equation $\Delta u + k^2u = 0$ with Boundary Conditions along Equidistant Curves and the Scattering of Electromagnetic Waves in a Waveguide.—V. P. Maslov. (*Dokl. Ak. Nauk S.S.S.R.*, 1st Dec. 1958, Vol. 123, No. 4, pp. 631–633.)

517.94 : 621.374.43 **2697**
Some Properties of Mathieu and Related Functions exemplified by the Regenerative Modulation Process.—Jungfer. (See 2509.)

MEASUREMENTS AND TEST GEAR

531.76 : 534.22-8 : 621.374.5 **2698**
The Measurement of the Time Delay of Ultrasonic Delay Lines.—J. Sears. (*J. Brit. Instn Radio Engrs*, April 1959, Vol. 19, No. 4, pp. 237–244.) "A method of determining the delay of wide-band ultrasonic delay lines to an accuracy of better than 0·1 microseconds is described. Alternative methods are discussed."

621.3.011.4 (083.74) **2699**
Determination of the Unit of Capacitance.—A. F. Dunn. (*Canad. J. Phys.*, Jan. 1959, Vol. 37, No. 1, pp. 35–46.) The absolute unit of capacitance has been determined to an accuracy within ± 0·0025 % using laboratory-maintained frequency and resistance units and a modified Wien-bridge measuring circuit.

621.3.018.41 (083.74) **2700**

Canadian Standard of Frequency.—S. N. Kalra, C. F. Pattenson & M. M. Thomson. (*Canad. J. Phys.*, Jan. 1959, Vol. 37, No. 1, pp. 10–18.) Three laboratories each with 100-kc/s Essen ring type crystal oscillators jointly compose the standard of frequency, intercomparison being made over telephone lines. Results indicate a frequency stability within about 2 parts in 10^{10} over short and long periods.

621.3.018.41 (083.74) **2701**

Canadian Caesium-Beam Standard of Frequency.—S. N. Kalra, R. Bailey & H. Daams. (*Nature, Lond.*, 28th Feb. 1959, Vol. 183, No. 4661, pp. 575–576.) The equipment is similar in principle to that described by Essen & Parry (204 of 1958) but differs in the methods of synthesis and measurement of frequency and in the detailed design of the atomic-beam apparatus.

621.317.34 : 621.372.2 **2702**

The Evaluation of Quadripole and Material Measurements with the Logarithmic Transmission-Line Chart.—K. Jost & G. Schiefer. (*Arch. elekt. Übertragung*, July 1958, Vol. 12, No. 7, pp. 295–300.) The method described facilitates the numerical evaluation of the permeability and dielectric constants of materials from high-frequency measurements using a slotted coaxial line. Logarithmic charts are given with curves of constant voltage s.w.r. and constant node displacement.

621.317.34.029.64 : 621.317.755 **2703**

A Microwave Reflectometer Display System.—G. M. Clark & R. D. Rookes: J. C. Dix & M. Sherry. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 300–301.) Comment on 1308 of April and authors' reply.

621.317.4 : 621.385.833 **2704**

Method of Measurement of the Induction and its Derivatives on the Axis of Magnetic Electron Lenses.—P. Durandeau, B. Fagot & M. Laudet. (*C. R. Acad. Sci., Paris*, 22nd Dec. 1958, Vol. 247, No. 25, pp. 2316–2318.) Application of a method proposed by Laudet (1560 of 1957). Accuracy depends on the precision of measurement of the flux, whatever the diameter of the search coil.

621.317.7 : 621.314.7 **2705**

Accurate Measurement of Transistor Cut-Off Frequency.—Y. Tarui. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 284–287.) Equipment is described for measurement of current gain, α, of transistors in the frequency range 1–20 Mc/s. The absolute value and phase of α can be measured to within 1·5 %. A rapid determination of α cut-off frequency is possible.

621.317.7 : 621.373.43 **2706**

Calibrated Source of Millimicrosecond Pulses.—E. J. Martin, Jr. (*Electronics*, 17th April 1959, Vol. 32, No. 16, pp. 56–57.) A simple generator using a coaxial discharge line.

621.317.715 : 621.375.13 **2707**

Galvanometer Feedback Systems.—J. A. Sirs. (*J. sci. Instrum.*, May 1959, Vol. 36, No. 5, pp. 223–227.) "The principles of applying feedback to a galvanometer, after optical and electronic amplification, are discussed. In particular, the galvanometer performance is examined when proportional, differential, compound and selective feedback systems are used. The latter method is compared with mechanical and series-capacitor tuning of the galvanometer response."

621.317.729.1 : 621.3.032.269.1 **2708**

Investigation of Planar Electron Guns with a Rheographic Tank by the Method of Current Injection.—J. Bonnerot. (*C. R. Acad. Sci., Paris*, 24th Nov. 1958, Vol. 247, No. 21, pp. 1824–1826.) A note on the method of simulating space-charge effects by means of current-injecting probes in an electrolyte tank.

621.317.737 **2709**

An Instrument for the Measurement of the Q-Factor of Inductances by the Method of Complex Compensation.—W. Wisch. (*NachrTech.*, May 1958, Vol. 8, No. 5, pp. 214–220.) The equipment described covers the frequency range 100 c/s–40 kc/s for inductances from about 1 mH to 10 H with a resistive component between 0·1 Ω and 5 kΩ.

621.317.74 : 621.373.43 **2710**

Pulse and Square-Wave Generators.—(*Electronic Radio Engr*, June 1959, Vol. 36, No. 6, pp. 211–219.) A review of circuits, special features, and performance limits of commercial instruments.

621.317.742.029.64 **2711**

An Automatic Standing-Wave Indicator for the 3-cm Waveband.—E. Laverick & J. Welsh. (*J. Brit. Instn Radio Engrs*, April 1959, Vol. 19, No. 4, pp. 253–262.) The instrument is based on the rotary type of indicator, the mechanical rotation of the detector being replaced by a ferrite polarization-rotating section and a fixed detector. Voltage s.w.r. or reflection coefficient is indicated directly on a meter. The errors of the instrument and its applications are discussed.

621.317.75 **2712**

Frequency Analyser uses Two Reference Signals.—T. B. Fryer. (*Electronics*, 1st May 1959, Vol. 32, No. 18, pp. 56–57.) Circuit details are given of an instrument which can be used from sub-audio up to radio frequencies. A wide range of filter bandwidths is available.

621.317.755 **2713**

Low-Voltage Oscilloscope Tubes.—F. de Boer & W. F. Nienhuis. (*Philips tech. Rev.*, 30th Nov. 1957, Vol. 19, No. 5, pp. 159–164.) An account of the development of two tubes requiring an anode voltage of 400 V.

621.317.794 **2714**

Radiation Measurements at Radio Frequencies: a Survey of Current Techniques.—W. A. Cumming. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 705–735.) A general survey of measuring techniques in diffraction, scattering, transmission and reflection, current distribution, aperture fields, radiation patterns and gain. 129 references.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

551.508.8 : 629.19 : 621.398 **2715**

Tracking Earth's Weather with Cloud-Cover Satellites.—R. Hanel, R. A. Stampfl, J. Cressey, J. Licht & E. Rich, Jr. (*Electronics*, 1st May 1959, Vol. 32, No. 18, pp. 44–49.) A description of the instrumentation of a satellite which scans the cloud cover and transmits pictures to the ground. See *Convention Rec. Inst. Radio Engrs*, 1958, Vol. 6, Part 5, pp. 136–141 (Hanel & Stampfl).

621.365.5 : 621.387 **2716**

A Method of Producing Eddy-Current Heating.—R. J. Armstrong. (*J. sci. Instrum.*, May 1959, Vol. 36, No. 5, pp. 246–247.) A circuit is described which uses two gas-filled triodes to generate trains of damped oscillations. In the particular case considered about 1 kW is generated at about 1 Mc/s and there are fifteen oscillations in each train.

621.38/.39 : 656.2 **2717**

Electronics in the Railway Industry.—B. K. Cooper. (*Elect. Rev., Lond.*, 20th Feb. 1959, Vol. 164, No. 8, pp. 335–339.) Signalling and traffic control applications of radar, television and automatic computers are discussed.

621.384.613 **2718**

On the Electron Capture Mechanism and Limiting Current in Betatrons.—A. N. Matveev. (*Zh. eksp. teor. Fiz.*, Aug. 1958, Vol. 35, No. 2(8), pp. 372–380.) In this investigation account is taken of the Coulomb interaction of electrons in the beam and of the electron losses on the walls of the vacuum chamber. A formula is derived for the limiting current which is also applicable to relativistic electron energies.

621.384.622.2 **2719**

22-MeV Electron Linear Accelerator.—N. A. Austin & S. C. Fultz. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 284–289.) An accelerator of the travelling-wave type with a wide range of pulse lengths, repetition rates and beam energies.

621.384.8 : 537.4 **2720**

Strong-Focusing Ion Source for Mass Spectrometers.—C. F. Giese. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 260–261.) Starting with an ion beam 0·375 in. square, the lens produces a line focus 0·60 in. long and 0·025 in. wide with a half-angle of divergence beyond the cross-over of 0·034 radians.

621.385.833 **2721**

Ion Focusing Properties of a Quadrupole Lens Pair.—H. A. Enge. (*Rev. sci. Instrum.*, April 1959, Vol. 30, No. 4, pp. 248–251.) The focusing properties of quadrupole lens pair have been studied, and the results of thick-lens calculations are presented in the form of graphs showing the field-strength parameters and magnifications as functions of object and image distances.

621.398 : 621.396.934 : 621.317.7 **2722**

Canaveral Test Range: Timing-Signal Transmission.—R. P. Wells. (*Bell Lab. Rec.*, March 1959, Vol. 37, No. 3, pp. 96–99.) The methods used to send range-timing signals over a single channel of the Canaveral–Puerto Rico cable system are described.

PROPAGATION OF WAVES

621.396.11 **2723**

Wave Propagation over an Irregular Terrain: Part 3.—K. Furutsu. (*J. Radio Res. Labs, Japan*, Jan. 1959, Vol. 6, No. 23, pp. 71–102.) General formulae for field strength obtained earlier, are used for estimating ridge and precipice effects on ground waves. The results are displayed in diagrams for practical use. Part 2: 1845 of 1958.

621.396.11 **2724**

Radio Echoes Observed on Sea Swell at the Casablanca Ionospheric Sounding Station.—A. Haubert. (*Ann. Géophys.*, July/Sept. 1958, Vol. 14, No. 3, pp. 368–372.) At ranges up to 100 km, echoes are observed which have a regular period that is inversely proportional to the radio frequency. Quantitative checks show that these echoes probably originate from sea waves; the beat frequency is consistent with a Doppler shift in the frequency of the reflected wave. See also 1237 of 1958 (Okamoto et al.).

621.396.11 : 551.510.52 **2725**

Models of the Atmospheric Radio Refractive Index.—B. R. Bean & G. D. Thayer. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 740–755.) Improved prediction of refraction effects, particularly over long distances and at great heights in the atmosphere, is given from the value of refractive index at the transmitting point by the two models described.

621.396.11 : 551.510.52 **2726**

Airborne Radiometeorological Research.—W. S. Ament. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 756–761.) A summary of the work of the Naval Research Laboratory in radar and propagation investigations.

621.396.11 : 551.510.535 **2727**

Investigations of Great-Circle Propagation between Eastern Australia and Western Europe.—H. J. Albrecht. (*Geofis. pura appl.*, Sept.-Dec. 1957, Vol. 38, pp. 169–180. In English.) Observations made during sunspot minimum in the frequency range 3–30 Mc/s are analysed and discussed with reference to ionospheric predictions.

621.396.11 : 551.510.535 **2728**

On the Reflection of Electromagnetic Waves in the Ionosphere with Vertical Sounding.—P. Szulkin. (*Bull. Acad. polon. Sci., Sér. Sci. tech.*, 1959, Vol. 7, No. 1, pp. 65–68. In English.) When the gradient of refractive index exceeds a certain value, a wave will undergo partial reflection at a height below the height of total reflection. For a typical case, it is shown that partial reflections will only occur at heights less than 2 km from the height of total reflection. This is confirmed by experimental observations.

621.396.11 : 551.510.535 (98) **2729**

The Development of Radio Traffic Frequency Prediction Techniques for Use at High Latitudes.—O. A. Sandoz, E. E. Stevens & E. S. Warren. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 681–688.) Some prediction techniques are described with particular attention to the problem in Northern Canada. The use of back-scatter and oblique-incidence sounding is discussed. 50 references.

621.396.11 : 551.510.535 : 523.78 **2730**

Propagation Mechanism of High-Frequency Waves related to the Annular Eclipse of 19th April, 1958.—S. Kanaya & K. Ueno. (*Rep. Ionosphere Res. Japan*, June 1958, Vol. 12, No. 2, pp. 188–195.) The propagation mechanism for distances up to 3 000 km can be studied by ionospheric sounding at the path midpoint and the measurement of angle of elevation at the receiving point.

621.396.11 : 551.510.535 : 551.594.6 **2731**

'Whistler Mode' Echoes Remote from the Conjugate Point.—R. L. Dowden & G. T. Goldstone. (*Nature, Lond.*, 7th Feb. 1959, Vol. 183, No. 4658, pp. 385–386.) Echoes of pulsed signals from Tokyo on a frequency of 17·44 kc/s, with delays of about 0·2 s, have been detected at Hobart, Tasmania, 3 500 km from the geomagnetic conjugate point to Tokyo.

621.396.11 : 551.510.535 : 621.396.812 **2732**

Determination of an Absorption Index of the Ionosphere from an Automatic-Statistical Analysis of Field-Strength Recordings.—H. Schwentek. (*Arch. elekt. Übertragung*, July 1958, Vol. 12, No. 7, pp. 301–308.) Description of a method of determining nondeviative ionospheric absorption using field-strength measurements of c.w. transmissions at oblique incidence. The main transmission paths are separated by a statistical analyser.

621.396.11.029.62 : 551.510.535 **2733**

IGY Observations of F-Layer Scatter in the Far East.—R. Bateman, J. W. Finney, E. K. Smith, L. H. Tveten & J. M. Watts. (*J. geophys. Res.*, April 1959, Vol. 64, No. 4, pp. 403–405.) Peculiar signal enhancements observed during transmissions at 36 to 50 Mc/s between the Philippines and Okinawa appear to represent F-layer scatter. These signals are observed nightly for periods of several hours during the months of September and October. Pulse tests indicate F-layer heights for these signals. Considerable pulse broadening is observed and the signals generally arrive from somewhat off the great circle path.

621.396.11.029.64 **2734**

Microwave Propagation over the Sea Beyond the Line of Sight.—K. Nishikori, A. Takahira & H. Irie. (*J. Radio Res. Labs, Japan*, Jan. 1959, Vol. 6, No. 23, pp. 57–70.) Measurements were taken for 3 years of the meteorological conditions over a 200-km path and from them refractive indices were calculated. These measurements were used to produce duct-shaped distributions. Using ray tracing methods and Fresnel expressions for reflection and penetration of ducts, quantitative values of attenuation and fading of microwave propagation over the same path were calculated; these agreed with the actual measured values.

621.396.11.029.45 : 551.510.535 **2735**

Diurnal Change of Ionospheric Heights Deduced from Phase Velocity Measurements at V.L.F.—J. R. Wait. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, p. 998.) The phase stability of 16-kc/s waves propagated over long distances [see e.g. 2524 of 1958 (Crombie et al.)] is deduced from waveguide-mode theory.

621.396.11.029.45 : 621.396.93 **2736**

V.L.F. Propagation Measurements for the Radux-Omega Navigation System.—C. J. Casselman, D. P. Heritage & M. L. Tibbals. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 829–839.) Measurements of the phase stability of 10–18-kc/s transmissions over a 4 200-km path are reported.

621.396.81.029.62 **2737**

Some Characteristics of Persistent V.H.F. Radio-Wave Field Strengths Far Beyond the Radio Horizon.—L. A. Ames, E. J. Martin & T. F. Rogers. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 769–777.) Ground and air measurements have been made at 220 Mc/s up to distances 700 miles beyond the horizon and at altitudes up to 40 000 ft. Dependence on 'angular distance' and the effect of tropospheric layers are discussed and prediction curves are presented. Weak fields measured beyond 800 miles are believed to be caused by ionospheric propagation. Airborne measurements on ionospheric scattering at 50 Mc/s up to 1 700 miles are also reported.

621.396.11.029.63 **2738**

Phase Front Fluctuations of Ten-Centimetre Radio Waves Propagated over a Sea Surface.—A. V. Men', S. Ya. Braude & V. I. Gorbach. (*Dokl. Ak. Nauk S.S.S.R.*, 11th April 1959, Vol. 125, No. 5, pp. 1019–1022.) A description of an experimental investigation of the propagation of vertically polarized waves over a range of 33 km. The phase fluctuations were measured in the range 0·01–100 c/s during the periods July–September (summer–autumn) and October–December (autumn–winter). Results indicate that with a few rare exceptions the fluctuations obeyed a normal-distribution law.

621.396.81.029.62: 551.510.535: 523.78 **2739**

Signal Intensity of Ionospheric Forward Scatter at V.H.F. during the Annular Eclipse of 19th April 1958.—(*Rep. Ionosphere Res. Japan*, June 1958, Vol. 12, No. 2, pp. 196–197.) No significant variation either in continuous wave signal strength or in the pattern of received pulses could be associated with the solar eclipse.

621.396.812.029.62 **2740**

A Year's Field-Strength Measurements in the V.H.F. Band at Kolberg near Berlin.—U. Kühn. (*Geofis. pura appl.*, Sept.–Dec. 1957, Vol. 38, pp. 157–168. In German, with English summary.) The results of measurements made in the frequency range 88–100 Mc/s are correlated with temperature, air pressure and air mass, and with the length of the propagation path. On the average the lowest field strengths were observed in June and December.

RECEPTION

621.396.621 : 621.396.666 **2741**

Investigations on Diversity Reception by the Aerial Selection Method.—R. Heidester & K. Vogt. (*Nachrichtentech. Z.*, June 1958, Vol. 11, No. 6, pp. 315–319.) Receiver switching methods are compared with an aerial switching method. Equipment for aerial diversity reception using transistors is described and operational results obtained with it are discussed.

621.396.666 : 621.396.62 : 621.314.7 **2742**

A Diode Circuit for Automatic Volume Control in Transistor Receivers.—R. Cantz. (*Telefunken-Röhre*, Sept. 1958, No. 35, pp. 31–42.) Difficulties experienced in a.v.c. for transistor receivers can be overcome by using a crystal diode in parallel with the transistor input stage. To reduce the d.c. consumption a transistor can be used in place of the diode.

621.396.822 **2743**

Measurement of Atmospheric Noise.—B. B. Ghosh & S. N. Mitra. (*J. Inst. Telecommun. Engrs, India*, Dec. 1958, Vol. 5, No. 1, pp. 2–16.) Measurements were made in India at four frequencies in the h.f. range by both objective and subjective methods. Data are given for more than two years. Comparison between the two methods indicates that a protection of 40 dB against noise is required for satisfactory broadcast reception 90 % of the time.

STATIONS AND COMMUNICATION SYSTEMS

621.39 : 358.236 **2744**

Progress and Problems in U.S. Army Communications.—R. E. Lacy. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 650–660.) Radio equipment, transmission, multiplexing, switching and data processing are discussed.

621.391 **2745**

Probability of Error for Optimal Codes in a Gaussian Channel.—C. E. Shannon. (*Bell Syst. tech. J.*, May 1959, Vol. 38, No. 3, pp. 611–656.) "A study is made of coding and decoding systems for a continuous channel with an additive gaussian noise and subject to an average power limitation at the transmitter. Upper and lower bounds are found for the error probability in decoding with optimal codes and decoding systems. These bounds are close together for signalling rates near channel capacity and also for signalling rates near zero, but diverge between. Curves exhibiting these bounds are given."

621.391 **2746**

The Reliability of Binary Transmissions for Various Types of Modulation.—H. J. Held. (*Nachrichtentech. Z.*, June 1958, Vol. 11, No. 6, pp. 286–292.) The error probability is calculated for the more important binary keying systems, and an approximation formula of general applicability is given. A comparison is made with experimental results.

621.391 **2747**

Note on Binary Decoding.—M. J. E. Golay. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 996–997.) The logical basis of decoding techniques is discussed in relation to time taken and equipment required.

621.395.44 : 621.315.1.052.63 **2748**

High-Frequency Transmission on High-Voltage Lines.—A. de Quervain. (*Tech. Mitt. PTT*, 1st Sept. 1958, Vol. 36, No. 9, pp. 349–354.) The suitability of power lines for multichannel carrier telephony is discussed, with particular reference to the causes of interference and to the means of overcoming major difficulties inherent in such systems.

621.396/.397 (047.1) **2749**

Sound and Television Broadcasting.—E. Schwartz. (*VDI Z.*, 11th Feb. 1959, Vol. 101, No. 5, pp. 189–191.) A progress review for 1958 with 85 references mainly to German literature.

621.396.1.029.51 **2750**

The Engineering of Communication Systems for Low Radio Frequencies.—J. S. Belrose, W. L. Hatton, C. A. McKerrow & R. S. Thain. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 661–680.) The factors affecting design of systems in the frequency range 80–200 kc/s are reviewed, with particular emphasis on propagation, aerial design and modulation system. Practical examples relating to arctic regions are given.

621.396.4 : 552.510.52 **2751**

Tropospheric Scatter System using Angle Diversity.—J. H. Vogelman, J. L. Ryerson & M. H. Bickelhaupt. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 688–696.) A microwave system is proposed using multiple primary feeds to each parabolic reflector with one transmitter or receiver for each feed. The results of a simple experiment show that the wave-angle diversity obtained should reduce the 'medium to aperture' coupling loss. The method also allows higher total transmitter power to be used at microwave frequencies.

621.396.4 : 621.376.2 **2752**

A Critical Analysis of some Communication Systems Derived from Amplitude Modulation.—W. D. Nupp. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 697–704.) A comparison is made between s.s.b. suppressed-carrier operation and a system of d.s.b. suppressed-carrier operation, as applied to aeronautical mobile services.

621.396.5 **2753**

'Frena', a System of Speech Transmission at High Noise Levels.—F. de Jager & J. A. Greefkes. (*Philips tech. Rev.*, 9th Oct. 1957, Vol. 19, No. 3, pp. 73–83.) A nonlinear system in which the frequency and amplitude components are transmitted in two separate channels. If the amplitude component is transmitted as a f.m. signal, the necessary minimum signal/noise ratio is 6 dB. This can be reduced to 4 dB using the 'Frenac' system in which the amplitude component is coded.

621.396.5 : 534.76 **2754**

Tunable F.M. Multiplex Adapter for Stereo.—W. B. Bernard. (*Electronics*, 10th April 1959, Vol. 32, No. 15, pp. 66–67.) 50-kc/s bandwidth is obtained at 455 kc/s, and a heterodyne system eliminates the need for filters. Muting is also provided.

621.396.5 : 534.78 **2755**

The Effect of Restricted Frequency Characteristics on Intelligibility of Speech in Radio Telephone Channels.—A. R. Ravi Varma & H. D. Krishna Prasad. (*J. Inst. Telecommun. Engrs, India*, Dec. 1958, Vol. 5, No. 1, pp. 38–43.)

621.396.65 **2756**

Swedish Norwegian S.H.F. Link.—(*Overseas Engr*, March 1959, Vol. 32, No. 375, p. 255.) The multichannel radio link being installed between Oslo and Karlstad (Sweden) operates in the frequency band 3·8–4·2 kMc/s and provides 600 telephone channels or one high-definition television channel. Two two-way paths and one reversible one-way path are to be constructed.

SUBSIDIARY APPARATUS

621.311.69 **2757**

The American Miniature Nuclear Generator, Snap III.—W. S. Eastwood, L. B. Mullett & J. L. Putman. (*Nature*, *Lond.*, 7th March 1959, Vol. 183, No. 4662, pp. 643–644.) A note on a power supply system for telemetry apparatus in a space rocket. The decay energy of a radioactive Po capsule is converted directly to electrical energy by means of thermocouples, giving an output of 5 W with efficiency 8–10 %.

621.316.721.078.3 : 621.318.381 : 621.314.7 **2758**

A Transistor Stabilized Supply to Feed an Electromagnet used for Nuclear Resonance Studies.—M. Sauzade. (*C. R. Acad. Sci., Paris*, 12th Jan. 1959, Vol. 248, No. 2, pp. 205–207.) A circuit is described for a 70-V 15-A supply, with current stabilized to 3 parts in 10^5.

621.316.722.1 : 621.314.63 **2759**

Zener Diodes Stabilize Tube Heater Voltages.—P. L. Toback. (*Electronic Ind.*, Dec. 1958, Vol. 17, No. 12, pp. 64–66.) Typical applications of Zener power diodes to the regulation of a.c. and d.c. heater supplies are described.

621.352.7.001.4 **2760**

Internal Resistance of Dry Cells.—(*Tech. News Bull. nat. Bur. Stand.*, Feb. 1959, Vol. 43, No. 2, pp. 23–25.) A description of a simple nondestructive pulse technique for investigating the effect of a momentary or continuous discharge on internal resistance. For a more detailed account see *J. electrochem. Soc.*, June 1959, Vol. 106, No. 6, pp. 471–475.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 **2761**

Television Study Centre.—M. Boella. (*Ricerca sci.*, Jan. 1959, Vol. 29, No. 1, pp. 25–39.) Report on the research work carried out between 1953 and 1957 at the National Research Council centre in Turin.

621.397.61 : 535.623 **2762**

An N.T.S.C. Colour Modulator for the C.C.I.R. Standard.—F. Jaeschke. (*Arch. elekt. Übertragung*, June 1958, Vol. 12, No. 6, pp. 271–288.) Design problems of the equipment are considered, particularly that of the conversion of the three monochrome signals supplied by the camera tube into the luminance and chrominance components. Measurement and control accessories are also discussed.

621.397.61.004.5 **2763**

Television Monitoring Installations.—O. Macek. (*Frequenz*, July 1958, Vol. 12, No. 7, pp. 214–223.) A survey of modern equipment and monitoring methods, mainly as used in Germany. 60 references.

621.397.62 **2764**

A Luminous Frame around the Television Screen.—J. J. Balder. (*Philips tech. Rev.*, 30th Nov. 1957, Vol. 19, No. 5, pp. 156–158.) The results of subjective tests are given.

621.397.62 **2765**

Contrast Filters for Television Sets.—R. Suhrmann. (*Elektronische Rundschau*, July 1958, Vol. 12, No. 7, pp. 227–232.) Measurements of filter transparency combined with subjective tests on television receivers indicate little difference in effectiveness between grey filters and selective filters placed in front of tube screens.

621.397.62 : 621.385.3 **2766**

The Pin Triode in the Frequency Range of the Television Bands IV and V.—Maurer. (See 2796.)

VALVES AND THERMIONICS

621.314.63 **2767**

The Time-Lag of Germanium Diodes and its Effect in Simple Rectifier and Limiter Circuits.—W. Heinlein. (*Frequenz*, May & June 1958, Vol. 12, Nos. 5 & 6, pp. 159–163 & 191–198.) The delay effects are investigated for diodes operating with sinusoidal voltages, and related to the results obtained for pulse operation (see 2907 of 1958).

621.314.7 **2768**

The H.F. Transistor and its Complex Characteristics in the Frequency Range 0–2 Mc/s.—G. Ledig. (*Frequenz*, May & June 1958, Vol. 12, Nos. 5 & 6, pp. 137–148 & 178–190.) Detailed theoretical treatment of *n*-pole networks by matrix methods. The transistor is considered as a three-pole network and its parameters and characteristics are calculated with particular regard to its practical application in receivers at 470 kc/s. Procedure and equipment for measuring the complex characteristics are described and equivalent circuits of the transistor are derived.

621.314.7 **2769**

Power Transistors.—D. S. Grant, R. O. Jones & T. Scott. (*Research, Lond.*, Jan. 1959, Vol. 12, No. 1, pp. 21–25.) Techniques for increasing emitter efficiency and heat dissipation are summarized.

621.314.7 **2770**

The Limits of Validity and the Model-Based Derivation of the Equivalent Circuit of Junction Transistors.—M. A. Nicolet. (*Helv. phys. Acta*, 10th March 1959, Vol. 32, No. 1, pp. 58–77. In German.) Parameters of the hybrid-II equivalent circuit of junction transistors were accurately measured for several types of transistor. The results can be interpreted by reference to the model of Webster (2798 of 1954) and Rittner (3390 of 1954) if account is taken of three additional effects. A modified equivalent circuit is proposed which has an extended range of application.

621.314.7 **2771**

Experimental Investigations on the Influence and Creation of Lattice Defects in Junction Transistors.—E. Baldinger, H. Bilger & M. A. Nicolet. (*Helv. phys. Acta*, 10th March 1959, Vol. 32, No. 1, pp. 78–88. In German.) Measurements have been made to determine the influence of lattice defects in Si and Ge transistors. Results are interpreted with reference to the theory of Sah et al. (3899 of 1957). The influence of fast and thermal neutron radiation on differential parameters and noise of Ge transistors is also investigated. Low-frequency noise appears to be mainly due to surface effects.

621.314.7 **2772**

Avalanche Transistors.—R. C. V. Macario. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 262–267.) Review of characteristics, equivalent circuits and applications.

621.314.7 **2773**

The Charge Storage in a Junction Transistor during Turn-Off in the Active Region.—R. S. C. Cobbold. (*Electronic Engng*, May 1959, Vol. 31, No. 375, pp. 275–277.) "Through the solution of the one-dimensional diffusion equation for a junction transistor, equations are derived for the emitter and collector currents that exist under conditions of minimum turn-off time. These equations are shown to be in fairly good agreement with the practical results."

621.314.7 **2774**

Calculation of Transients in Transistors.—A. A. Grinberg. (*Fiz. Tverdogo Tela*, Jan. 1959, Vol. 1, No. 1, pp. 31–43.) Calculation of the transient conductivity of a transistor with consideration of the limiting value of the internal resistance, the collector capacitance and the load resistance.

621.314.7 : 517.7 **2775**

Jacobians—a New Computational Tool.—T. R. Nisbet & W. W. Happ. (*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 69–71.) Tables are given for converting transistor parameters for each of the three circuit configurations.

621.314.7 : 621.317.7 **2776**

Accurate Measurement of Transistor Cut-Off Frequency.—Tarui. (See 2705.)

621.314.7 : 621.318.57 **2777**

A Storing and Switching Transistor.—W. v. Münch & H. Salow. (*Nachrichtentech. Z.*, June 1958, Vol. 11, No. 6, pp. 293–299.) Insertion of a tungsten point into the collector of a *p-n-p*-type transistor during the alloying process produces a switching transistor with thyratron-like input characteristics. The transistor has a differential input impedance of 1 MΩ in the blocked state, and < 10 Ω in the conducting state. See also 3244 of 1956 (Salow & v. Münch).

621.314.7 : 621.318.57 **2778**

Transistors for Electronic Switching.—T. R. Robillard & R. W. Westberg. (*Bell Lab. Rec.*, March 1959, Vol. 37, No. 3, pp. 88–92.) A description of design problems in the development of transistors for the new electronic telephone switching system.

621.314.7 : 621.385.4 **2779**

Field-Effect Tetrode.—(*Electronic Radio Engr*, June 1959, Vol. 36, No. 6, p. 210.) A new four-terminal semiconductor developed by Bell Telephone Laboratories which can function as a transformer, gyrator, isolator, non-distorting modulator, or short-circuit-stable negative resistance.

621.314.7.004.15 : 629.19 **2780**

An Estimate of Transistor Life in Satellite Applications.—A. J. Heeger, T. R. Nisbet & W. W. Happ. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, p. 991.) Transistors used in satellite instruments will be affected by the radiation bands in the outer atmosphere. Their expected life is of the order of 10^4 hours.

621.314.7.012.8 **2781**

The Junction Transistor as a Network Element at Low Frequencies: Part 2—Equivalent Circuits and Dependence of the *h* Parameters on Operating Point.—J. P. Beijersbergen, M. Beun & J. te Winkel. (*Philips tech. Rev.*, 9th Oct. 1957, Vol. 19, No. 3, pp. 98–105.) Part 1: 1719 of May.

621.314.7.029.3 **2782**

Low Frequencies Vary T-Parameters.—G. N. Kambouris. (*Electronic Ind.*, Dec. 1958, Vol. 17, No. 12, pp. 69–71.) An examination of certain a.f. transistors indicates that, due to feedback, their characteristic impedances are not constant above 50 c/s. Measurements of equivalent T-parameters should therefore be made at frequencies at which the feedback impedance can be neglected.

621.383.2 : 535.371.07 **2783**

Image Converters and Image Intensifiers for Military and Scientific Use.—M. W. Klein. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 904–909.)

621.385.029.6 **2784**

Microwave Power Tubes—a Survey.—(*Electronic Ind.*, Nov. 1958, Vol. 17, No. 11, pp. 103–131.) The survey includes characteristics of different types of valve, listed under the manufacturer's name, and a glossary of microwave terms and abbreviations.

621.385.029.6 **2785**

The Multigap Klystron, a Generator of Short Electromagnetic Waves.—R. Hechtel. (*Telefunken-Röhre*, Sept. 1958, No. 35, pp. 5–30.) Theoretical investigation of the characteristics of a new type of klystron in which the drift tube has a number of cylindrical gaps. A comparison with the reflex klystron shows the advantages of this valve with regard to bandwidth and efficiency.

621.385.029.6 **2786**

Surface Waves in Electron Beams in the Presence of a Magnetic Field.—G. Mourier. (*C. R. Acad. Sci., Paris*, 1st Dec. 1958, Vol. 247, No. 22, pp. 1978–1980.) A study of electron velocities in a beam of finite thickness, using a dynamic model.

621.385.029.6 **2787**

An Experimental C.W. Power Travelling-Wave Tube.—M. O. Bryant, J. F. Gittins & F. Wray. (*J. Electronics Control*, Feb. 1959, Vol. 6, No. 2, pp. 113–129.) Design, performance and constructional methods are detailed. Power output is about 1 kW at X-band frequencies with gain up to 22 dB. Bandwidth at fixed beam voltage is about 1·5%. With varying beam voltage a 6% range of frequencies is covered. The slow-wave structure is of the clover-leaf type. Ferrite isolators in the feed waveguides ensure stability.

621.385.029.6 : 061.3 **2788**

International Convention on Microwave Valves.—(*Proc. Instn elect. Engrs*, Part B, 1958, Vol. 105, Supplement No. 10, pp. 401–608.) The text is given of the following papers which were among those read at the I.E.E. Convention held in London 19th–23rd May 1958.

Travelling-Wave Tubes—I:

(*a*) Design of a 100-mW Helix Travelling-Wave Amplifier at 50 Gc/s.—W. E. Danielson, H. L. McDowell & E. D. Reed (pp. 405–408).

(*b*) Experimental Medium-Power High-Gain Travelling-Wave Tubes of Very Short Structure, with Permanent-Magnet Focusing, for the 4- and 6-Gc/s Bands.—W. Klein & H. Neufischer (pp. 408–411).

(*c*) An Application of the Spatial-Harmonic Wave on a Large-Diameter Helix.—J. Koyama (pp. 412–414).

Discussion (pp. 415–418).

Magnetrons:

(*d*) A New Design of High-Power S-Band Magnetron.—H. A. H. Boot, H. Foster & S. A. Self (pp. 419–425).

(*e*) A 200-kW 80-Watt Q-Band Magnetron.—R. Zwobada (pp. 426–428).

(*f*) An Investigation into the Factors Affecting the Life of Magnetrons.—F. C. Thompson (pp. 429–430).

(*g*) The Theory of Circular Magnetrons with Uniformly Rotating Space Charge.—R. Dunsmuir (pp. 431–439).

(*h*) A 3·8 mm-Wavelength Pulsed Magnetron.—A. J. Monk (pp. 440–442).

Discussion (pp. 443–445).

Travelling-Wave Tubes—II:

(*i*) Theory of Travelling-Wave Tubes using Paralleled Electron Streams.—C. K. Birdsall (p. 445).

(*j*) Power Limitations in Helix Travelling-Wave Tubes and the Application of Fluid Cooling.—G. M. Clarke (pp. 446–454).

(*k*) An M-Type Pulsed Amplifier.—O. Doehler, A. Dubois & D. Maillart (pp. 454–457).

(*l*) A 20-kW Pulsed Travelling-Wave Tube.—J. D. Pearson & H. S. Cockroft (pp. 458–463).

(*m*) Constructional and Design Difficulties of Wide-Band High-Power Pulsed Travelling-Wave Tubes in the 10-Gc/s Region.—C. H. Dix & R. G. Robertshaw (pp. 464–467).

(*n*) Some Transient Phenomena in Microwave Tubes.—A. V. Brown (pp. 468–474).

(*o*) A 4-Gc/s Travelling-Wave Tube for Microwave Radio Links.—P. F. C. Burke (pp. 475–479).

(*p*) A Package-Type Travelling-Wave Amplifier with a New Magnetic Focusing System.—M. Kenmoku & S. Yasuda (pp. 480–484).

Gas-Discharge Valves and Plasma: See 2800 below.

Backward-Wave Oscillators—I (Devices):

(*q*) Frequency Pushing in Crossed-Field Oscillators.—R. I. Buick, A. Reddish & I. J. Zucker (pp. 525–529).

(*r*) Operation Characteristics of the Carmatron Tube.—O. Doehler, B. Epsztein & J. Arnaud (pp. 529–533).

(*s*) Results Obtained on Cross-Field Carcinotrons under Pulsed Operation.—M. Favre (pp. 533–537).

(*t*) Anomalous Behaviour in the M-Type Carcinotron.—J. Nalot & R. Visocekas (pp. 538–542).

Discussion (pp. 542–543).

Grid Control Valves:

(*u*) Electrode Spacing in Disc-Seal Triodes.—M. R. Gavin, W. Fulop & L. J. Herbst (pp. 544–549).

(*v*) Advances in the Techniques and Applications of Very-High-Power Grid-Controlled Tubes.—M. V. Hoover (pp. 550–558).

(*w*) Magnetic Coupling by Parallel-Wire Grids and Soldered Cross-Lateral Grids in Disc-Seal Triodes.—J. Kellerer (pp. 559–562).

(*x*) Measurement of the Active Admittances of a Triode at 4 Gc/s.—M. T. Vlaardingerbroek (pp. 563–566).

(*y*) A New Method of Making Accurate Fine-Wire Grids for Use in Radio Valves.—A. E. Widdowson, D. C. Gore & C. H. Butcher (pp. 567–576).

(*z*) Microwave Triodes.—H. Groendijk (pp. 577–582).

(*aa*) A New Method of Measuring the Input Admittance of Ultra-High-Frequency Triodes.—K. Takashima & T. Misugi (pp. 583–585).

Discussion (pp. 585–587).

Backward-Wave Oscillators—II (Theory):

(*bb*) The Magnitude of the Locking-Signal for Backward-Wave Oscillators.—E. A. Ash (pp. 588–593).

(*cc*) Interaction of Electromagnetic Waves and Electron Beams in Systems with Centrifugal-Electrostatic Focusing.—Z. S. Chernov (pp. 594–596).

(*dd*) A New Crossed-Field Travelling-Wave Tube, the M-J.—C. C. Johnson & C. K. Birdsall (p. 597).

(*ee*) An Experimental Study of Large-Signal Behaviour in M-Type Valves in the Presence of Space Charge by the Use of an Analogue Method.—B. Epsztein (pp. 598–604).

(*ff*) Aspects of M-Type Interaction with Particular Reference to the Backward-Wave Magnetron Amplifier.—J. W. Klüver (pp. 605–608).

621.385.029.6 : 537.533 : 621.396.822 **2789**

Investigations of the Anomalous Noise in Magnetically Focused Electron Beams.—H. P. Louis. (*Tech. Mitt. PTT*, 1st Sept. 1958, Vol. 36, No. 9, pp. 333–

348.) The excess noise produced by scalloped beam amplification in Brillouin-focused beams is investigated experimentally.

621.385.029.65 **2790**

BWO uses Ridge-Loaded Ladder Circuit.—J. A. Noland & L. D. Cohen. (*Electronics*, 1st May 1959, Vol. 32, No. 18, pp. 66..69.) The backward-wave oscillator described operates in the frequency range 60–75 kMc/s and uses the ladder circuit described by Karp (1212 of 1955 and 2647 of 1957). It was found to be reliable and suitable for use with permanent-magnet focusing.

621.385.032.213.13 **2791**

Dispenser Cathodes.—(*Philips tech. Rev.*, 23rd Dec. 1957, Vol. 19, No. 6, pp. 177–190.)

Part 1: Introduction.—A. Venema (pp. 177–179).

Part 2: The Pressed Cathode.—R. C. Hughes & P. P. Coppola (pp. 179–185).

Part 3: The Impregnated Cathode.—R. Levi (pp. 186–190).

Two types of cathode are described using the same principle as, and having similar characteristics to the L-cathode [see e.g. 773 of 1951 (Lemmens et al.)] but having the advantage that they can be more readily manufactured.

621.385.032.213.13 **2792**

Evaporation of Barium from Cathodes Impregnated with Barium-Calcium-Aluminate.—I. Brodie, R. O. Jenkins & W. G. Trodden. (*J. Electronics Control*, Feb. 1959, Vol. 6, No. 2, pp. 149–161.) The rate of barium formation during the life of impregnated cathodes has been shown to decay in a manner dependent on the thickness and porosity of the tungsten.

621.385.1.002 **2793**

Modern Trends of Construction in the Development of Valves.—H. Katz. (*Nachrichtentech. Z.*, June 1958, Vol. 11, No. 6, pp. 281–285.) Review of structural design and electrode and envelope materials suitable for operation at high temperatures and high frequencies, and capable of meeting exacting performance specifications.

621.385.2.029.6 : 621.376.23 **2794**

Vacuum-Diode Microwave Detection.—N. E. Dye, J. Hessler, Jr, A. J. Knight, R. A. Miesch & G. Papp. (*Electronics*, 24th April 1959, Vol. 32, No. 17, p. 110.) Tests with specially constructed coaxial diodes indicate that noise figures do not compare favourably with crystal performance.

621.385.3 : 621.365.5 **2795**

The Use of Oscillator Triodes in High-Frequency Generators with Variable Load.—E. G. Dorgelo. (*Elektronische Rundschau*, July 1958, Vol. 12, No. 7, pp. 241–247.) The performance of triodes specially designed for use in r.f. heating equipment is discussed.

621.385.3 : 621.397.62 **2796**

The Pin Triode in the Frequency Range of the Television Bands IV and V.—R. Maurer. (*Telefunken-Röhre*, Sept. 1958, No. 35, pp. 43–62.) The limitations imposed by lead inductances on the use of pin triodes for frequencies in the range 470–800 Mc/s are discussed. Equations are derived and evaluated for the triode Type PC86, which has been designed for operation in this range.

621.385.3.029.63 **2797**

A Transmitting Triode for Frequencies up to 900 Mc/s.—P. J. Papenhuijzen. (*Philips tech. Rev.*, 16th Oct. 1957, Vol. 19, No. 4, pp. 118–128.) Detailed description of the construction and characteristics of an air-cooled disk-seal triode Type TBL 2/300 which delivers as an oscillator a power of 405 W at 470 Mc/s and 155 W at 900 Mc/s.

621.385.3.029.64 **2798**

A 4 000-Mc/s Wide-Band Amplifier using a Disc-Seal Triode.—J. P. M. Gieles. (*Philips tech. Rev.*, 30th Nov. 1957, Vol. 19, No. 5, pp. 145–156.) Description of an amplifier based on a valve Type EC56 or EC57 with which an output power of 0·5 or 1·5 W respectively can be obtained, with a gain of 8 dB.

621.385.832.032.9 **2799**

Automatic Electrical Welding of Cathode-Ray Tubes.—(*Engineering, Lond.*, 24th Oct. 1958, Vol. 186, No. 4833, pp. 554–555.) Description of a new mass-production glass sealing plant incorporating many automatic processes which include electrical welding to make pressure-tight joints between the main glass sections of the bulb.

621.387 + 537.56] : 061.3 **2800**

International Convention on Microwave Valves.—(*Proc. Instn elect. Engrs*, Part B, 1958, Vol. 105, Supplement No. 10, pp. 485–524.) The text is given of the following papers which were among those read at the I.E.E. Convention held in London, 19th–23rd May 1958. For titles of other papers included in Supplement No. 10 see 2788 above.

Gas-Discharge Valves and Plasma:

(*a*) A New Form of X-Band Pre-T.R. Cell.—A. B. Parker (pp. 488–491).

(*b*) A Microwave Pulsed Attenuator using an R.F.-Excited Discharge.—P. D. Lomer & R. M. O'Brien (pp. 500–504).

(*c*) Active Microwave Duplexing Systems.—P. O. Hawkins (pp. 505–507).

(*d*) Passive Protection Cells.—P. D. Lomer (pp. 508–509).

(*e*) A Wide-Band Multi-way Electronic Switch for Use at Microwave Frequencies.—S. M. Hamberger (pp. 510–515).

(*f*) Electromechanical Modes in Plasma Waveguides.—R. W. Gould & A. W. Trivelpiece (pp. 516–519).

(*g*) One-Dimensional Nonstationary Flow in a Plasma.—G. Kalman (p. 520).

Discussion (pp. 521–524).

MISCELLANEOUS

061.3 : 621.3 **2801**

Highlights of '59 I.R.E. Show.—J. M. Carroll, W. E. Bushor & S. Weber. (*Electronics*, 1st May 1959, Vol. 32, No. 18, pp. 39–43.) Description of some of the developments reported in papers read at the I.R.E. National Convention, New York 1959, including guidance of land vehicles by radar, microwave computers, plans for tracking space ships, inertial navigation, and microwave measurements on semiconductors. For the Convention program and summaries of papers, see *Proc. Inst. Radio Engrs*, March 1959, Vol. 47, No. 3, pp. 456–489.

061.4 : 621.396.6 **2802**

Radio Components Show.—(*Wireless World*, May 1959, Vol. 65, No. 5, pp. 211–218.) Detailed review of components and accessories at the R.E.C.M.F. Exhibition held in London, 6th–9th April 1959.

621.3 : 623.8 **2803**

Some Representative Electronic Subsystem Development Problems in Naval Ordnance.—E. H. Beach, M. J. Parker, P. Yaffee & R. B. Knowles. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 929–945.) Design problems are discussed for equipment required for underwater and missile-carried ordnance, for field-system evaluation and test sets.

621.38.004.15 **2804**

Numerical Approach to Electronic Reliability.—J. J. Naresky. (*Proc. Inst. Radio Engrs*, May 1959, Vol. 47, No. 5, Part 1, pp. 946–956.) A description of progress made at Rome Air Development Centre, N.Y., in obtaining maximum reliability in electronic equipment.

621.38.004.15 : 061.3 **2805**

American Electronics Reliability Symposium.—R. Brewer. (*Brit. Commun. Electronics*, March 1959, Vol. 6, No. 3, p. 197.) A brief report of the Fifth National Symposium on Reliability and Quality Control in Electronics held in Philadelphia, Pa, January 1959.

621.38.004.6 **2806**

Predicting the Reliability of Complex Electronic Equipment.—A. G. Field. (*Trans. Soc. Instrum. Technol.*, March 1959, Vol. 11, No. 1, pp. 18–23.) The reliability is calculated on the basis of a constant component failure rate. The effects of ambient conditions on failure rate are discussed.

629.19 : 621.396 **2807**

The Challenge of Space.—H. A. Manoogian. (*Electronics*, 24th April 1959, Vol. 32, No. 17, pp. 65–80.) Includes a short description of possible systems of communication, navigation and guidance for interplanetary flight.

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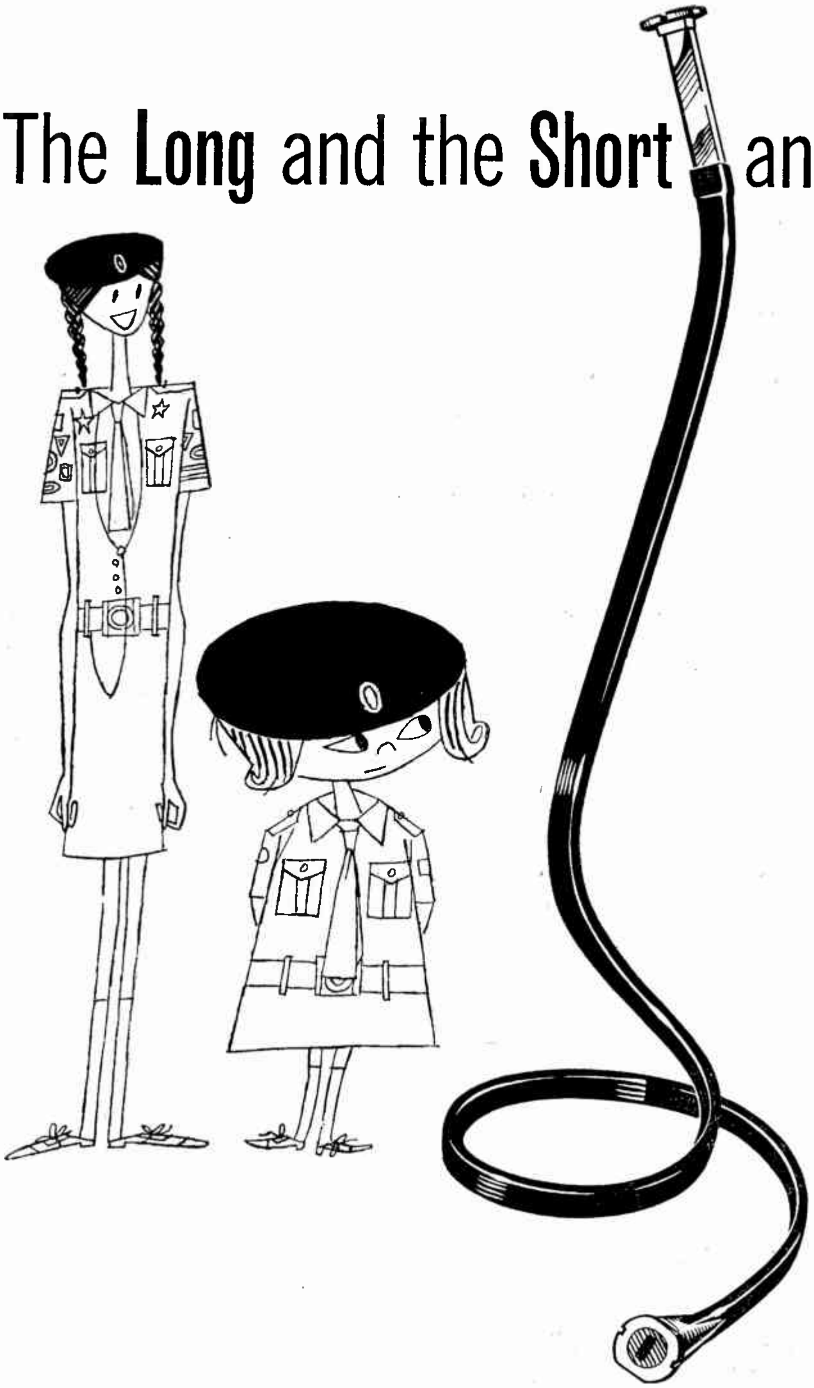

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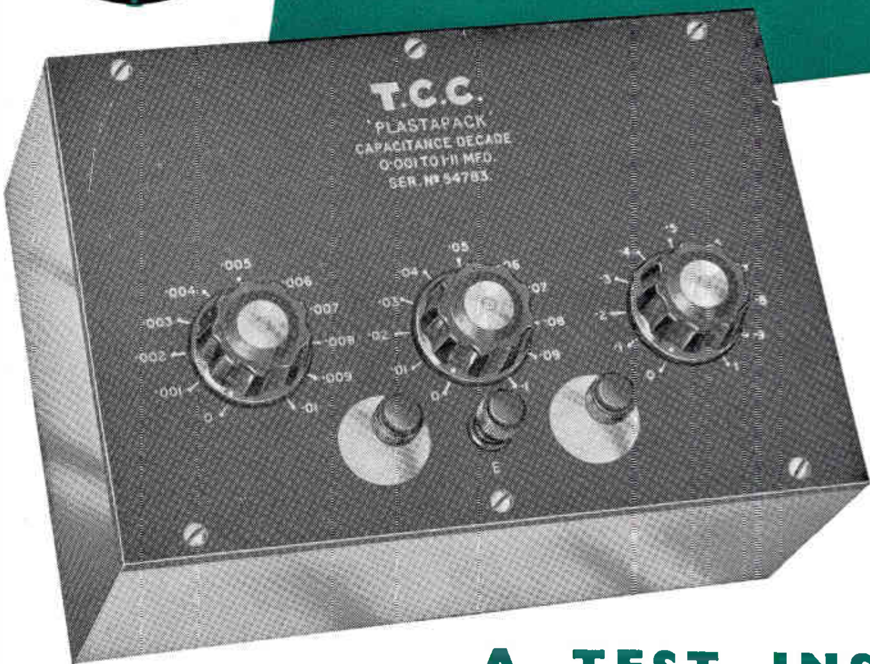

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